Metamaterials Science and Technology
Series Editors: Balamati Choudhury
Vinod K. Tewary

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Shiv Narayan
Arun Kesavan
Editors

Handbook of Metamaterial-Derived Frequency Selective Surfaces

Springer

Metamaterials Science and Technology

Series Editors

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This Series of Handbooks focuses on all aspects of the emerging field of Metamaterial Science and Technology.

Metamaterials are artificially constructed materials exhibiting unique EM properties, which are generally not found in nature. In case of electromagnetic wave propagation through metamaterials, the electric field vector E, the magnetic field vector H, and the wave vector k constitute a left-handed triplet. This is not the case for naturally occurring materials. Many basic electromagnetic phenomena, such as Doppler shift, Snell's law, and Cherenkov radiation get reversed in metamaterials. These unusual but potentially useful properties open up unique possibilities to implement this novel structure in numerous applications such as, lenses and antennas design, imaging with super-resolution, microwave devices, dispersion compensating interconnects, radars, and most importantly in the seemingly sci-fi area of invisibility cloaking.

The handbook series covers all the aspects of metamaterial science and technology from theoretical modeling to fabrication techniques, and from measurements procedures to hardware realization. Each volume in this series is edited by experts, with more global experts contributing chapters based on their work. The volumes in this series will prove useful to researchers and practicing engineers alike.

Shiv Narayan • Arun Kesavan
Editors

Handbook of Metamaterial-Derived Frequency Selective Surfaces

With 506 Figures and 33 Tables

Springer

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To the supreme God who has given us energy, enthusiasm, and patience to complete this book.

Series Preface

Metamaterials are artificially constructed materials exhibiting unique electromagnetic properties, which are generally not found in nature. In case of electromagnetic wave propagation through a metamaterial, the electric field vector E, the magnetic field vector H, and the wave vector k constitute a left-handed triplet that is not the case for naturally occurring materials. Many basic electromagnetic phenomena, such as *Doppler shift*, *Snell's law*, *Cherenkov radiation,* get reversed in metamaterials. These unusual but potentially useful properties open up unique possibilities to implement this novel structure in numerous applications such as lenses and antennas' design, imaging with super-resolution, microwave devices, dispersion compensating interconnects, radar, and most importantly the dream come true research of invisibility cloaking.

The emergence of artificial materials during the 1950s and 1960s has opened up the area of development of new structures with unusual electromagnetic properties. Recent experimental demonstration of artificial materials with negative constitutive parameters has resulted in a new science and technology of materials. It has created an entirely new class of materials, called metamaterials. These new materials have the most fascinating physical properties. Since then, the study of metamaterials has evolved into a subject of intense interest.

Over 45 years ago, a Russian theorist Prof. Veselago proposed such a kind of hypothetical material. After the realization of metamaterials in 2001 by Prof. David R. Smith, this field has received considerable attention from many research groups worldwide. As this technology is going to be the high-performance contributor over the entire wave spectrum, ranging from acoustic to microwave and to optical, a handbook series covering all the applications toward hardware realization is of intense interest to the scientific community. The aim of this handbook series is just that. It will cover almost all aspects of metamaterial science and technology, starting from theoretical modeling, fabrication techniques, and measurement procedures to hardware realization.

The tentative plan for this handbook series is to bring out 10 volumes, comprising modeling and simulation, nano-metamaterials, metamaterial-based frequency selective surfaces, metamaterial antennas, photonic metamaterials, invisibility cloaking, etc. The editors of these handbook series volumes will be chosen from the pioneers/leaders of the Metamaterial Science and Technology Platform. Each volume will

provide complete guidelines of the respective metamaterial technology and up-to-date literature references.

We are confident that this handbook series will be very useful to researchers, students, and engineers, engaged in various applications of metamaterials over a wide frequency spectrum, covering microwave, terahertz, infrared, and optical frequencies. We are indebted to Late Dr. Rakesh Mohan Jha, the founder and Chief Scientist of Centre for Electromagnetics, CSIR-NAL, for envisaging this handbook series for researchers of the next generation.

Volume Preface

It was in my mind for a long time to contribute a book on *metamaterial-frequency selective surfaces* (MTM-FSS). For this, I was waiting to be mature in this area and now I have more than 10 years of experience in the design and development of frequency selective surfaces (using conventional as well as metamaterial elements) for aerospace applications. Meanwhile, my colleague Dr. Balamati Choudhury (Series editor of Metamaterial Science and Technology) offered me to edit a book on metamaterial-frequency selective surfaces and I accepted the offer. To accomplish such an endeavor, I requested my very dynamic and talented student, Dr. Arun Kesavan from INRS, University of Quebec, Montreal, Canada, to be a coeditor, and he accepted my invitation.

Frequency selective surfaces (FSS) consist of a two-dimensional array of apertures on metallic screen or metallic patches that exhibit filter characteristics for the electromagnetic wave impinging on them. During the past two decades, FSS find various applications starting from defense to medical to industry. FSS structures based on metamaterial element further opened its scope in millimeter, THz, and optical frequency regions. In order to cover current and near-past advancements in the field of MTM-FSS technology, we have prepared the content of the book. Moreover, the content was prepared in such a way that the book should be useful to researchers/students/academicians/scientists/engineers working in the area of MTM-FSS technology. In view of this, the content of the book is divided into four sections namely: (1) Defense and Security Systems, (2) Microwave Circuits and Devices, (3) Wireless Communication Systems, and (4) Metamaterial-FSS for Industrial Applications. To accomplish this project, we have invited researchers/academicians/scientists worldwide, working in different areas of electromagnetics. We are very thankful to the authors who have contributed chapters to this book.

Section I of the book contains 1–7 chapters; Chap. ▶ 1 presents the EM design and analysis of various metamaterial-FSS structures for airborne radome applications including MTM-FSS structures loaded with active devices. Chap. ▶ 2 discusses various metamaterial absorber designs in the GHz, THz, infrared, and optical regime and explains various absorption mechanism types. Moreover, it presents a detailed study on mathematical modeling of the metamaterial absorber which can be helpful to understand wave prorogation and its absorption. Chapter ▶ 3 describes the basic properties of graphene, the working of FSSs and RASs, and the applications of

graphene-based MTMs in the defense and security systems among stealth, communication, and imaging. Chapter ▶ 4 provides a review of various active metamaterial frequency selective surface (FSS)-based switchable and tunable radar absorbing structures (RASs). It describes the basic mechanism of active absorber geometries and their working principles, design topology, biasing mechanism, and experimental procedure of FSS-based switchable and tunable RAS designs. Chapter ▶ 5 presents the design and analysis of SIW-integrated metamaterial-FSS for radome applications. Further, the technique to reduce and improve the bandwidth of FSS structure using SIW is discussed. In Chap. ▶ 6, metamaterial-based high-performance engineered surface is discussed and illustrated some designs to validate the efficiency of the approach for RCS reduction applications and further describes the RCS reduction over a wide frequency band using polarization converter and reflection type metamaterial. Chapter ▶ 7 investigates the potential of designing multiangle retrodirective metasurfaces, studies their reliability in terms of retro-reflection (i.e., RCS) enhancement at a wide range of incident angles, and demonstrates the potential of retrodirective metasurfaces for applications dedicated to protect vulnerable road users such as cyclist and pedestrians.

Section II includes the discussion on *Microwave Circuits and Devices* through Chaps. ▶ 8–▶ 11. Chapter ▶ 8 discusses the polarizing functionalities of FSS structures in the background of metamaterial (MTM) literature. The design method of MTM-FSS polarizing structures, origin and mechanism of polarization, and computational and experimental characterization methods are explained. The significant developments made in MTM-FSS polarizer and its various application aspects are presented. In Chap. ▶ 9, EM analysis of metamaterial-based microwave coupler is presented based on an equivalent lumped Lange coupler (LC) circuit model. Further, the enhancement of fabrication limits of metamaterial coupler has also been elaborated. Chapter ▶ 10 aims at the recent developments of theoretical, numerical, and experimental aspects of metamaterial lens structures, for enhancing the imaging resolution of far-field microwave sensors. It summarizes the breadth of research done in the field of metamaterial lens tailored toward microwave nondestructive evaluation (NDE) of composites. Chapter ▶ 11 presents the simulation studies of the design of different types of bandpass filters including dual-band filters with metamaterial-based structures and complementary split-ring resonators.

Section III of the book describes the *Wireless Communication Systems* that contains Chaps. ▶ 12 through ▶ 19. Basically, this section includes a detailed discussion on the design of antennas and antennas array using metamaterial-FSS for modern wireless communication systems. Chapter ▶ 12 discusses the design methodologies of several broadband and compact antenna configurations, including millimeter wave antennas. In Chap. ▶ 13, the design of phase-gradient metasurface is described in detail for the beam steering applications and gives insight into the theoretical foundation and design methodologies. Moreover, it describes a recent demonstration of passive and active beam steering antenna array using phase gradient metasurface. Chapter ▶ 14 investigates the MTM-FSS and its anticipated role in successful future THz wireless communication and discusses motivation toward the design and development of MTM-FSS and their different applications

in wireless communication. Chapter ▶ 15 introduces the latest advancements in beam-tilting antenna system designs based on FSS for wireless communications and describes the various techniques of beam-tilting antennas based on passive and active FSS. In Chap. ▶ 16, the importance of circularly polarized (CP) millimeter-wave (MMW) antennas is presented based on metamaterial-FSS for wireless communication systems, and emphasis has been put to achieve high gain, linear-to-circular polarizer, wider CP bandwidth, spatially coupling reduction in MIMO antennas, and broadband radar cross-section (RCS) reduction in MMW antennas. Chapter ▶ 17 presents an overview of microwave-imaging technique and the role of metamaterial-FSS antennas in image sensing and discusses the operating principle, design considerations, and advantages of different categories of image-sensing antennas. Furthermore, a brief overview of metamaterial-based sub-diffraction imaging is presented. Chapter ▶ 18 describes the method for radiation enhancement of MIMO antenna systems using metamaterials and explains the design, characterization, and experimental validation of the method. Further, the chapter also describes the techniques to enhance the isolation between antenna elements in MIMO systems. Chapter ▶ 19 summarizes the different techniques used to enhance the radiation performances of antennas using metamaterial-FSS structures.

Section IV of the book describes the modern industrial applications of metamaterial-FSS structures that consist of chapters from Chaps. ▶ 20 to ▶ 24. Chapter ▶ 20 discusses the metamaterial-FSS based energy-harvesting system using the concept of absorption for the railway environment. Different MTM-FSS designs for single and multiple frequencies have been discussed. Chapter ▶ 21 describes the overview of use of metamaterials in medicine both in diagnosis and in therapeutics, and further, its usage in MRI for shielding effectiveness is discussed. In Chap. ▶ 22, an overview of metamaterial-FSS for various THz applications has been presented, including the design flow and simulation setup using EM simulation software. Chapter ▶ 23 explains the RFID technology its evolution and classifications. It describes the design and characterization of magneto-dielectric substrate (MDS) metamaterials for the miniaturization of HF RFID antennas. Finally, Chap. ▶ 24 describes how better performing MTM-FSS can be prototyped using the *additive manufacturing* method emphasizing different applications. It also includes a discussion on the types and advantages, and challenges of this technology along with their future advancements.

I hope the effort made by us and the contributing authors to bring out this book in the present form would be definitely useful to readers working in the area of electromagnetics.

Bangalore, India — Shiv Narayan
Ottawa, Canada — Arun Kesavan
January 2023

Acknowledgments

First, we would express our thanks to the authors who have contributed book chapters to complete this endeavor: Dr. Raveendranath U Nair and Ms. Sreeja J for contributing Chap. ▶ 1; Mr. Nilotpal R. and Dr. Bhattacharyya S for Chap. ▶ 2; Singh G and Prof. Sardana N for Chap. ▶ 3; Dr. Saptarshi Ghosh for Chap. ▶ 4; Dr. Krushnakanth V and Prof. Raghavan S for Chaps. ▶ 5 and ▶ 21; Dr. Samadi F for Chap. ▶ 6; Mr. Kalaagi M and Prof. Seetharamdoo for Chaps. ▶ 7 and ▶ 20; Dr. Yogesh N for Chap. ▶ 8; Dr. Arun Kumar G and Dr. Gopi Ram for contributing Chap. ▶ 9; Dr. Saptarshi Mukherjee et al. for Chap. ▶ 10; Dr. Vakula D and Sowjanaya A for Chap. ▶ 11; Dr. Reddy G for Chap. ▶ 12; Dr. Arnab Pattanayak for Chap. ▶ 13; Pandit N et al. for Chap. ▶ 14; Dr. Borhani-Kakhki M and Prof T Denidni for Chap. ▶ 15; Amit Kumar et al. for Chap. ▶ 16; Dr. Athira Raveendran and Dr. Raman S for Chap. ▶ 17; Ms. Kaur H and Dr. Singh H for Chap. ▶ 18; Dr. Sarin V and Prof. Vasudevan K for Chap. ▶ 19; Mr. RK Jaiswal for Chap. ▶ 22; Dr. Jamal Zaid for Chap. ▶ 23; and Dr. Karthikeyan S and Ms. Mellita R for contributing Chap. ▶ 24.

We would like to express our sincere gratitude to Dr. Ghanshyam Mishra, Mr. Saket Kaushal, Dr. Vipin Velayudhan, Dr. Vinod VKT, Dr. Sai Krishna Reddy, and Prof. Satish Sharma for their voluntary contribution as reviewers for the chapters in this book.

Finally, we would express our thanks to Series Editors, Dr. Balamati Choudhury and Dr. Vinod K. Tewary, and Springer publications for their collaboration.

Shiv Narayan
Arun Kesavan

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About the Series Editors

Dr. Balamati Choudhury is a scientist at the Centre for Electromagnetics of CSIR-National Aerospace Laboratories, Bangalore, India. She obtained her M.Tech. (ECE) degree from the National Institute of Science and Technology, India, and Ph.D. (Engg.) degree in Microwave Engineering from Biju Patnaik University of Technology (BPUT), India, in 2013. During 2006–2008, she was a senior lecturer in the Department of Electronics and Communication at NIST, Orissa, India.

Her active areas of research and teaching interests are in the domains of Soft Computing Techniques in Electromagnetic Design and Optimization, Computational Electromagnetics for Aerospace Applications, Metamaterial Design Applications, RF, and Microwaves. More specifically, the topics of the sponsored projects she has contributed to are development of Ray Tracing Techniques toward RF Analysis of Propagation in an Indoor Environment, Low RCS Design, Phased Arrays and Adaptive Arrays, and Conformal Antennas. She was the recipient of **ICCCES/ Outstanding Young Investigator Award** for the year 2016–2017 for her contributions in Metamaterial Science. She was also the recipient of the **CSIR-NAL Young Scientist Award** for the year 2013–2014 for her contribution in the area of Computational Electromagnetics for Aerospace Applications. Her work has been appreciated in various conferences (LAMSYS 2016, ISAMPE-2017, IRSI-2006, ANSH-2013) through best paper award.

Dr. Balamati has authored or co-authored over 190 scientific research papers and technical reports, besides two books, five SpringerBriefs, and three book chapters.

The book entitled *Soft Computing in Electromagnetics: Methods and Applications* was published by the Cambridge University Press, Cambridge, UK (2015). Another book *Metamaterial Inspired Electromagnetic Applications: Role of Intelligent Systems* was published by Springer, Singapore. She was the author of five SpringerBriefs published by Springer, Germany, which cover all the cutting-edge technologies such as Terahertz technology, invisibility cloaking, and HF ray–based methods for space applications. She has also book chapters in a monograph entitled *Recent Advances in Metamaterial Science and Technology* (Tech Science Press, USA, 2015), which is in press.

Dr. Vinod Tewary is a theoretical physicist at the Applied Chemicals and Materials Division of the National Institute of Standards and Technology (NIST) in Colorado, USA. He received his Ph.D. in Solid State Physics from Delhi University and has since then worked as a research associate in the Atomic Energy Research Establishment, in Harwell, UK; a professor of Physics and dean of Research at Birla Institute of Technology and Science, Pilani, India; and as an adjunct professor at Ohio State University, Columbus, before joining NIST.

His areas of research interest include multiscale modeling of nanostructures, modeling of mechanical properties of semiconductors and solid-state devices, crystal defects and phonons in nanostructures, with special focus on lattice, continuum, and multiscale Green's functions. He has been a recipient of several awards for his contributions to the field, including the Distinguished Fellow and Lifetime Achievement Awards, and the Eric Reissner Medal by ICCES (International Conference on Computational Engineering and Sciences) in 2017, 2013, and 2004, respectively. He received a Bronze Medal from the US Government (Department of Commerce) for Scientific Research and Technical Leadership in 1996, the Multicultural Award by Boulder (Colorado State) in 1994, and the Pride of India Award for Excellence in Sciences in 1992. He was elected fellow of the Institute of Physics, UK, in 1974.

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His research interests are broadly in the field of electromagnetic applications, and the topics include frequency selective surfaces (FSS), RAS, radome, metamaterials, numerical methods in electromagnetics, antennas, pattern synthesis of antenna array, and EM material characterizations. Particularly, he is working on the design and analysis of FSS structures based on advanced numerical techniques (i.e., MM-GSM, TLTMM, and FDTD method) for aerospace applications for the last eight years. Dr Shiv is the author/coauthor of over 100 technical documents, including peer-reviewed journal and conference papers. He has published Technical Briefs (Books) with Springer on various topic such as *FSS Technology for the Design of High Performance Antennas* and *FDTD Modeling Inside Microwave Cavities*. He was awarded with "Excellence Award in Research" for the year 2018–2019 from CSIR-NAL for his significant contributions to the *electromagnetic design and development of FSS structures for airborne*

applications such as radomes, RAS, and antennas. He was also awarded with prestigious "CSIR-Raman Research Fellowship" for the year 2019–2020 and visited College of Engineering, San Diego State University, San Diego, CA, USA.

He is the reviewer of several prestigious international journals published by IEEE Transactions on AP, IEEE AWPL, IEEE MTT, IEEE Sensors, Elsevier, Cambridge University Press, Springer Nature, Tech Science Press, etc.

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Part I

Defense and Security Systems

Metamaterial-FSS for the Design of High-Performance Radome

1

Shiv Narayan, J. Sreeja, and Raveendranath U. Nair

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Abstract

This chapter deals with the design and analysis of various hybrid-element metamaterial FSS and active metamaterial FSS for airborne radome applications. The EM performance analysis of several metamaterial FSS unit cells has been carried out based on TLTMM method and validated using full-wave analysis method. A unique left-handed metamaterial has been introduced based on the theory of tightly coupled resonators for narrow X-band application consisting of two inverted hybrid element conductors on each side of the substrate. The advantages of the proposed novel structure have been compared to standard double square loop FSS in terms of bandwidth and EM response. Further, varactor diodes are incorporated in the metamaterial unit with a broadband transmission characteristic (90% bandwidth: 1.4 GHz) in order to tune the EM characteristics. By varying

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the reverse bias voltage across varactor diode, the resonance frequency of proposed metamaterial unit cell has been tuned from 10 GHz to 11.43 GHz. Reconfigurable metamaterials find application in active radome design to enable the tuning of its operating frequency in view of electronic warfare.

Introduction

Artificial material characteristics which involve artificial dielectrics, artificial magnetics, and artificial plasma have been dated back by well-known scientists. In 1948, Kock suggested a three- dimensional array of metallic structures to simulate the crystalline lattices of dielectric materials. In this, metal elements (also called lattice inclusions) become oscillating dipoles under applied field similar to oscillating molecular dipoles of a natural dielectric. Another phenomenon suggested by Schelkunoff and Friis (1940) invokes magnetism without magnetic constituents with today's acquainted structure "Split-Ring Resonators (SRR)" depicting artificial magnetism for the resonant frequency range. Rotman and Brown (1960) reported a medium with negative permittivity (wire medium), and it is familiar currently by the name artificial plasma.

Electromagnetic metamaterials (MTMs) are theoretically described as artificial effectively homogeneous electromagnetic structures having unusual properties, which are not freely available in the nature (Veselago 1967). The size or periodicity of effectively homogeneous structure is used to be much smaller than the guided wavelength. For such structure, the refractive phenomena will dominate over other phenomenon (scattering/diffraction), when a EM wave propagates through the effectively homogeneous medium. The four possible combinations in the pair of permittivity and permeability (ε, μ) are (+, +), (+, −),(−, +), and(−, −). Among these, the MTM with both negative is the recent topic of research consideration, also known as *left-handed* (LH) medium and/or *Negative refractive Index* (NRI) material.

The negative-ε/positive-μ structure as proposed by Pendry in 2000 is a metal thin-wire structure. As electric field is applied parallel to the axis of the wires, an induced current along the wire length gives rise to equivalent electric dipole moments, thus exhibiting a plasmonic-type permittivity frequency function as follows,

$$\varepsilon_r(\omega) = 1 - \frac{\omega_{pe}^2}{\omega^2 + \tau^2} + j\frac{\tau\omega_{pe}^2}{\omega(\omega^2 + \tau^2)} \tag{1}$$

where $\omega_{pe} = \sqrt{2\pi c^2}/p^2\, ln\left(\frac{p}{a}\right)$; "$a$" is the radius of wires and "p" is the periodicity.

The positive-ε or negative-μ MTM is realized with the split-ring resonator (SRR) structure when the excitation of magnetic field, H, is perpendicular to the plane of the rings; thus, it induces resonating currents in the SRR loop to produce magnetic dipole moments and exhibits a plasmonic-type permeability function as follows,

$$\mu_r = \left(1 - F\omega^2\right)/\left(\omega^2 - \omega_{0m}^2 + j\omega\tau\right) \tag{2}$$

where $F = \pi(r/p)^2$, "r" is the inner radius of the smaller ring and.

$\omega_{0m} = c\sqrt{3p/\pi \ln(2\omega r^3/\delta)}$ where ω is the width of the rings and δ is the radial spacing between the rings.

Many interesting effects can be seen in the propagation of EM waves through the LH material, for example, backward wave propagation due to antiparallel group and phase velocity resulting in backward travel of wave fronts while the energy is moving in the forward direction. Also, size reduction up to a great extent can be achieved for relevant conditions pertaining to particular design. First experimentation validating the LH behavior done by Smith (Smith et al. 2000) was influential to the research community in this field (Shelby et al. 2001).

The approach used in our work is the subwavelength resonator-based as used by Pendry et al. (1999) for SRR designs two decades ago. Several different type of resonators have been proposed by people until now such as omega-shaped, S-shaped, labyrinth type, fishnet structures, tri-loaded dipole structure, Jerusalem cross, etc. The primary advantage offered by these compact structures is flexibility in EM design practice that makes them suitable for miniaturization. They can be coupled with coplanar, stripline, and microstrip line in many different ways (Radonić et al. 2007). This is known as resonator-loading approach. The main consideration is the geometrical parameters in any metamaterial design. Permittivity and permeability can be controlled based on the design, thickness, gap, length, and width of unit cell.

Scientists have used several techniques in the design of metamaterial structures spanning from RF to millimeter wave technology for various applications. In this work, we have utilized two of the well-known implementations, namely, complementary MTM structures based on Babinet's principle (Zhang et al. 2013) and metamaterials based on tightly coupled scatterers (Donzelli et al. 2009).

Another type of periodic structures which exhibit spatial filtering in microwave and millimeter wave regions is frequency selective surfaces (FSS). There are two types of FSS, namely, patch-type and aperture-type which show low pass filter and high pass filter characteristics, respectively. Such type of FSS structures is useful in the design of radar absorbers, radomes, and reflector antennas. Generally, LH behavior of metamaterial is the prime aim where applications have known to show extraordinary results due to the changing physical phenomenon (Engheta and Ziolkowski 2006), but we are more interested in designing metamaterial-engineered frequency-selective surface proving sharp roll-off characteristics (outside operating band) for radome applications with positive-ε/negative-μ or negative-ε/positive-μ or both negative MTM structures.

The two major constraints in our design of MTM-FSS structure for radome applications are polarization stability at oblique angle of incidence and single layer requirement as compared to multilayer metamaterial-FSS spatial filter response (Oraizi and Afsahi 2009a; b).

Cube-shaped structures with SRR/wire/composite or any other isotropic design on each face on the cube with 90^o rotation along with high dielectric constant across the gaps have been experimented and theorized to be polarization independent with all axis and angle up to 85^o (Capolino 2009). This behavior is due to quasi-periodical location and higher density of particles. But, realization of 3D structures for radome

applications and its EM characterization is quite complex. Thus, taking into account the limitations of our design, work has been carried out using 2D planar structures. One major point, which needs to be considered, is the metal losses and thickness limitations while working in millimeter wave region.

These limitations can be substantiated using MTM-FSS structures which contain resonating cells exhibiting ε-negative, μ-negative or both negative at resonant frequencies. Sharp roll-off characteristics at out-of-band can be achieved with a combination of these elements designed to bear FSS-like wallpaper characteristics. Also the out of band transition can be reduced to a great extent, thus, reducing the RCS for stealth applications effectively (Narayan and Jha 2015).

Effective Parameter Retrieval for Metamaterial Unit Cell

Application of Babinet's principle has been implied in the design of artificial planar metamaterials (Suwailam et al. 2013) depicting effective negative electric permittivity. Previously, complementary SRRs have been shown to be an alternative to conventional metallic wires to accomplish negative permittivity (ε). Complementary metamaterials show properties identical to their inverse structures. For example, the transmission coefficient of CSRR, t^c, is related to the transmission coefficient of SRR, t as $t + t^c = 1$. Also, Babinet's principle states that their scattered electric and magnetic fields as well as transmission and reflection are interchanged.

The fields scattered by two complementary planar structures of arbitrary shape of thin perfect conducting sheets can be related by Babinet's principle as follows. If we consider an incident electromagnetic (EM) field (E_0, B_0), then its complementary EM field $\left(E_0^c, B_0^c\right)$ is defined as $E_0^c = -cB_0^c$, *and* $B_0^c = E_0^c/c$, which relates to a 90° rotation around the propagation direction. The reflected (E_r, H_r) and transmitted (E_t, H_t) electric and magnetic fields satisfy the condition as discussed. If we have charge distribution on the thin metallic surface, the perpendicular reflected electric field is opposite to the transmitted electric field, i.e., $n.\ E_r = -n.\ E_t$ where n is a vector normal to the metallic surface. The parallel component of electric field is continuous across the thin metallic surfaces. According to the Babinet's principle, there are two related complementary plane structures: One is related to surface electric charge, and the other is related to the magnetic flux associated with complementary structure.

Basically, Babinet's principle defines the relation between field distributions across the direct structure to its complementary structure for EM wave propagating perpendicular to the screen. Both the patch-type hybrid unit cell and its complementary structure have been designed and validated using the Babinet's principle approach. For the patch-type hybrid structure, the excitation is along the axis of I-shaped, i.e., parallel to the central arm of I-shaped and C-shaped structure. Thus, symmetric structure of the resonators with respect to orientation of E-field supports polarization (TE and TM)-independent metamaterials.

Further, *transmission line transfer matrix method* (TLTMM) is used to study the macroscopic behavior of MTM hybrid unit cell for specific angle of incidence, polarization, and frequency of incident plane wave (Oraizi and Afsahi 2009a; b). Here, characteristic impedance and propagation constant is estimated using extracted effective permittivity and permeability parameters. As our design consists of only a single layer of metamaterial unit cell, it can be modeled as a three-layer (air, MTM, air) model. Accordingly based on the computation of field amplitude, transmission matrix, and wave amplitude transmission matrix, reflection and transmission coefficients for each internal layer are calculated and finally transmittance and reflectance graphs are obtained.

In order to retrieve the effective permittivity and permeability of a metamaterial laminate, it is required to characterize it as an effective homogeneous laminate (Caloz and Itoh 2006). Here, the permittivity and permeability of MTM laminate can be retrieved from reflection and transmission coefficient data. For a plane wave impinging normally, thus traversing the thickness, d (either in plane parallel or perpendicular to the MTM cell), S11 is termed as reflection coefficient and S21 is related to the transmission coefficient, T by $S_{21} = Te^{ik_0 d}$, where k_0 represents the free space wave number.

In the literature, several techniques for retrieval of effective metamaterial parameters are reported (Ling et al. 2012; Smith et al. 2005; Szabo et al. 2010). However, these methods produce a scalar electric permittivity and magnetic permeability. While in practical scenario, the electromagnetic behavior of most of the metamaterial structures is anisotropic that requires a tensor to accurately describe it. Even when one is able to determine all the elements of full permittivity and permeability, the model can predict only the far-field behavior whereas the near-field behavior of the metamaterial will vanish. Therefore, a full-wave simulation is required to predict correct EM fields with inclusion of coupling effects between MTM cell and surrounding structures.

In our technique, Kramers-Kronig relations have been used along with the standard well-known S-parameter-based retrieval method (Chen et al. 2004a), to confirm the uniqueness of the solution. For varying thickness limits, the technique is well suited to account for the calculation of refractive index which involves more than *one branch* of complex logarithmic function. However, the technique has two limitations:

1. For the determination of real part of refractive index, a cumbersome iterative method is required, which is based on Taylor series.
2. One cannot find effective material parameters in some frequency regions when standard passive material conditions $\mathrm{Im}(\varepsilon) \geq 0$ and $\mathrm{Im}(\mu) \geq 0$ are imposed.

Instead, as noted in papers cited above, the magnetic and electric dipoles induced in metamaterials depend on each other. Here passivity condition can be satisfied for $\mathrm{Im}(\varepsilon) \leq 0$ and $\mathrm{Im}(\mu) \leq 0$. Relaxing the conditions allows us to calculate the effective parameters even in regime (i.e., resonant frequency band). The procedure of the permittivity and permeability extraction technique is described below:

Step 1: Wave impedance is uniquely determined from S-parameters for a plane wave incident as described above.

$$S_{11} = R_{01}\left(1 - e^{i2N_e k_0 d_e}\right)/\left(1 - R_{01}^2 e^{i2N_e k_0 d_e}\right) \tag{3}$$

$$S_{21} = \left(1 - R_{01}^2\right)e^{iN_e k_0 d_e}/\left(1 - R_{01}^2 e^{iN_e k_0 d_e}\right) \tag{4}$$

where $R_{01} = (Z_e - 1)/(Z_e + 1)$ and Z_e is the complex wave impedance.

$$Z_e = \sqrt{\left((1 + S_{11})^2 - S_{21}^2\right)/\left((1 - S_{11})^2 - S_{21}^2\right)} \tag{5}$$

$$e^{iN_e k_0 d_e} = S_{21}/1 - S_{11}R_{01} \tag{6}$$

The sign of wave impedance is estimated by assuming the conditions,

$$Re\,(Z_e) \geq 0 \tag{7}$$

$$Im(N_e) \geq 0 \tag{8}$$

Or equivalently, $\mid e^{iN_e k_0 d_e} \mid \leq 1$.

Step 2: The complex refractive index, N_e, can be expressed as,

$$N_e(\omega) = n_e(\omega) + iK_e(\omega) \tag{9}$$

where n_e is the refractive index and K_e represents the extinction coefficient.

The complex refractive index can be calculated as,

$$N_e = \frac{1}{k_0 d_e}\left\{Im\left[\ln\left(e^{iN_e k_0 d_e}\right)\right] + 2m\pi - Re\left[\ln\left(e^{iN_e k_0 d_e}\right)\right]\right\} \tag{10}$$

where m is an integer denoting the *branch index*.

Step 3: Separating the real and imaginary parts of the above expression, the refractive index and extinction coefficient becomes,

$$n_e = \frac{1}{k_0 d_e}\left\{Im\left[\ln\left(e^{iN_e k_0 d_e}\right)\right] + 2m\pi\right\} = n_e^0 + f\frac{2m\pi}{k_0 d_e} \tag{11}$$

$$K_e = -Re\left[\ln\left(e^{iN_e k_0 d_e}\right)\right]/k_0 d_e \tag{12}$$

where n_e^0 represents the refractive index w.r.t. *principal branch* of logarithmic function. Here, the advantage comes into picture once the branches of logarithmic function do not affect the imaginary part of refractive index given in the above

eq. (11). Now, we can determine the real part of the refractive index from the imaginary part by implying Kramers-Kronig relation,

$$n^{kk}(\omega') = 1 + \frac{2}{\pi} P \int_0^{\infty} \frac{\omega K_e(\omega)}{\omega^2 - \omega^2} d\omega \tag{13}$$

where P denotes the principal value of improper integral.

The integration of the above equation can be determined numerically with the help of trapezoidal rule of integration. In order to avoid singularity of improper Kramers-Kronig integral, the integral is performed using Cauchy's method as follows,

$$\Psi_{i,j} = \frac{\omega_j K_e(\omega_j)}{\omega_j^2 - \omega_i^2} + \frac{\omega_{j+1} K_e(\omega_{j+1})}{\omega_{j+1}^2 - \omega_i^2} \tag{14}$$

$$n^{KK}(\omega_i) = 1 + \frac{\Delta\omega}{\pi} \left(\sum_{j=1}^{i-2} \Psi_{i,j} + \sum_{j=i+1}^{N-1} \Psi_{i,j} \right) \tag{15}$$

Using the above equations, the branch number "m" can be calculated as,

$$m = Round \left[\left(n^{KK} - n_e^0 \right) \frac{k_0 d_e}{2\pi} \right] \tag{16}$$

where the function, *Round* [], rounds to the nearest integer. Further, the branch number m is substituted in eq. (11), and the exact value of refractive index is calculated.

Generally, there will be a sharp discontinuity at the end of the limit of frequency range, and this can be removed by increasing the frequency interval of simulation. But if it still persists, that is an indication that effective medium theory reaches its limit and effective parameters in the region above the discontinuity cannot be calculated accurately.

Step 4: Finally, the effective permittivity and permeability can be calculated as,

$$\mu_{eff} = N_e Z_e; \quad \varepsilon_{eff} = N_e / Z_e \tag{17}$$

A suitable approximation for real part of the refractive index is achieved with Kramers-Kronig relations. The signs of the real and imaginary parts need to be taken into account (Woodley and Mojahedi 2010). Also, when the thickness of metamaterial is compared to the operating wavelength, it will include more than one branch to approximate. But, the effective medium theory cannot be applied when the optical thickness of metamaterial becomes comparable to the operating wavelength, as its limit is reached.

This discontinuity can further be removed by *phase unwrapping technique* (Barroso and Hasar 2012). They had pointed out specifically that the cause of

discontinuity is due to improper calculation of branch indices. They expressed refractive index as follows,

$$n = \frac{1}{k_0 d}(-iln|T|+\Phi),\ -\pi < \Phi < \pi \tag{18}$$

$$n = \frac{1}{k_0 d}(-iln|T|+2m\pi), \pi \leq \Phi \leq -\pi \tag{19}$$

In Eq. (19), 2mπ is added to the complex transmission coefficient expression, T = T exp (iΦ) = exp (ink$_0$$d$), to make the logarithmic function single valued and continuous at T. The resonant frequency at which the phase of T jumps from -π to π is constantly unwrapped by shifting the intervals -2π, - 4π, - 6π. . . . . - 4π, - 6π. Thus, once complex transmission coefficient T has been determined, we can again use the abovementioned technique to calculate the corresponding refractive index without considering any discontinuities over the identified frequency range.

Few points that need to be kept in mind while retrieving parameters (Hasar et al. 2012; Andrei et al. 2010) are as follows:

1. Determination of the port boundary from MTM cell. In case of one-dimensional symmetrical MTM cell, the determination of the boundary can be done as described in section B (Chen et al. 2004b). For two-dimensional and asymmetric 1D metamaterials, it is required to determine effective boundaries of the slab from optimization. However, in our case, where wave propagation is in the plane perpendicular to slab, the distance from the port should be optimized (0.011–0.5 mm) as effective parameters are due to near-field excitations. Moreover, the dimensions of the box in the plane where MTM cell is located should be optimized accordingly (0.1–0.8 mm). However, phase can always be adjusted.
2. Even if you need to find effective parameter of any array or multilayer structure, it should be done using *wave port* instead of *Floquet port* for an infinite array of unit cell. For a strong resonant metamaterial, it is required to find out the locations of effective boundaries between the laminate and air. The EM wave propagating through metamaterial laminate will not behave like plane wave due to non-uniform scattering fields produced by induced currents flowing across metallic unit cell. Also, retrieving parameters at oblique angles of incidence can be accomplished (Menzel 2008).

Design of Patch/Aperture Type Hybrid MTM Element-Based FSS Unit Cell

In order to realize high-performance broadband and multiband radomes, it is essential to design hybrid metamaterial element-based FSS. This is due to the fact that with hybrid unit cell, broadband and multiband characteristics can be achieved with single-layer configuration apart from cascaded MTM-FSS structure. In view of this,

the detailed design and analysis of patch−/aperture-type hybrid MTM element-based FSS is discussed in the subsequent sections.

Patch-Type I-Shaped Hybrid Metamaterial-FSS Unit Cell

In this work, the analysis of I-shaped hybrid metamaterial cell consisting of two C-shaped resonators separated equally by the middle resonator has been carried out as shown in Fig. 1. Generally, all metal conductors have electric plasma frequency in the ultraviolet region. But arranging the cut wire pairs in an array considerably reduces the plasma frequency in the microwave region due to electron density and effective mass drop along with spatial dispersion effects. Thus in the proposed design, cut wire pairs are combined according to spatial filter requirement to lower the electric plasma frequency.

The dimensions of the I-shaped (composed of three thin wire structures) resonator have been optimized for X-band resonance. Excitation of C-shaped resonators combined with two opposite arms of I-shaped resonator provides the resonance at Ka-band.

The transmission and reflection characteristics in the proposed patch type-hybrid unit cell backed with Rogers RT5880 ($\varepsilon_r = 2.2$, $tan\delta = 0.0009$) with height of dielectric substrate, 0.254 mm are studied at normal incidence as shown in Fig. 2. The proposed structure exhibits dual bandstop characteristics at the designed frequencies. The dimensions of central I-shaped resonator are in accordance with the center frequency of 10.2 GHz and the C-shaped resonators are optimized to resonate at the higher band at central frequency of 31.8 GHz.

With reference to the default dimensions of the hybrid MTM unit cell, it is essential to note that the lateral dimensions of the unit cell (L × W), at the fundamental resonant frequency, 10.2 GHz, equal $L = \lambda/5.77$ and $W = \lambda/8$, ie., they are significantly smaller than the wavelength. All parameters such as periodicity in x- and y-direction, thickness of the dielectric substrate, gap between two-unit cells in horizontal as well as vertical direction, and width of the arms have a definite effect on the effective material parameters as well as resonant frequency.

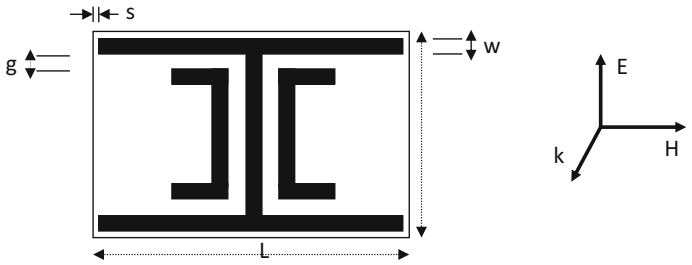


Fig. 1 Schematic of hybrid patch-type MTM-FSS unit cell

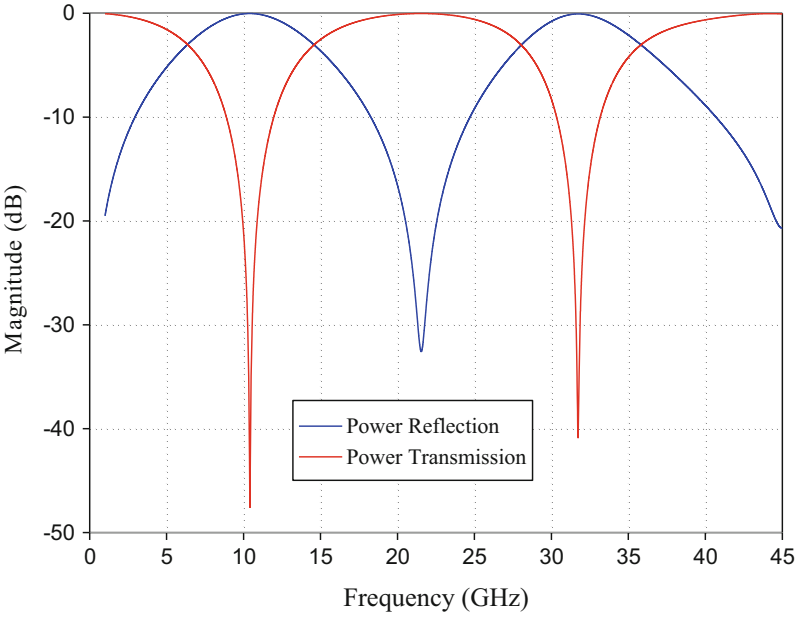


Fig. 2 Transmission and reflection characteristics of patch type hybrid MTM-FSS backed with dielectric at normal incidence using TLTMM

Aperture-Type Hybrid MTM-FSS Unit Cell

The main objective of this work is to design a dual band-pass in X and Ka-band region which can be useful for radome applications, and authors investigate in the same direction to get enhanced EM performance characteristics. Thus, applying Babinet's principle as discussed above, we have successfully computed the transmission and reflection characteristics for the aperture-type hybrid unit cell, i.e., complementary of the above shown unit cell (Fig. 1). The proposed aperture type hybrid MTM-FSS unit cell is shown in Fig. 3 with design parameters same as discussed in Table 1.

The gap between the unit cells in an array of infinite structures in both the directions has been maintained to be 0.2 mm which considerably involves the effect of mutual coupling along with the intraelement unit cell coupling. Also, increasing the periodicity in vertical direction shifts the resonant frequency to the higher frequency region and vice versa. The length of the middle arm of the I-shaped resonator is inversely proportional to the shift in the lower resonant frequency while the length of C-shaped resonators is inversely proportional to the shift in higher band frequency, and hence corresponding shift in the transmission band. At specific resonant frequencies, E-field excites a combination of various elements due to the asymmetry of the structure in horizontal (X) and vertical (Y) direction. Therefore, the MTM-FSS pattern is polarization-dependent with E-field perpendicular to the axis of I-shaped resonator as shown in Fig. 3. The transmission characteristics of the

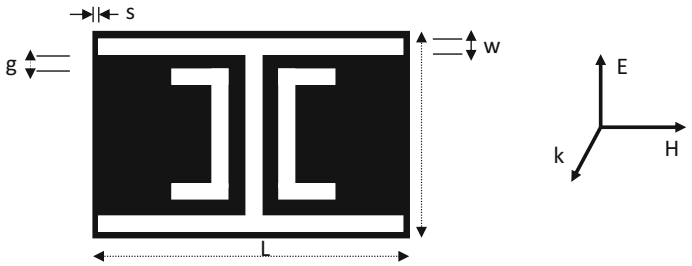


Fig. 3 Schematic of hybrid aperture-type MTM-FSS unit cell

Table 1 EM design parameters of patch-type hybrid MTM-FSS unit cell

Design parameters	Optimized dimension (mm)
Periodicity (*dx*); *L*	5.20
Periodicity (*dy*); *W*	3.754
Width of the arms of hybrid unit cell; *w*	0.254
Gap between two hybrid unit cells	0.20
Gap between I and C-shaped resonator; *g*	0.254
Gap between horizontal/vertical I-shaped and unit cell; *s*	0.10

aperture-type hybrid MTM-FSS unit cell have been validated using both full wave analysis method and TLTMM method at various angles of incidence at specific polarization (i.e., TM) as shown in the subsequent figures.

The extraction of effective parameters from the transmission (S21) and reflection (S11) coefficient is discussed in subsection "Effective Parameter Retrieval for Metamaterial Unit Cell," and for simultaneous application of TLTMM to validate the EM characteristics simulated from full-wave analysis (CST Microwave Studio), the work flow chart is given below:

As depicted in Fig. 4, the lower X-band transmission region extends from 10.21 GHz to 11.868 GHz with central frequency at 11.04 GHz, and the upper Ka-band transmission region extends from 31.84 GHz–33.65 GHz with central frequency at 32.75 GHz with 1.658 GHz and 1.806 GHz bandwidth, respectively, w.r.t 0.5 dB transmission loss. Thus, operating in dual passband with transmission efficiency greater than 90% is the desired bands. Moreover, this structure also exhibits superior wideband reflection characteristics in the region between two passbands as can be visualized from Fig. 5. Metamaterials, also known as artificial subwavelength resonator, are shown here to exhibit strong resonance in X-band and Ka-band with a wide band separation. Hence, we characterize these resonators as metamaterial-FSS structures due to their operation as spatial filter in a similar fashion as well-known *frequency selective surfaces (FSS)*.

Figure 6 shows that the proposed hybrid aperture MTM-FSS unit cell shows ε-negative/μ-positive behavior close to 18.64 GHz, which is lower than the electric plasma frequency as can be figured out from power transmission graph shown in Fig. 4. For waves propagating normally to the hybrid-MTM/FSS unit cell structure,

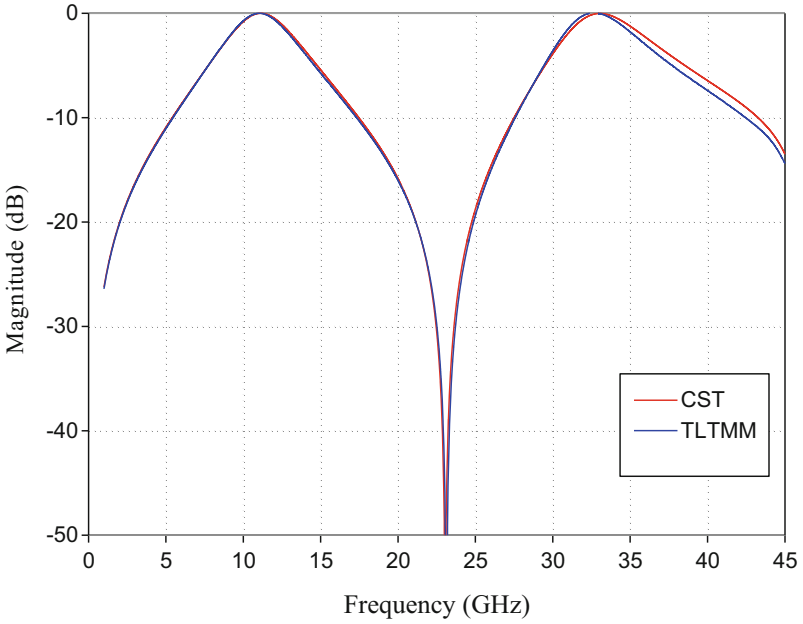


Fig. 4 Transmission characteristics of hybrid aperture-type MTM-FSS unit cell for TM polarization at normal incidence using full-wave and TLTMM method

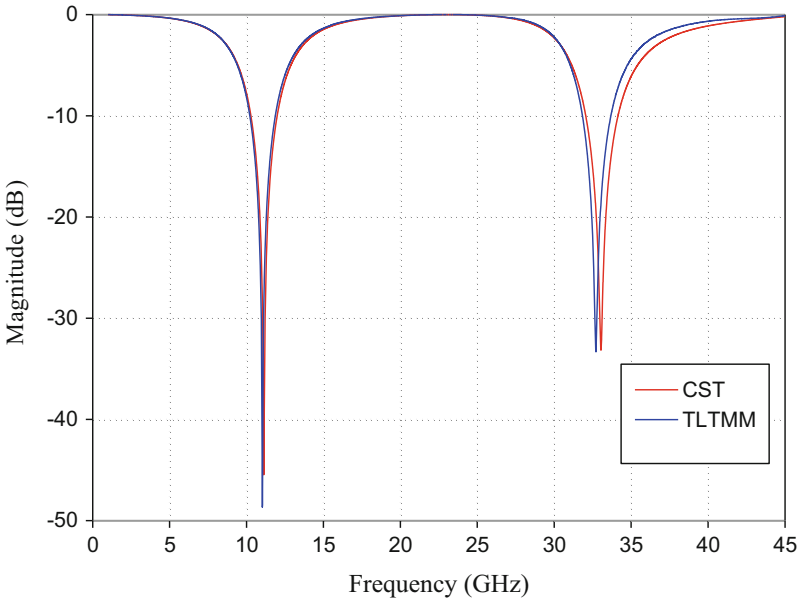


Fig. 5 Reflection characteristics of hybrid aperture-type MTM-FSS unit cell for TM polarization at normal incidence using full-wave and TLTMM method

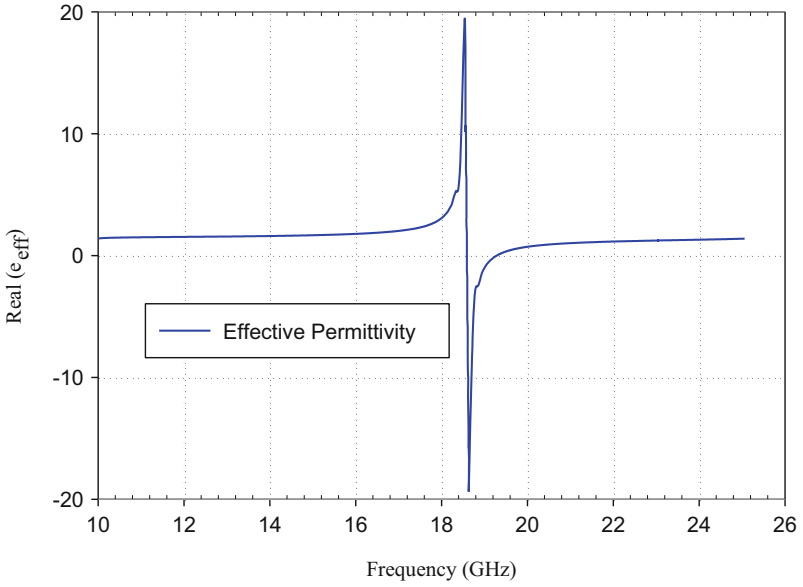


Fig. 6 Extracted material characteristics of hybrid aperture-type proposed MTM-FSS unit cell

and in absence of losses, a homogeneous effective MTM cell exhibits negative effective permittivity at $f < f_P$, and such a medium shows artificial dielectric as opaque whereas for $f > f_P$, effective permittivity is positive, and it shows translucent behavior. Another important property of the proposed structure is the variation of the electric resonant frequency based on the interaction between the unit cells (i.e., between the horizontal arms of I shaped). The spacing between the arms can be reduced or strip widths at the edge can be increased in order to achieve the interaction which results in an additional inter-pair capacitance, thus influencing the electric resonant frequency.

Taking into account the possibility of utilizing the structure for radome applications, it has also been validated at 30^{o} and 45^{o} oblique incidence at TM polarization due to its anisotropic nature. In the subsequent graphs, the validation from FDTD-based full-wave analysis simulation software and TLTMM method has been shown. Further, attempts have been made in the direction of making the structure polarization independent with respect to TE/TM polarization at oblique angle of incidence, discussed briefly in the next section (Figs. 7, 8, and 9).

To summarize, the band separation between the two bands as shown above can be controlled with optimization of I-shaped and C-shaped resonators independently. However, intraelement mutual coupling should not be ignored when dealing with structural optimization. One can also get a clear picture with the help of surface current distribution and excited E-/H-field about inter and intra mutual coupling. With the increase in length of both the resonators, band shifts toward higher frequency side and vice versa. Transmission resonance for thicker substrates occurs

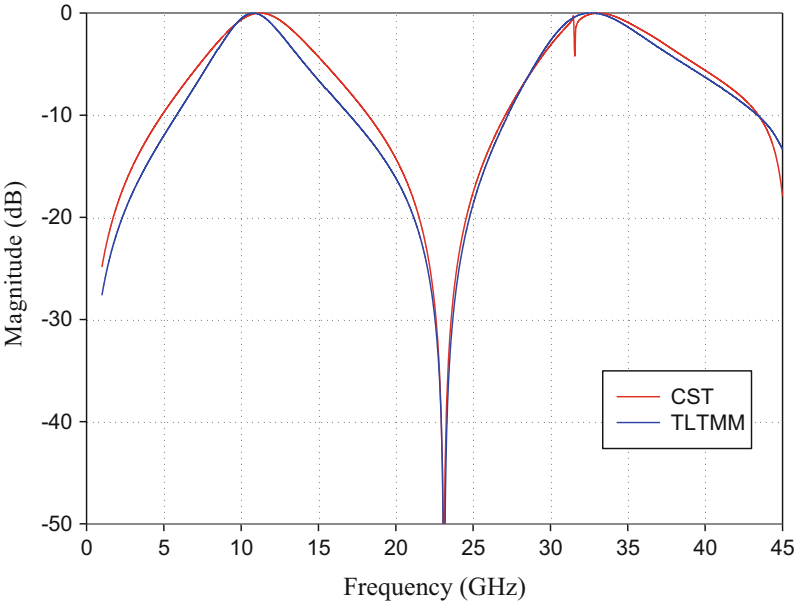


Fig. 7 Transmission characteristics of hybrid aperture-type MTM-FSS unit cell for TM polarization at 30° using full-wave and TLTMM method

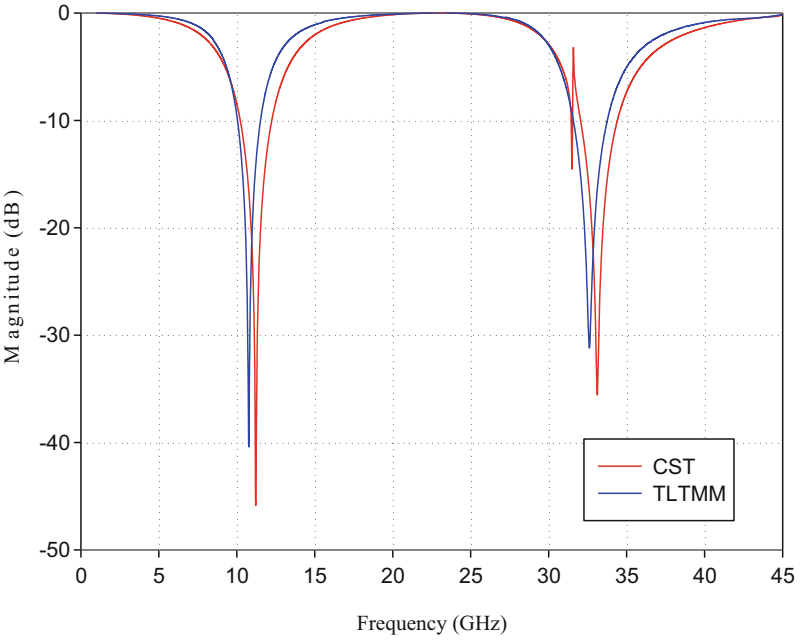


Fig. 8 Reflection characteristics of hybrid aperture-type MTM-FSS unit cell for TM polarization at 30° using full-wave and TLTMM method

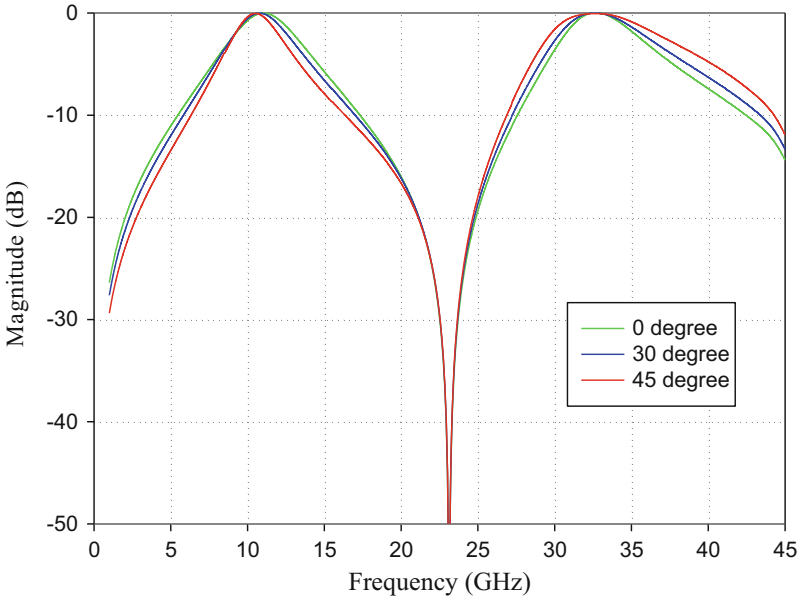


Fig. 9 Transmission characteristics of hybrid aperture-type MTM-FSS unit cell for TM polarization using TLTMM method at 0^{o}, 30^{o}, and 45^{o}

at lower frequencies which leads to increased bandwidth. Inductance present in the unit cell is due to the surface current induced, and capacitance is due to the E-field path. If transmission greater than unity is achieved for a structure with no active elements and gain, then it indicates an error. This fault can be explicated by lensing effect together with experimental error caused by the entire size of the structure. For a strong resonant metamaterial, the spacing between the effective boundaries of the slab and air needs to be determined. However, polarization-independent MTM-FSS structures are essential for radome applications, which is discussed in the following section.

Polarization-Independent Hybrid MTM-FSS Unit Cell

Further, attempts were carried out to make the structure polarization independent (TE/TM). The property of skew symmetry in which one unit cell consists of four cells which are diagonally symmetric with 90^{o} rotation as shown in Fig. 10. The concept derives from the current distribution and the E-field excitation (Feng et al. 2015). For one polarization, one set of diagonal elements will excite while for the other polarization, another set of diagonal with 90^{o} rotated cells will excite, thus maintaining the symmetry behavior.

Figure 11 depicts the power transmission with respect to both TE and TM polarizations at incidence angle 30^{o}. It can be concluded that asymmetrical

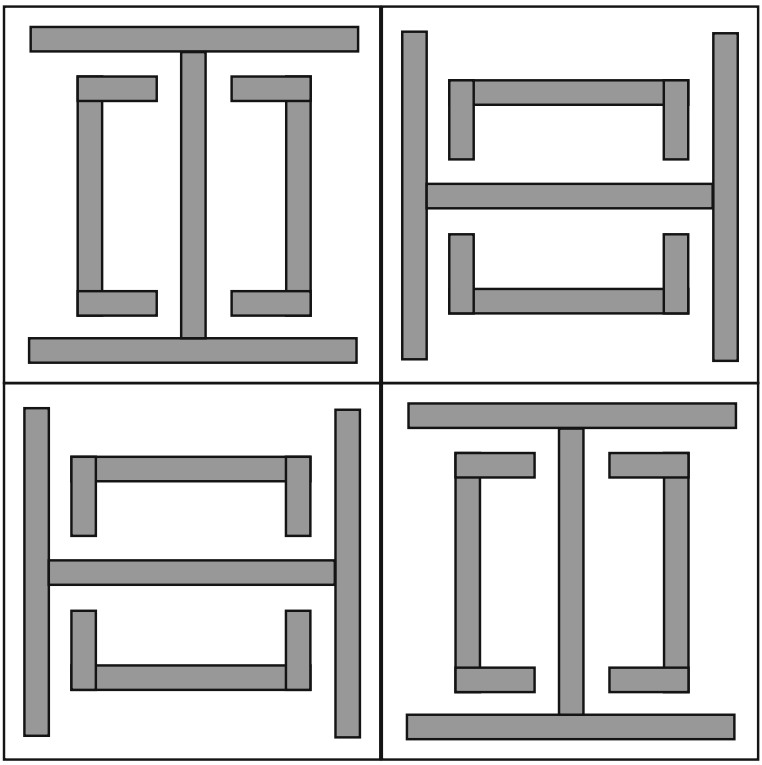


Fig. 10 Schematic of hybrid MTM-FSS unit cell backed with glass epoxy ($\varepsilon_r = 4$, $tan\ \delta = 0.015$) with skew symmetry (height of substrate =0.254 mm)

anisotropic structure can also be made polarization independent with the application of skew symmetry property as shown. However, limitations are always present. At oblique angle of incidence with infinite periodic structures, this property can deviate from its standard EM behavior and the distortions in the form of sharp discontinuities (kink) can be observed at higher angle of incidence at specific polarization. Hence, this geometry was not taken into consideration due to stringent EM performance requirement. Also, more symmetrical structures such as standard single and double SRRs which are dependent on direction of E-field propagation along the axis of SRR can be made polarization independent up to a large extent with the help of this skew symmetry and rotation technique.

Tightly Coupled Conductors-Based MTM-FSS Unit Cell

The concept of coupled particle approach dates back to 2005, where an alternative to SRR and wire design, a pair of metallic Nano rods, was investigated to provide a diamagnetic response. Yet, this concept was not feasible in the design of NRI

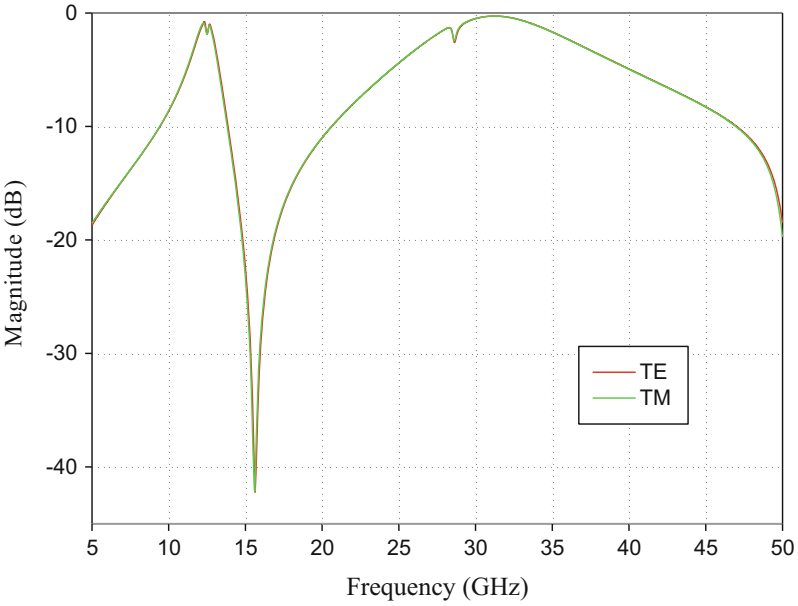


Fig. 11 Transmission characteristics of four element skew symmetry hybrid-FSS unit cell for TE/TM polarization at incidence angle 30°

material due to wide electric resonance of strips as compared to magnetic one. Thus, it is required to find the magnetic resonance within the negative permittivity region to make it NRI for that particular band. However, separate tuning of electric and magnetic resonance using short pairs is not feasible as it is controlled by a single parameter (length of the wires). Hence, novel structures where we can independently control the electric and magnetic resonant frequencies have been investigated in literature (Ziolkowski 2003; Chen et al. 2004a, b; Vallecchi et al. 2009; Yurduseven et al. 2011).

The physical significance behind the pair of tightly coupled resonators is crucial to our design. The electric resonant frequency (f_e) of this type of structure mainly depends on the inter- element capacitance which creates electric dipole moment due to the symmetric currents on top and bottom patch. lt exhibits a much wider stop band at its electric resonant frequency. On the other hand, the magnetic resonance frequency (f_m) depends on the intraelement capacitance formed between top and bottom pairs of resonators and the self-inductance of the pairs. Hence, magnetic resonant frequency is attributed to the resonant TL mode or the asymmetric mode due to the asymmetric currents flowing on top and bottom conductors. f_m is strongly influenced if the width of the arms of the design is comparable to the dielectric thickness separating the two conductors. When the strips become narrower enough, (<H), f_m shifts toward high frequency side due to the reduction of capacitance between the above and below conductor. Thus, if arms are thin, it can prove to be an advantage for achieving NRI material. With the increase in spacer dielectric

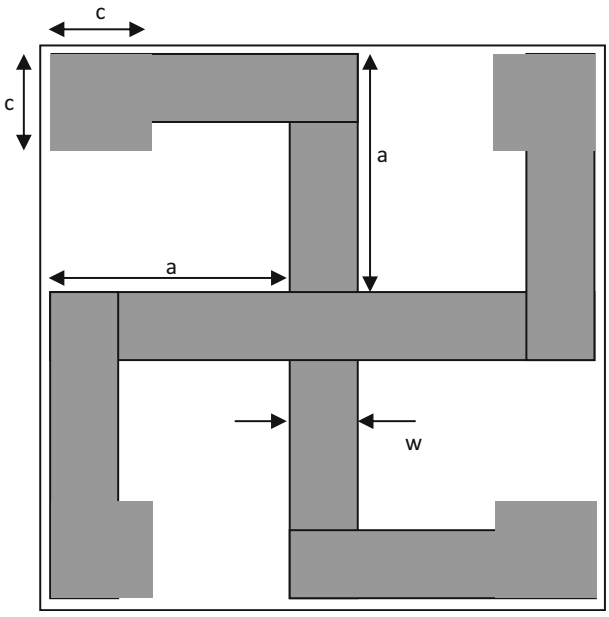


Fig. 12 Schematic of top view of tightly coupled conductors-based hybrid MTM unit cell

permittivity, both electric and magnetic resonances shift toward lower frequencies. This technique offers an extra degree of geometrical parameter reconfigurability with planar design as compared to previously proposed structures in order to achieve NRI/LH metamaterial.

Figures 12 shows the top view of the hybrid MTM unit cell, and Fig. 13 shows the 3D view of the proposed unit cell. The dielectric substrate used for the proposed design is Arlon ($\varepsilon_r = 3.42, \quad tan\delta = 0.0035$), and the dimensions of MTM unit cell are given in Table 2. The proposed structure consists of two overlapping arms with an inverted image constituting the capacitance between the two metallic conductors on each side of substrate. In the vicinity of the applied magnetic field, the capacitances between these conductors enable the induced current to flow around the unit cell.

The transmission and reflection characteristics of the proposed MTM hybrid unit cell at normal incident wave are shown in Fig. 14, followed by the application of retrieval procedure to retrieve effective parameters as shown in Fig. 15. If the capacitance between the centre arms of proposed overlapping inverted MTM cell is small or does not exist, then the magnetic resonant frequency will be separated from the electric resonant frequency and an LH transmission band will not be possible. There exists a magnetic resonance around the transmission peak at 10.35 GHz with asymmetric currents flowing on top and bottom conductor and the electric resonant frequency at 9.38 GHz with a narrow band of negative permeability existing from 10.2 GHz–10.46 GHz while a band of negative permittivity exists from 10.3 GHz–10.45 GHz. The current flowing through the top and bottom surface of the unit cell in asymmetric mode is shown in Fig. 16. Thus, the narrow negative

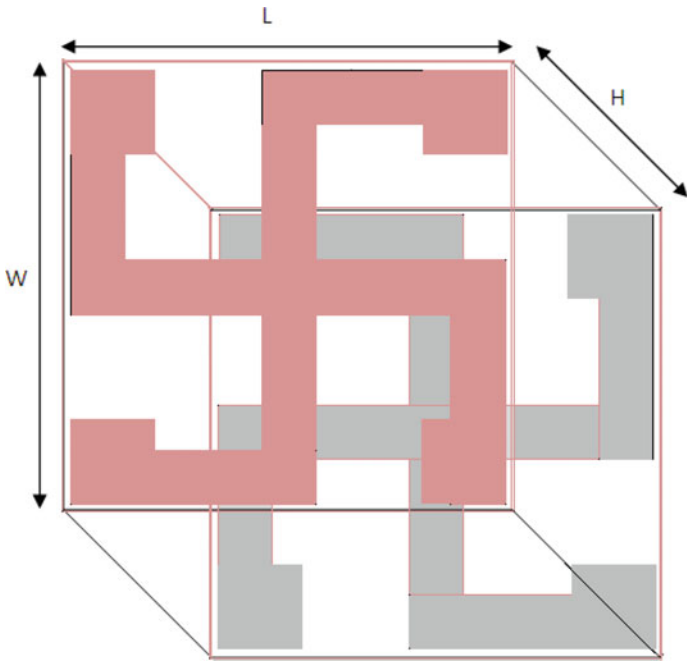


Fig. 13 Schematic of 3D view of MTM element consisiting of two hybrid unit cells placed on each side of the dielectric substrate

Table 2 Dimensions of tightly coupled hybrid patch MTM unit cell

Design parameters	Dimensions (mm)
Periodicity (*dx*); L	8.3
Periodicity (*dy*); W	8.3
Width of the conductor; w	1.0
Spacing between two unit cells	0.3
Substrate height with vertically placed inverted unit cells on top and bottom.; H	0.3
Length of extended arm from the center; a	3.5
Side of square-shaped patch at the edges; c	1.5

permeability band existing around the resonance has been made to overlap the negative permittivity band in order to achieve NRI material. Also, the bandwidth of LH transmission band can be increased by increasing the height of the substrate, but the band will shift toward lower frequency side up to a certain limit and will gradually shift toward higher frequency side upon further increase in the substrate height (H).

Here, the metamaterial structure is shown to exhibit negative permeability for the wave incident perpendicular to the plane of metamaterial slab which offers an

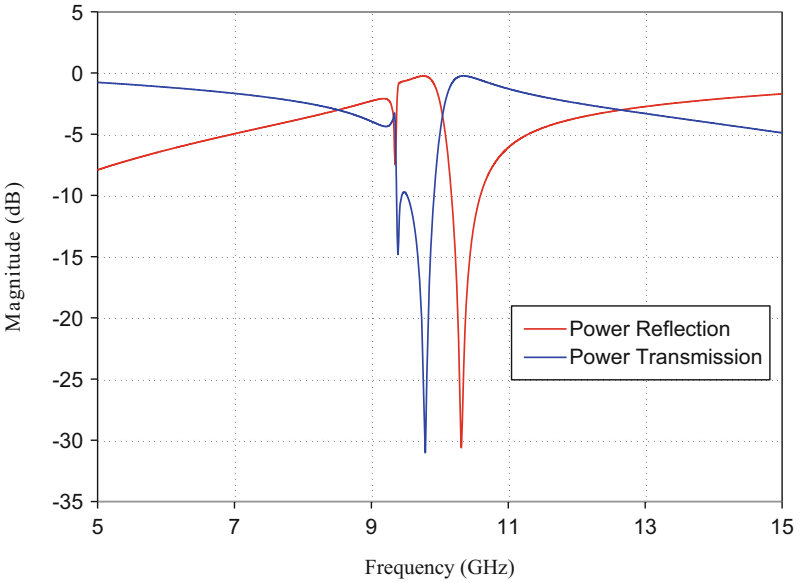


Fig. 14 Transmission and reflection characteristic of the proposed LH MTM unit cell using TLTMM method at normal incidence

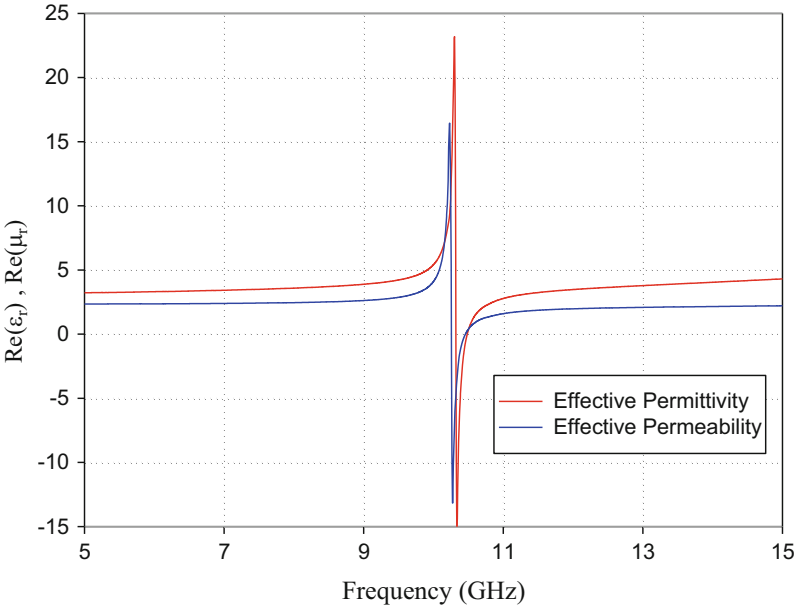


Fig. 15 Extracted effective material parameters of single-layer tightly coupled inverted proposed LH MTM unit cell

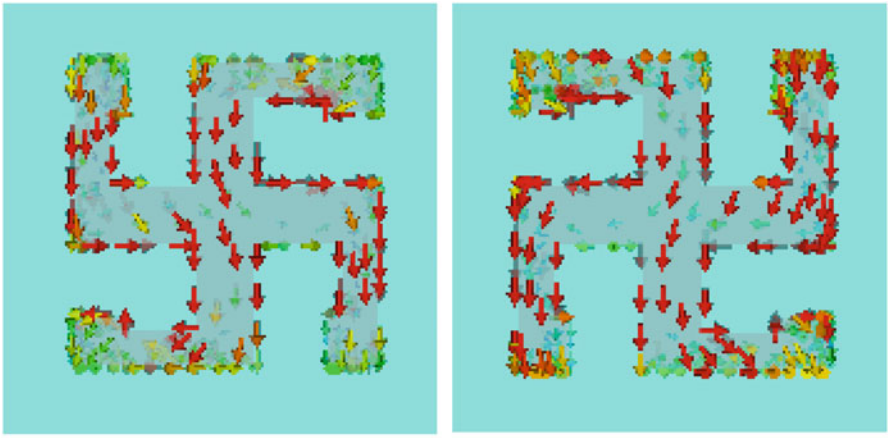


Fig. 16 Asymmetrical current distribution on a pair of tightly coupled proposed hybrid MTM cells (**a**) top surface (**b**) bottom surface near f_m

advantage over the standard SRRs, double SRRs, Labyrinth resonator, and omega-shaped resonator where magnetic field perpendicular to the MTM slab is required for inducing magnetic moment into the cell. Also, composite wire structure is not needed to accomplish separate negative permittivity.

Thus, the proposed hybrid LH MTM cell exhibits isotropic behavior at oblique angle of incidence over a narrow band of around 350 MHz as shown in Fig. 17. Due to the bandwidth limitation, steps have been taken to achieve FSS-type MTM-based element with improved performance for radome applications. Thinner substrates have high attenuation due to closer field confinement to the substrate between the conductors which leads to increased incident wave dissipation. However, the effect of losses is rather negligible for thicker substrates at higher frequencies, and it does not adversely affect the fundamental properties of the design characteristics.

Hybrid MTM-Based FSS Structure with Shorted Edges

In order to achieve flat band-pass characteristics, the design has been further modified by employing short-circuit strips at the open-ended edges of proposed tightly coupled MTM-FSS structure as shown in Fig. 18. This structure consists of a pair of similar conductors on the top and bottom of the dielectric substrate ($\varepsilon_r = 3.42$, $tan\delta = 0.0035$). All the other design parameters are same as shown in Table 2 except for the periodicity (dx = dy = 9.5 mm) and height of the substrate which is optimized to minimize insertion loss at the transmission band as 1.8 mm.

The magnetic resonance frequency is significantly increased by the introduction of short circuit strips, although the resonance frequency is marginally disturbed due to the symmetric flow of currents on conductors on both sides of the substrate. This is due to reduction in the capacitance between the pair of tightly coupled conductors. Thus, negative permittivity and permeability regions occurred at two distinct

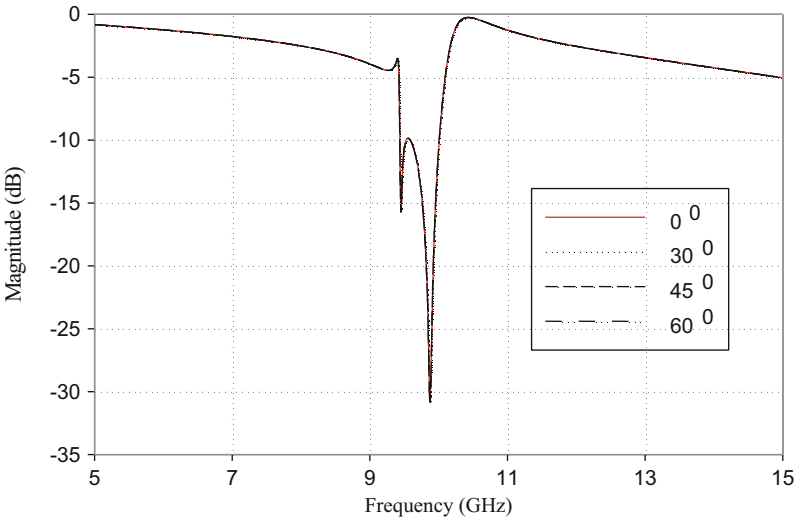


Fig. 17 Transmission characteristics of proposed LH MTM unit cell for TE polarization at oblique angle of incidence. (Unit cell dimensions for computation: 0.8 × 0.8 × 2 mm)

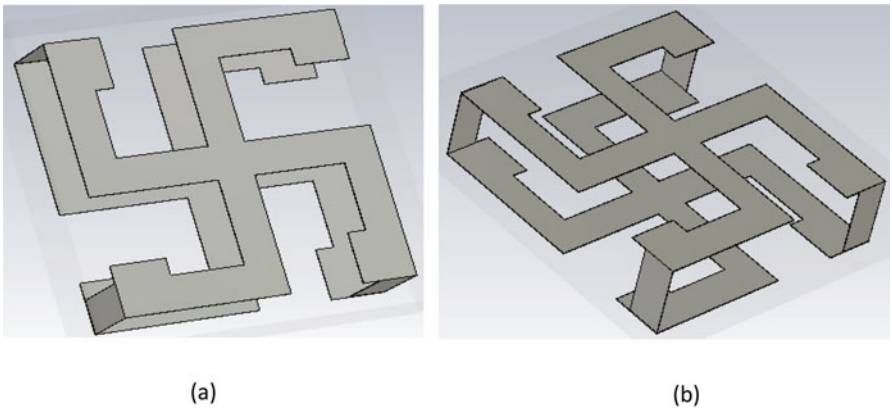


Fig. 18 (**a**) Schematic top view and (**b**) 3D view of the proposed MTM-based FSS unit cell with shorted strips at the edges

frequencies unlike LH band. The extracted effective parameters are depicted in Fig. 19 such that the unit cell box dimensions, 10.5 × 10.5 × 2.8 mm, are taken into consideration for computation and extraction. The shorted structure exhibits a strong electric resonance when compared to magnetic resonance as can be visualized from Fig. 19. Moreover, electric and magnetic resonances are shifted in opposite directions. The region between both the resonances crossing each other is the transmission band which was quite narrow in case of previously proposed tightly coupled inverted LH MTM structure.

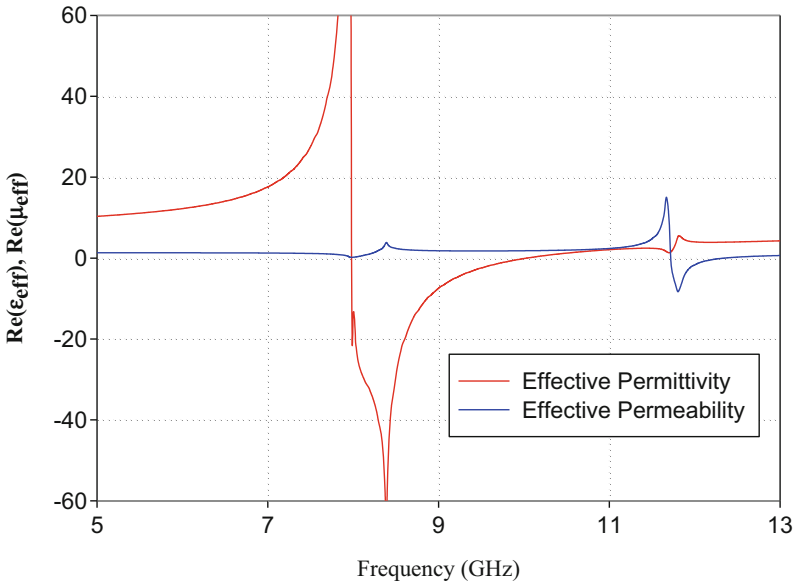


Fig. 19 Material characteristics of the proposed shorted hybrid MTM-FSS element extracted from its transmission and reflection characteristics

Figures 20 and 21 depict the validation of CST-based simulated and TLMM-based computed power transmission characteristics at normal and 30° incidence angles for TE polarization. Further, it is observed from Fig. 20 that the proposed structure exhibits a bandwidth of 1 GHz from 10.43 GHz to 11.435 GHz corresponding to −0.5 dB insertion loss at normal incident.

In view of radome application, the EM characteristics of proposed structure have been performed at oblique angle of incidence up to 45° for TE polarization with high transmission characteristics such that the radome does not degrade the performance of the antenna (Fig. 22). Further, reflection characteristics are also studied at oblique incidence for TE polarization shown in Fig. 23.

Figures 24 and 25 show the power transmission characteristics of proposed shorted MTM-FSS structure at oblique angles of incidence (30° and 45°) for TM wave polarization that exhibit similar bandwidth (1 GHz w.r.t. 0.5 dB transmission loss) as TE case. Thus the proposed shorted MTM- FSS unit cell structure exhibits isotropic behavior due to its symmetry in X-and Y-direction and hence is well-suited for airborne radome application in X-band.

Comparison Between Double Square Loop FSS and Proposed Shorted MTM-FSS Structure

On further investigation, the proposed shorted MTM-FSS was found to show superior performance compared to known double square loop FSS (Luo et al. 2005) in terms of bandwidth and sharp roll of characteristics. The proposed cell approximately mimics the behavior of ideal bandpass characteristics and, thus, can be utilized for radome applications with reduced RCS for stealth technology in the

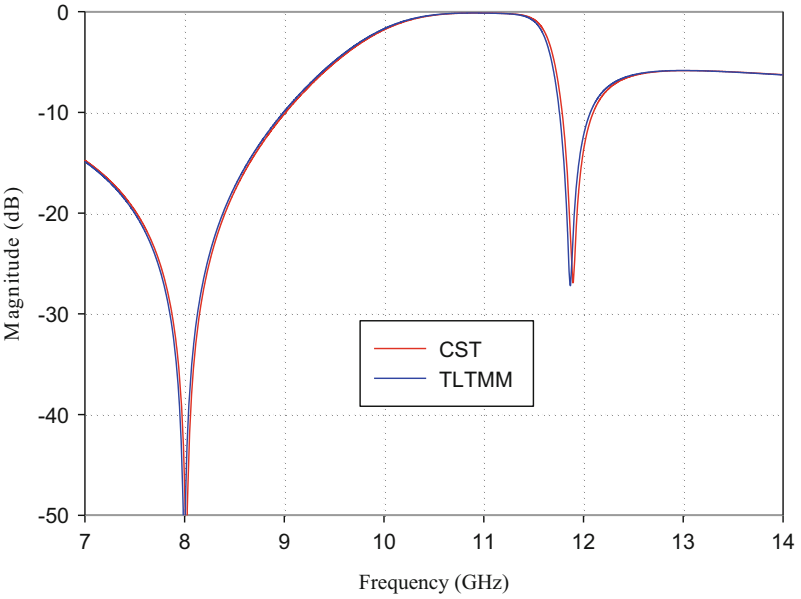


Fig. 20 Transmission response of hybrid-shorted MTM-FSS unit cell using CST and TLTMM method at normal incidence

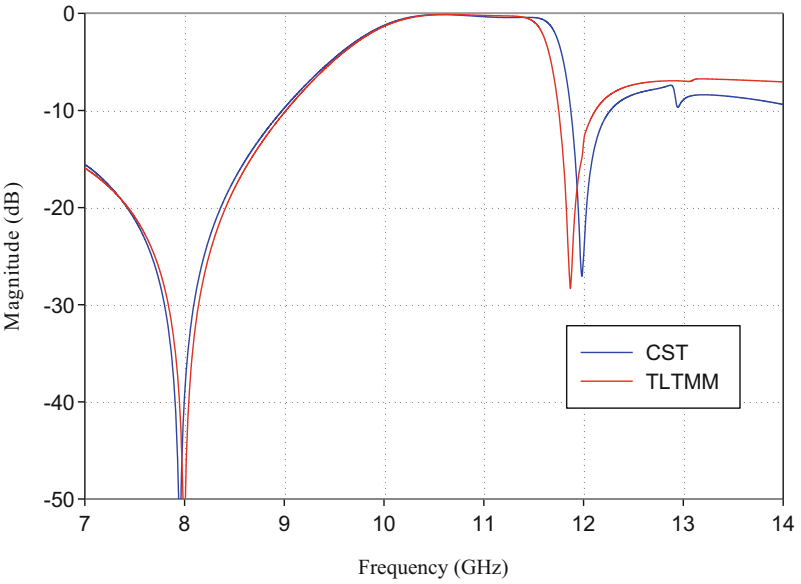


Fig. 21 Transmission response of hybrid-shorted MTM-FSS unit cell for TE polarization at 30^{o} using CST and TLTMM method

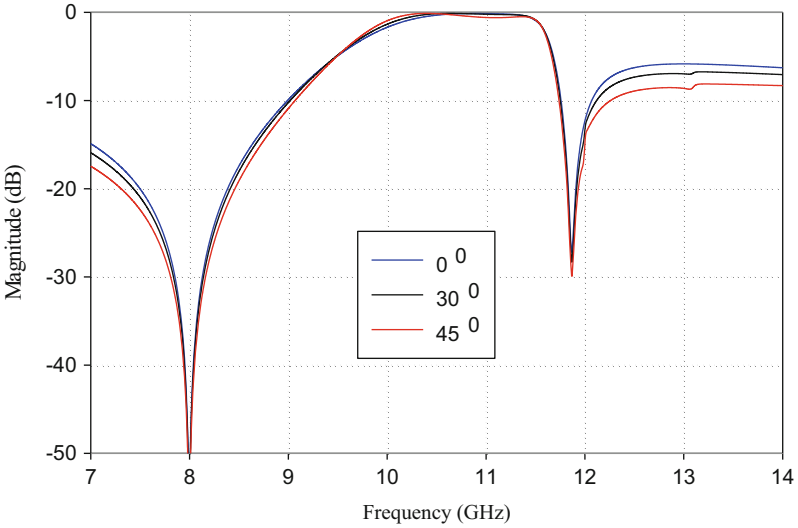


Fig. 22 Stability of transmission characteristics of hybrid-shorted MTM-FSS unit cell for TE polarization at various incident angles (0°, 30°, and 45°)

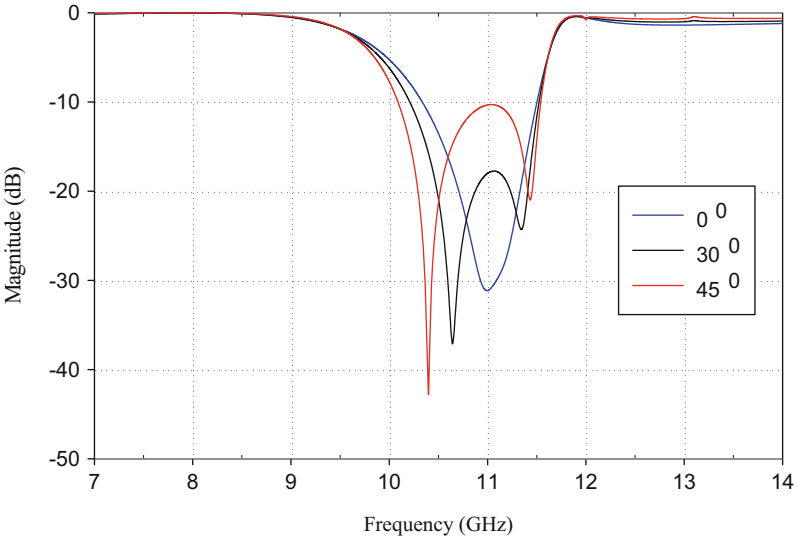


Fig. 23 Reflection characteristics of proposed hybrid-shorted MTM-unit cell at various angle of incidence (0°, 30°, and 45°) for TE polarization using TLTMM method

X-band. To get the desired accurate mid transmission frequency band in the X-band region, the structure was optimized with slight changes in flexible design parameters as depicted in Table 3 when compared to Table 2 to shift the transmission band to lower frequency side.

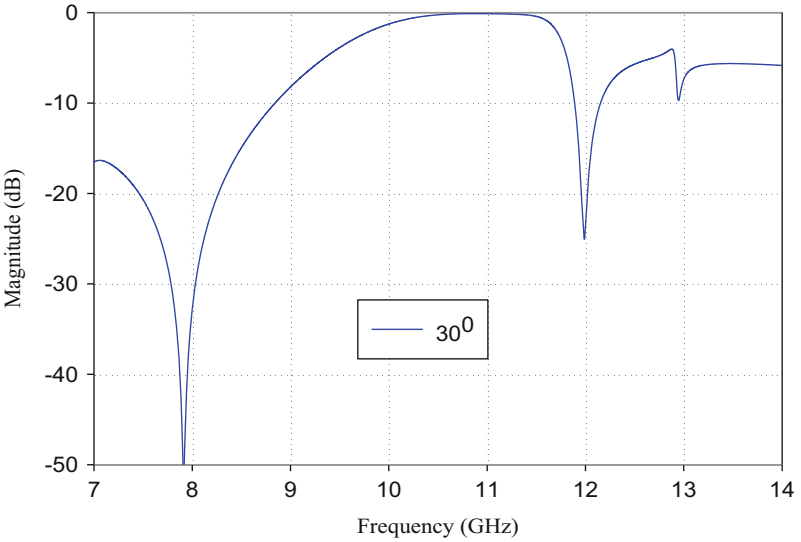


Fig. 24 Transmission characteristics of proposed hybrid-shorted MTM-FSS unit cell at 30° for TM polarization

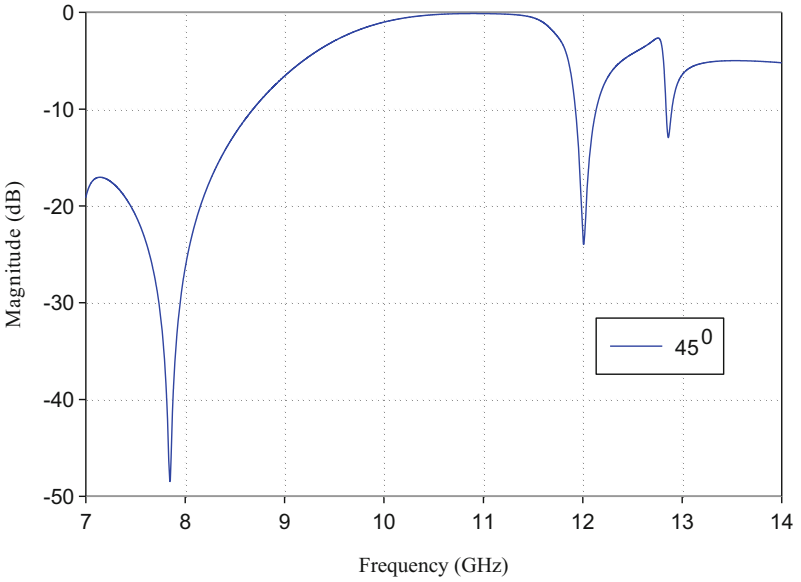


Fig. 25 Transmission characteristics of proposed hybrid-shorted MTM-FSS unit cell at 45° for TM polarization

Table 3 Design parameters for optimized proposed MTM-FSS cell

Design parameters	Optimized dimensions (mm)
Periodicity (*dx*); L	9.8
Periodicity (*dy*); W	9.8
Width of the conductors; w	1.0
Spacing between two-unit cells	1.3
Substrate height; H	1.8
Length of extended arm from the center; a	3.75
Side of square-shaped patch at the edges; c	1.5

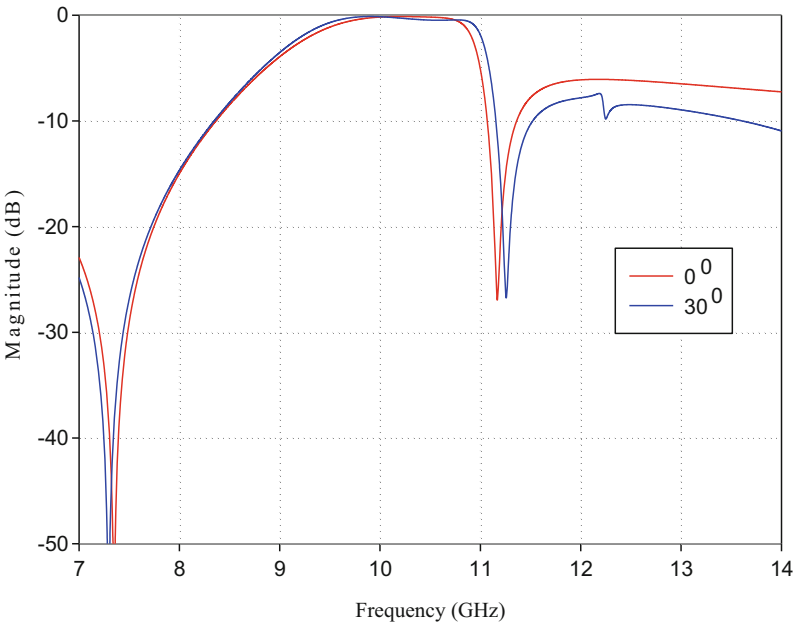


Fig. 26 Power transmission for proposed optimized MTM-FSS structure at 0^o and 30^o using TLTMM method

The optimized MTM-FSS unit cell exhibits a flat transmission band (~ -0.5 dB loss) from 9.74 GHz - 10.745 GHz with a bandwidth of l GHz at central frequency 10.2 GHz as shown in Fig. 26 for normal incidence and 30^o wave incidence for TE polarization. The substrate between the top and bottom pair of conductors consists of Arlon ($\varepsilon_r = 3.42$, $tan\delta = 0.0035$). The EM performance of proposed MTM-FSS structure is compared with the standard double square loop FSS (Luo et al. 2005) structure optimized at central frequency (10.2 GHz) same as the shorted proposed MTM-FSS using identical dielectric constant shown in Fig. 27.

Design parameter considerations for the optimized standard double square loop at the desired central frequency are listed in Table 4. The transmission performance of proposed MTM-FSS-shorted cell is found to be superior as compared to the standard

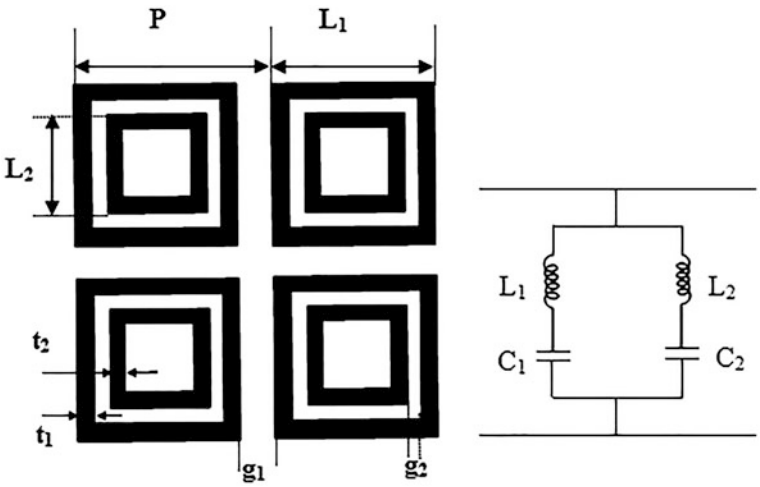


Fig. 27 Schematic of optimized double square loop FSS

Table 4 Design considerations for optimized double quare loop FSS

Design parameters	Optimized dimensions (mm)
Periodicity (p)	9.7
Length of outer square-loop (d_1)	8.7
Length of inner square-loop	6.5
Width of outer square-loop (t_1)	0.8
Width of inner square-loop (t_2)	0.8
Gap between the unit cells (g_1)	1.0
Spacing between inner and outer square-loop (g_2)	1.1
Height of dielectric slab	1.8

DSL-FSS as shown in Fig. 28. In comparison with the double square loop with 750 MHz bandwidth, the proposed shorted MTM structure provides a bandwidth of l GHz in the transmission band at 10.2 GHz. Besides, the sharp roll off behavior outside the transmission band aids to reduce the out-of-band RCS to lower value as compared to the double square loop FSS as RCS reduction of radome is one of the key objectives of this work. In addition to that, the proposed shorted MTM structure gives an extra degree of freedom in terms of optimization with various geometrical parameters such as arm widths, gap between adjacent unit cells, and height of the substrate. Such optimized geometry provides a strong electric and magnetic frequency at particular frequency that indicates regions of negative permittivity and permeability, respectively. Thus the proposed MTM structure exhibits a transmission band between electric and magnetic resonances as well. The proposed MTM-FSS structure array has band spacings, $f_t/f_{r1} = 1.39$ and $f_{r2}/f_t = 1.10$, while the standard double square loop array has band spacings, $f_t/f_{r1} = 1.49$ and $f_{r2}/f_t = 1.44$.

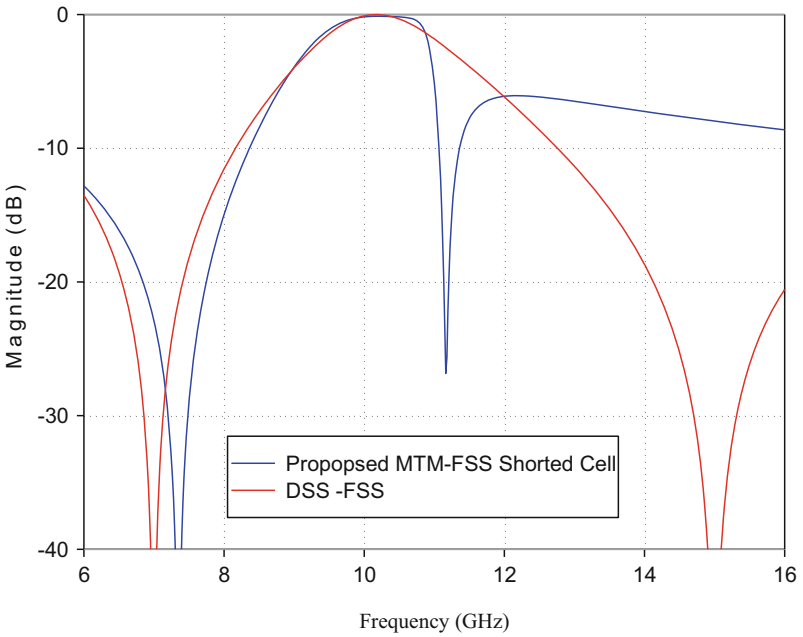


Fig. 28 Power transmission for proposed MTM-FSS cell and standard double square loop FSS cell at normal incidence

To summarize, Subsection "Tightly Coupled Conductors-Based MTM-FSS Unit Cell" illustrates the design of novel LH MTM unit cell consisting of tightly coupled inverted conductors on both sides of dielectric substrate exhibiting narrow LH transmission band in the X-band. Further, taking into account the dependency of electric and magnetic resonance on structural parameters of the proposed cell due to symmetric and asymmetric current flowing on top and bottom conductors, novel MTM-based FSS structure with shorted strips at the edges has been implemented to achieve a flat-top transmission band of 1 GHz BW in the X-band with central frequency 10.2 GHz between the regions of negative permittivity and negative permeability. Finally, the superiority of the proposed polarization independent MTM-FSS with shorted strips at edges is compared with standard optimized double square loop FSS with respect to overall bandwidth and sharp roll-off in the out-of-band transmission characteristics (subsection "Comparison Between Double Square Loop FSS and Proposed Shorted MTM-FSS Structure").

Patch-Type Hybrid Double Square Metamaterial-FSS Unit Cell

A modified double square loop, with inner loop consisting of four T-shaped conductors as shown in Fig. 29, is printed on a 1-mm-thick substrate with dielectric constant, 3.28, and loss tangent, 0.001. The dimensions of the proposed unit cell are

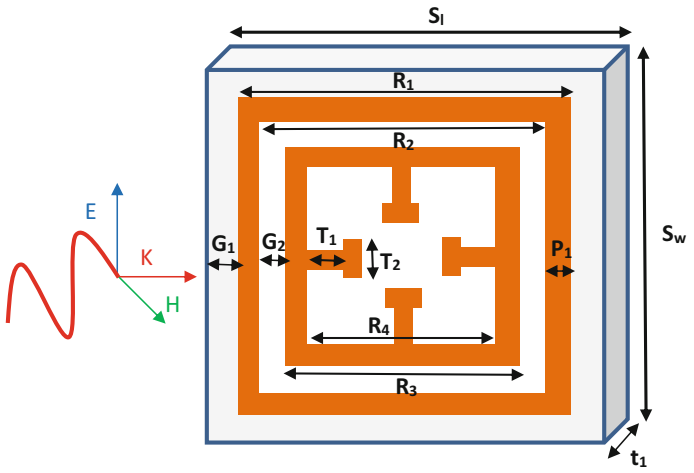


Fig. 29 Schematic of hybrid patch-type metamaterial-FSS unit cell

Table 5 Design parameters for hybrid patch MTM-FSS unit cell

Design parameters	Dimension (mm)
SI	9.00
Sw	9.00
Pl	0.50
Gl	1.00
G2	0.50
Tl	0.50
T2	1.00
Rl	8.00
R2	7.00
R3	4.96
R4	3.96

optimized to resonate at X-band. The optimized design parameters of hybrid MTM-FSS are given in Table 5.

As EM plane wave illuminates the metamaterial surface, the field induces across the metamaterial structure as shown in Fig. 30. The surface current density of the MTM-FSS at resonance frequency of 10.025 GHz is analyzed where TE plane wave is impinging normal to the surface (incident polarization angle is 0°). At resonance frequency, two electric dipoles are generated on the MTM-FSS structure (one is due to the outer ring, and other is due to the inner ring) with corresponding oscillating current which is shown in Fig. 30 with blue arrow lines. Both dipoles' oscillating currents are flowing in opposite direction. The outer metal ring is observed to have low-surface current density compared to the inner resonating metal ring. These excited dipoles partially reflect and transmit the incident electromagnetic plane wave.

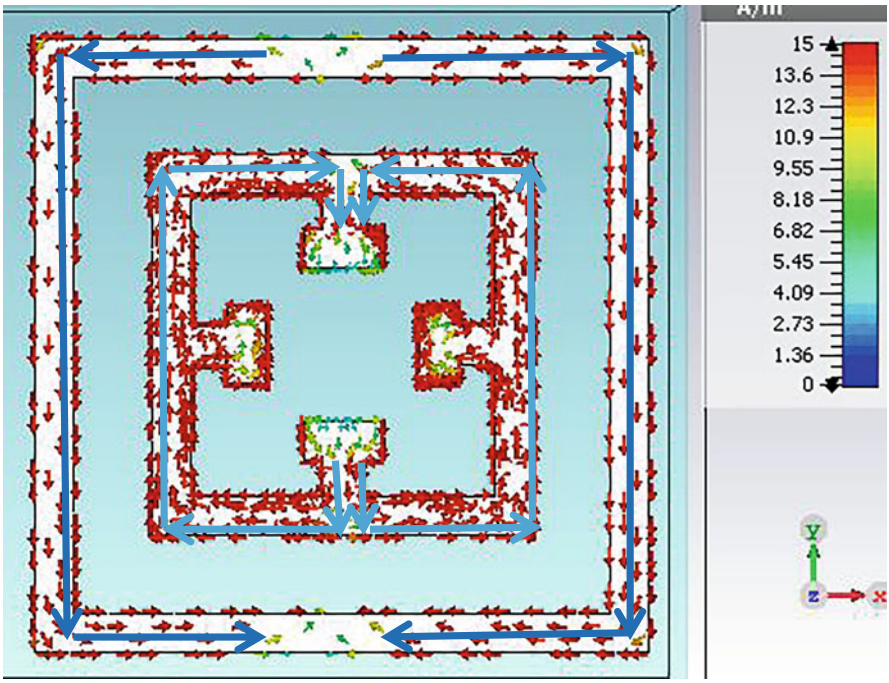


Fig. 30 Schematic of surface current density on MTM-FSS at 10.025 GHz at normal incident for TE polarization

Figure 31 illustrates the surface current density of the MTM-FSS at resonance frequency of 10.025 GHz for TM-polarized wave impinging normal to the surface (incident polarization angle is 0°). Here, two electric dipoles are created on the surface of MTM-FSS due to incident TM wave that are 90° out of phase with respect to the dipoles created in case of TE polarized.

Equivalent Circuit Model of Hybrid MTM-FSS Structure

Figure 32 illustrates the equivalent electrical circuit model of proposed MTM-FSS structure. As EM plane wave illuminates the FSS structure, an electric current is induced and starts flowing through the rectangular metallic bars in the MTM-FSS structure. It is observed that the self-inductance produced here is due to the current induced in the metallic bar, which is approximately proportional to the length of the bar. The equivalent circuit model of the proposed structure can be represented with two series combination of LC-resonant circuits in parallel as shown in Fig. 32. The inductance and capacitance provided by the outer loop is represented by L_1 and C_1 in the equivalent circuit model, respectively. The inner square loop with T-shaped conductors can be characterized by two series inductors, L_2 and L_2' and two parallel capacitors, C_2 and C_2'.

The expressions (Sangeetha et al. 2014) given below are used to estimate the numerical values of L_1, C_1, L_2, and C_2.

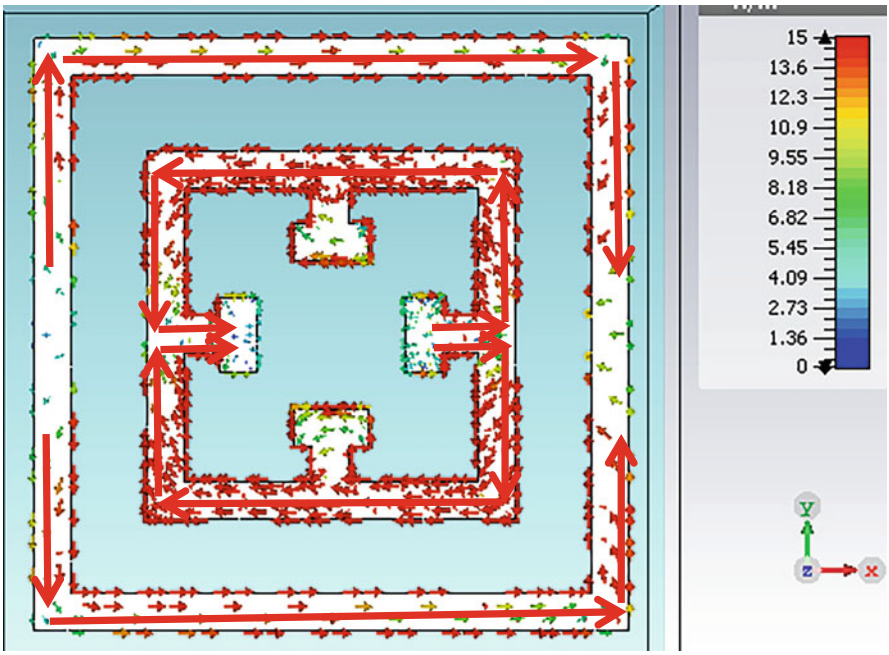


Fig. 31 Schematic of surface current density on MTM-FSS at 10.025 GHz at normal incidence global correction for TM polarization

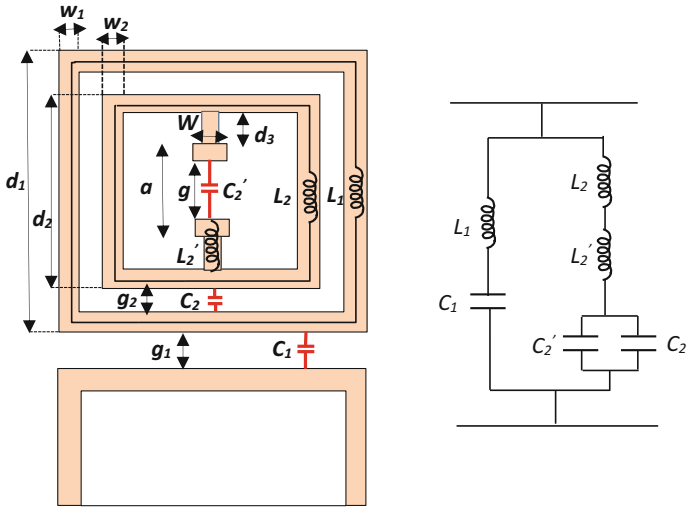


Fig. 32 Equivalent circuit model of MTM-FSS at 10.025 GHz for normal incident EM plane wave

$$X_1 = \omega L_1 = F(p, w_1, \lambda) \times \left(\frac{d_1}{p}\right) \tag{20}$$

$$X_2 = \omega L_2 = F(p, w_2, \lambda) \times \left(\frac{d_2}{p}\right) \tag{21}$$

$$X_3 = \omega L_2' = F(p, w_3, \lambda) \times \left(\frac{d_3}{p}\right) \tag{22}$$

where X_1, X_2, and X_3 are the inductive reactance of outer square-loop, inner square-loop, and inner T-type metal strip, respectively. p is the periodicity of the hybrid-element unit cell. w_1, w_2, and w_3 are the width of inductive grids of square-loop, respectively, where the reactance is reduced by a factor of $\frac{d_1}{p}$, $\frac{d_2}{p}$, and $\frac{d_3}{p}$, respectively, owing to the conductor not being continuous. F is the function of periodicity, width, and wavelength, which can be computed for TE and TM polarizations (Sangeetha et al. 2014).

Further, the capacitive susceptance of outer and inner square loop can be determined by Eqs. (23) and (24), respectively, as

$$B_1 = \omega C_1 = 0.75 \times B_{1p} \times \left(\frac{d_1}{p}\right) \tag{23}$$

$$B_2 = \omega C_2 = \frac{B_{1p} \times B_{2p}}{B_{1p} + B_{2p}} \times \left(\frac{d_2}{p}\right) \tag{24}$$

where B_{1p} and B_{2p} are given as,

$$B_{1p} = 4.0 \times \varepsilon_{eff} \times F(p,\ g_1, \lambda) \tag{25}$$

$$B_{2p} = 4.0 \times \varepsilon_{eff} \times F(p,\ g_2, \lambda) \tag{26}$$

and the effective dielectric constant, $\mathcal{E}_{eff}$, is given by

$$\mathcal{E}_{eff} = \varepsilon_{r_h} + (\varepsilon_{r_h} - 1)\left[\frac{-1}{\exp^N(x)}\right] \tag{27}$$

where $\varepsilon_{r_h} = \frac{\varepsilon_r + 1}{2}$ and $x = \frac{10h}{p}$.

The numerical value of exponential factor "N" is 1.3 for ring-type structures and 1.8 for the cross-type of structure. The thickness of the substrate is represented by h, and periodicity of the proposed MTM-FSS unit cell is p.

The gap capacitance C_2' is calculated as (Hosseini and Hakkak 2008),

$$C_2' = \frac{2W}{\pi} \varepsilon_0 \varepsilon_{reff} \cosh^{-1}\left(\frac{a}{g}\right) \tag{28}$$

where

$$\varepsilon_{reff} = \frac{\varepsilon_r + 1}{2} + \frac{\frac{\varepsilon_r - 1}{2}}{\sqrt{1 + 10\left(\frac{h}{W}\right)}} \tag{29}$$

For TE and TM mode of incident wave, FSS structure provides different inductive and capacitance reactances. So, it is necessary to determine the reactance for both TE and TM polarizations independently.

Now from Fig. 32, the admittance of hybrid-element MTM-FSS is determined as

$$Y = j\left[\left(\frac{B_1}{1 - X_1 B_1}\right) + \left(\frac{(B_2 + B_3)}{1 - (X_2 + X_3))(B_2 + B_3)}\right)\right] \tag{30}$$

The equivalent impedance of the hybrid-FSS unit cell can be calculated as:

$$Z = \frac{1}{Y} \tag{31}$$

The Eq. (31) is used to estimate the transmission and reflection characteristics of proposed MTM-FSS structure. In order to determine resonant frequency of proposed MTM-FSS structure, the equivalent circuit method is used since it can be represented by lumped LC resonant circuit as shown in Fig. 32. The resonance frequencies of outer and inner square-loop elements is calculated as,

$$f_{r1} = \frac{1}{2\pi\sqrt{L_1 C_1}} \tag{32}$$

$$f_{r2} = \frac{1}{2\pi\sqrt{\left(L_2 + L_2'\right)\left(C_2 + C_2'\right)}} \tag{33}$$

To examine the efficacy of the equivalent circuit approach, the transmission characteristic of hybrid-element MTM-FSS is compared with the full-wave analysis method as shown in Fig. 33. Using the ECM, calculated values of f_{r1} and f_{r2} are 7.26 GHz and 13.74 GHz, respectively. From full-wave method, the resonance frequencies f_{r1} and f_{r2} are observed to be 7.13 GHz and 14.20 GHz, respectively. lt is observed that the two resonance frequencies estimated based on full- wave method are very close to the ECM-based results. Thus, a good agreement is obtained between computed and simulated results.

EM Performance Analysis of Hybrid MTM-FSS Unit-Cell

The proposed hybrid MTM-FSS unit cell as shown in Fig. 32 seems to be symmetrical along X- and Y-directions. Thus the proposed structure is polarization independent with E-field directed perpendicular to the structure. The effective parameters are extracted from absolute value and phase of transmission, S_{21} and reflection, S_{11} coefficient. Figures 34 and 35 show the extracted metamaterial behavior,

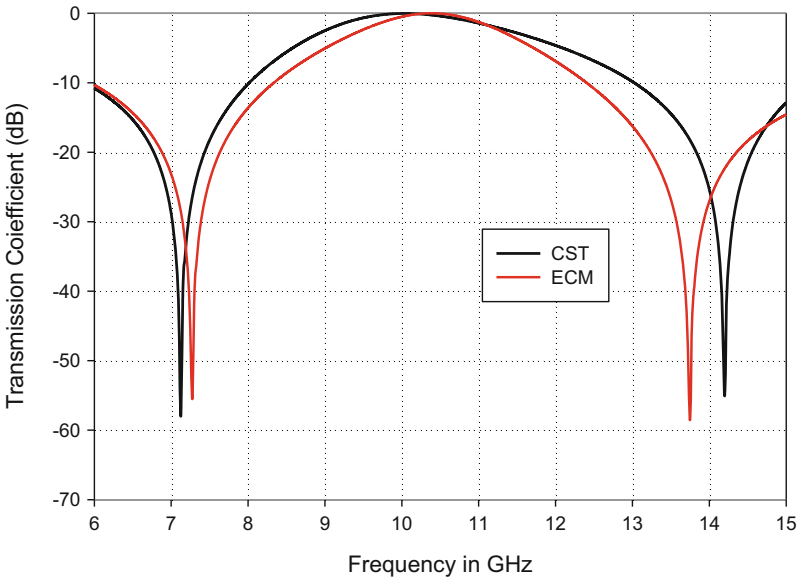


Fig. 33 Transmission coefficient of metamaterial structure using full-wave method (CST microwave studio) and ECM

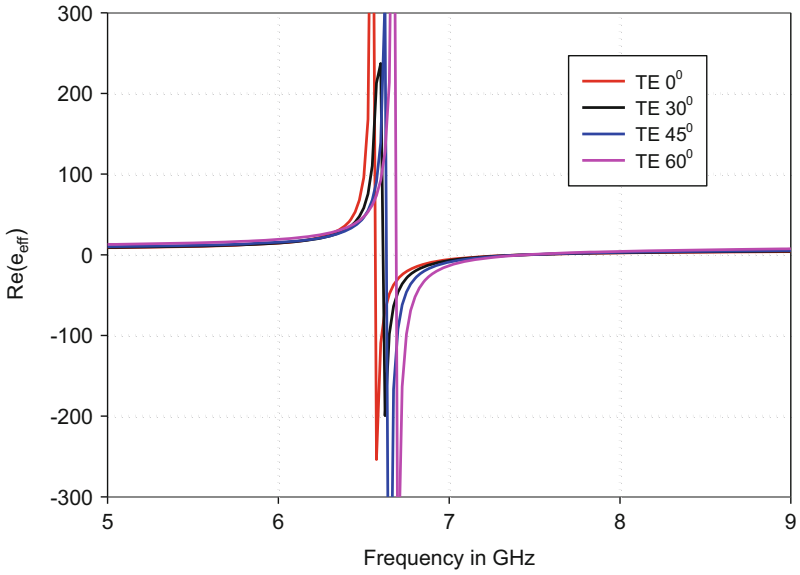


Fig. 34 Extracted metamaterial characteristics of MTM-FSS unit cell at different angles of incidence (0°, 30°, 45°, and 60°)

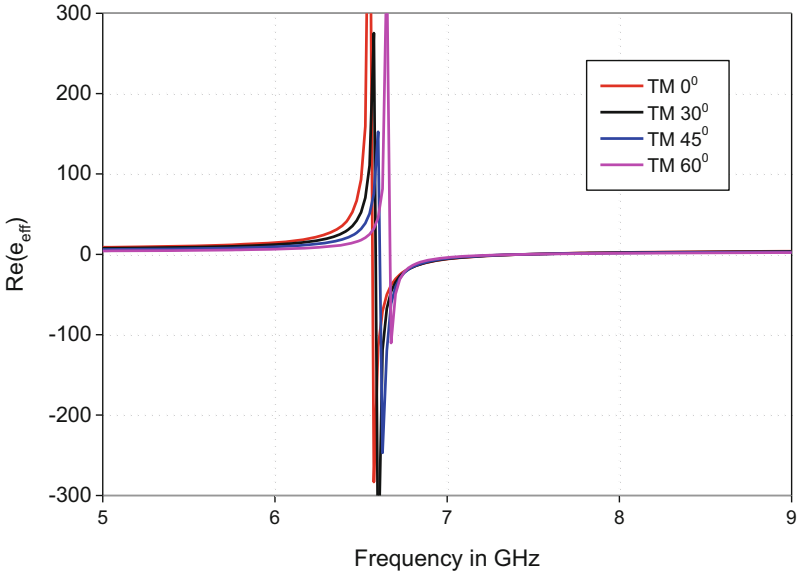


Fig. 35 Extracted metamaterial characteristics for proposed hybrid MTM-FSS unit cell at different angles of incidence (0°, 30°, 45°, and 60°)

respectively, for TE and TM mode of polarization at various angles of incidence. The proposed structure exhibits ε-negative behavior around 6.7 GHz at different TE and TM polarizations below the electric plasma frequency. Further, the transmission characteristic of proposed hybrid MTM-FSS is studied at oblique angles of incidence for TE and TM polarizations. The transmission line transfer matrix method (TLTMM) is used to perform EM analysis of hybrid MTM-FSS structure at different angles of incidence. Figures 36 and 37 depict the power transmission characteristics of the proposed unit cell for TE and TM polarizations. Here, it is observed that the resonance frequencies of proposed structure do not vary *w.r.t.* the incidence angle for both polarizations.

Metamaterial-FSS Structures Loaded with Active Devices

Metamaterial is composed of densely packed composite materials or metal resonators. These artificial composites exhibit a predominant characteristic where their effective permittivity and permeability can be arbitrarily tuned to a positive or negative value. These are extensively used in the design of antennas, filters, FSS, and absorbers due to their capability to influence electromagnetic waves. The implementation of switching and active elements into the metamaterial including diodes and transistors in the recent years enables it to have a tunable refractive index (Jiang et al. 2012), and such materials are called "active meta-material."

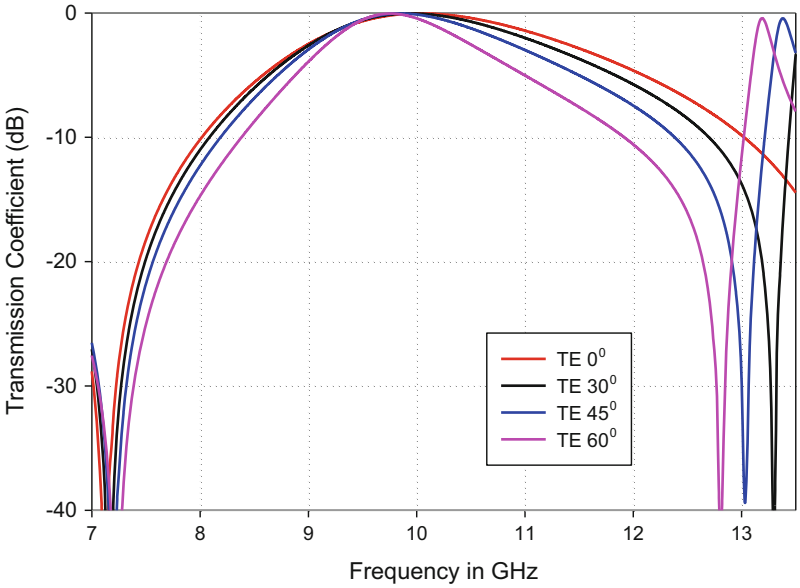


Fig. 36 Transmission characteristics of hybrid MTM-FSS at different angles of incidence (0°, 30°, 45°, and 60°) for TE polarization

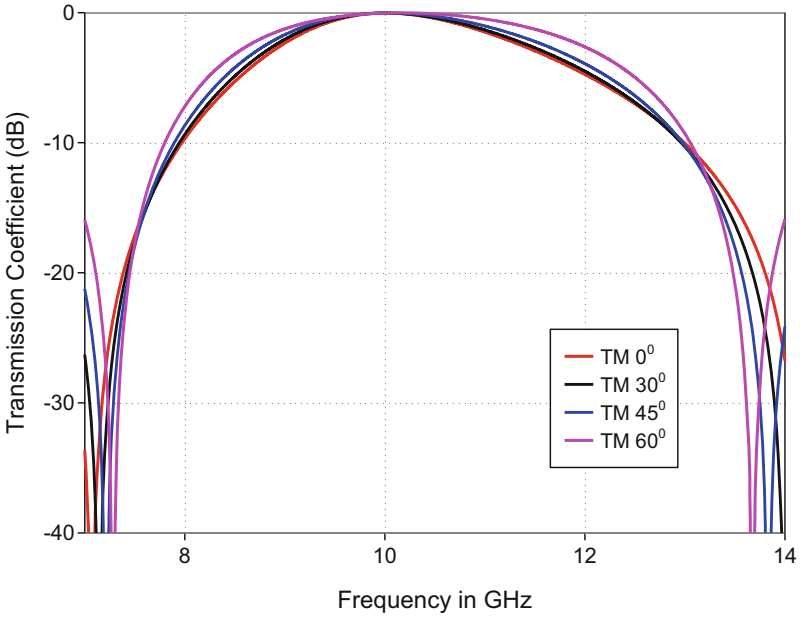


Fig. 37 Transmission characteristics of hybrid MTM-FSS at various angles of incidence (0°, 30°, 45°, and 60°) for TM polarization

The standard approaches to modify the resonant frequency of a metamaterial include varying the geometrical parameters of the metallic elements such as strip width, spacing between elements, periodicity, split width, etc. or changing the material properties (thickness and dielectric constant). However, it is impossible to alter the printed metallic structure and substrate properties once it is fabricated unless some refined practices are used for real-time tuning (Jeremiah et al. 2004). Varactor diodes, due to their ease of integration into various kinds of materials, are preferred as a tuning technique. For real-time tuning of resonant frequencies of SRRs or other kinds of metamaterials, varactor diodes are used due to their inherent voltage-controlled capacitance characteristics (Fig. 38).

The electromagnetic properties of conventional metamaterials are performed at a specified fixed frequency. To operate the designed MTM structure at other frequency, it is required to redesign the structure to hold the same electromagnetic characteristics at that frequency, which leads to a less efficient and expensive design. Tunable metamaterials would be the better choice to resolve this problem. Tunable metamaterial shows the real-time control of its EM characteristics by the loading of active tunable semiconductor devices, such as RF MEMS, RF PIN diodes, ferrite material, varactor diodes, etc. Among these various methods, varactor diodes are used as its status can be changed in real time and the diode current is negligible in reverse bias (Niel et al. 2016). The varactor diode is incorporated between the elements in a unit cell to achieve real-time tuning of EM characteristics of proposed metamaterial structure (Fig. 39). Here, the metamaterial-FSS structure is analyzed with and without loading the active device and is designed such that the proposed structure can be tuned over different frequencies in presence of the diode, and it exhibits similar response at oblique angles of incidence for both TE and TM polarizations.

The varactor diode is a P-N junction semiconductor diode, which changes its junction capacitance with applied reverse bias voltage. This property of dynamic

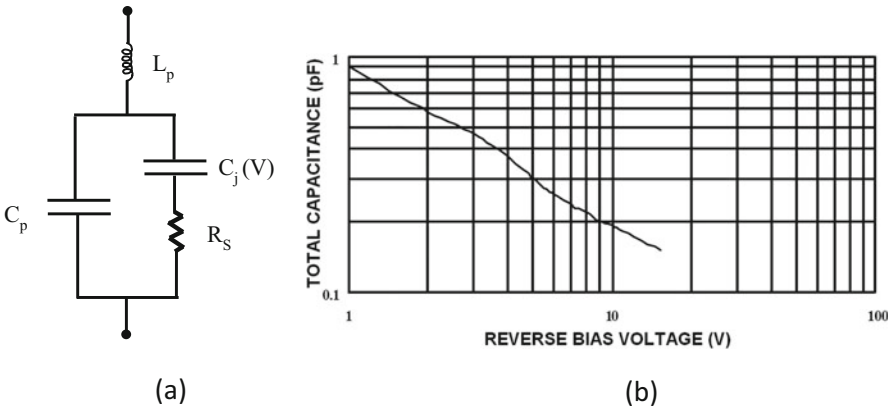


Fig. 38 (**a**) Equivalent electrical circuit model of a varactor diode, (**b**) Total capacitance of varactor diode as a function of applied reverse bias voltage

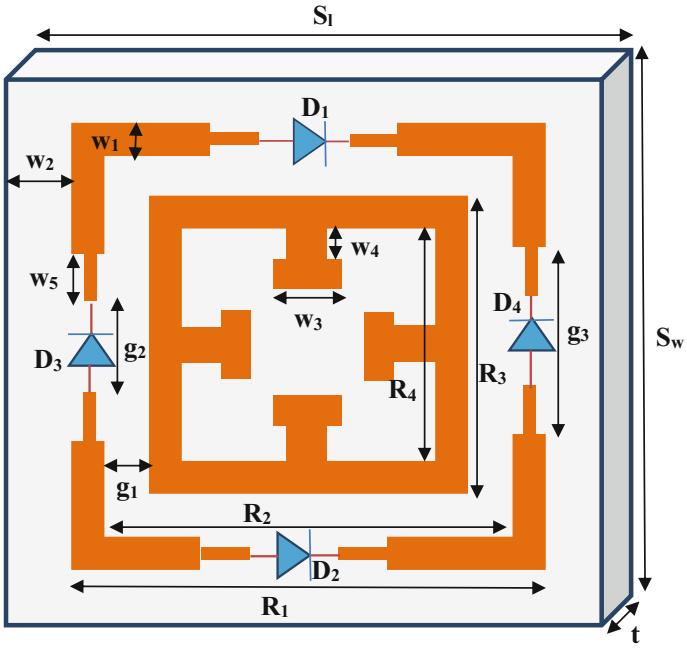


Fig. 39 Schematic of MTM-FSS unit cell loaded with varactor diodes

change in capacitance *w.r.t.* applied voltage can be used to achieve a change in the frequency or the phase of an electrical circuit, or both. An equivalent circuit model of varactor diode is shown in Fig. 38 (a), where C_p is the capacitance due to the diode package, C_j is the diode junction capacitance, L_p is the inductance due to the diode package, and Rs is the series resistance.

The MTM-FSS unit cell is shown in Fig. 39 which is loaded with four varactor diodes across four gaps in the outer metal ring. Here, the model of used varactor diode is MACOM's MA46H120-ODS1203 (http://www.mouser.com/access/?pn=MA46H120). Its junction capacitance, C_j, varies from 1.14 pF to 0.14 pF as the reversed bias voltage is varied from 0 V to 15 V. Varactor diode under unbiased condition has higher junction capacitance C = 1.14 pF at 0 volts. As the reverse voltage is increased, its capacitance decreases gradually as shown in Fig. 38 (b). The optimized design parameters of varactor diodes-loaded MTM-FSS unit cell are described in Table 6.

EM Performance Analysis of Varactor-Loaded MTM-FSS Structure

The surface current density of varactor-loaded MTM-FSS at resonance frequency of 10.025 GHz is analyzed as shown in Fig. 40. The incident plane wave is considered to be TE polarized. Two electric dipoles at a resonance frequency are created on the

Table 6 Optimized design parameters for varactor-loaded MTM-FSS unit cell

Design parameters	Dimension (mm)
SI	8.00
Sw	8.00
Wl	0.35
W2	0.80
W3	1.20
W4	0.57
W5	0.55
Rl	5.70
R2	6.40
R3	4.54
R4	3.84
gl	0.58
g2	0.70
g3	1.80

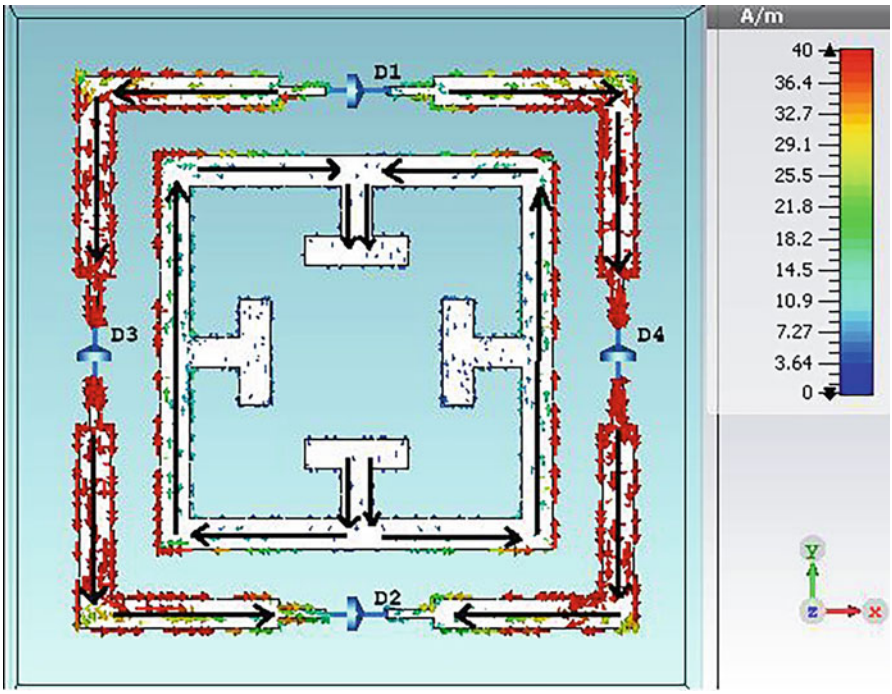


Fig. 40 Schematic of surface current density on varactor-loaded MTM-FSS at 10.025 GHz illuminated with TE-polarized wave

proposed structure (one is due to the outer ring, and other is due to the inner ring) with corresponding oscillating current which is shown in Fig. 41 with black arrow lines. The diodes D3 and D4 at outer ring are excited and add up the capacitive

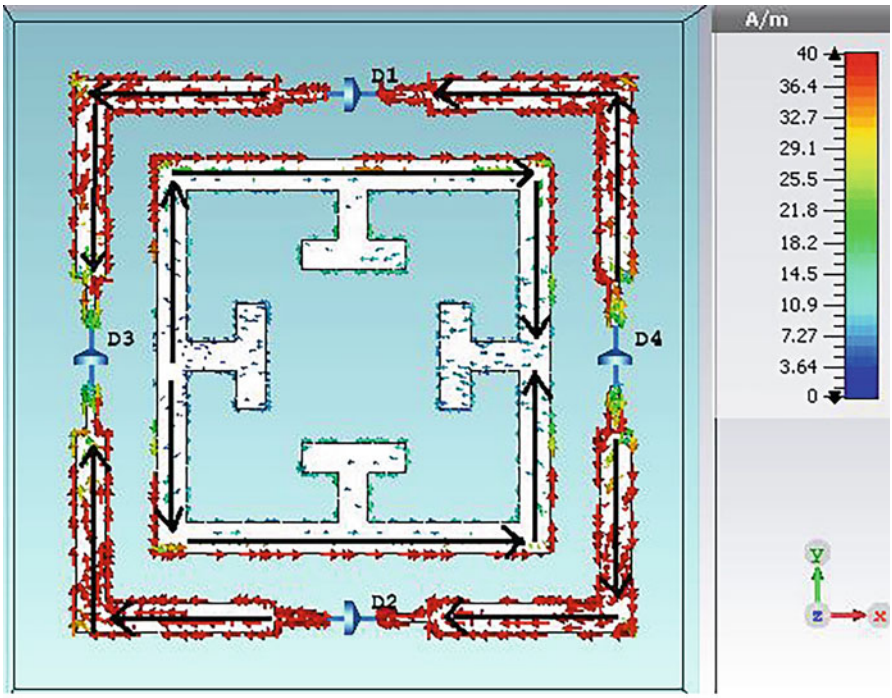


Fig. 41 Schematic of surface current density on varactor-loaded MTM-FSS at 10.025 GHz illuminated with TM-polarized wave

effect. Because of this capacitance, the outer ring has more dominant conduction current as compared to the inner ring.

Figure 41 illustrates the surface current density of the proposed varactor-loaded unit cell at resonance frequency of 10.025 GHz, where the incident plane wave is TM polarized. Therefore, diodes Dl and D2 at outer ring are excited and add the capacitive effect. Here, two electric dipoles are generated on the surface of structure due to incident TM polarized wave, which are 90^{o} out of phase with respect to the dipoles created in case of TE polarized.

Further, the single band frequency-tuning capability of the proposed varactor-loaded MTM-FSS structure is studied. Figures 42 and 43 illustrate the extracted metamaterial properties (effective permittivity) of the proposed design at various applied reverse voltages to the varactor diodes for TE and TM polarization, respectively, under normal incidence. At normal incident plane wave, it is observed that the proposed unit cell with varactor diodes shows metamaterial ENG properties, i.e., ε-negative. In conclusion, the resonant frequency of the metamaterial structure can be tuned to other frequencies by changing the varactor diode capacitance.

To show the tuning capability of proposed varactor-loaded MTM-FSS structure for both TE and TM polarization, the junction capacitance of the varactor diode is varied between 1.14 pF − 0.14 pF *w.r.t;* the applied reverse bias voltage ranges from

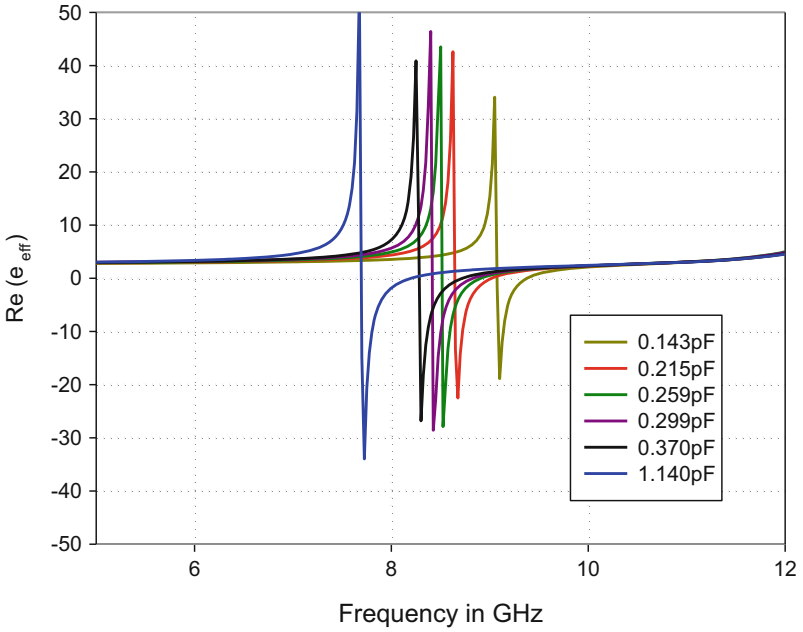


Fig. 42 Effective permittivity of varactor-loaded MTM-FSS unit cell at different bias voltages for TE polarization

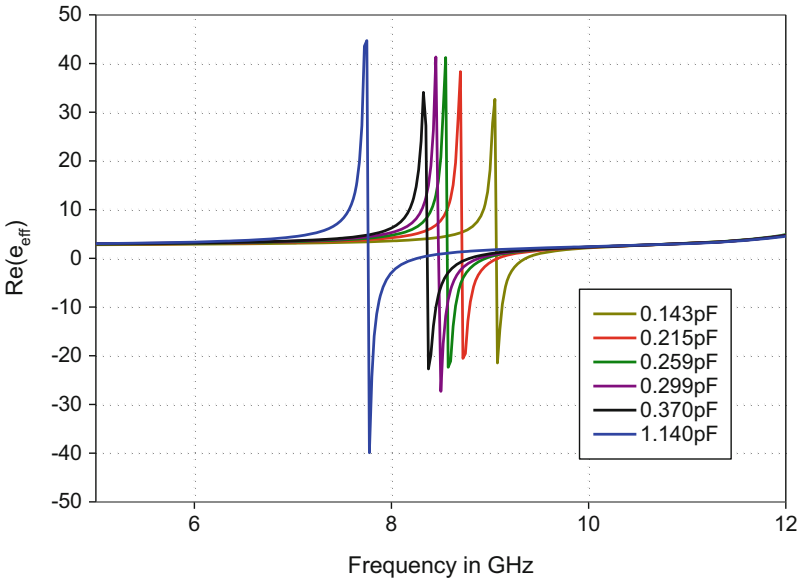


Fig. 43 Effective permittivity of varactor-loaded hybrid MTM-FSS unit cell at different bias voltages for TM polarization

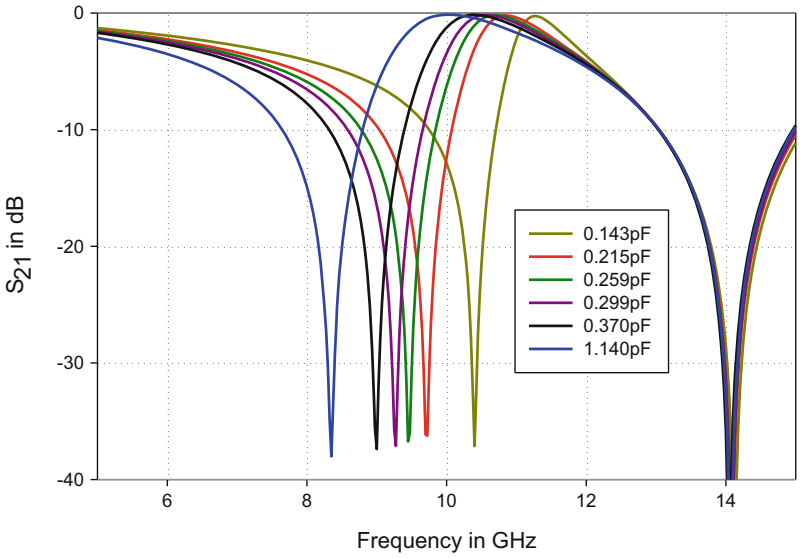


Fig. 44 Transmission characteristics of varactor-loaded MTM-FSS unit cell at different bias voltages for TE polarization

0 to 15 volts. It is observed that the resonance frequency of proposed structure is tuned from 10.025 GHz to 11.425 GHz as shown in Figs. 44 and 45.

It is also observed that the first reflection resonance shifts toward the higher frequency side as the diode reverse bias voltage increases from 0 to 15 volts, while second reflection resonance is observed to be stable. Hence the transmission bandwidth of proposed varactor-loaded MTM- FSS structure decreases with the increase in diode reverse bias voltage. The resonance frequency of the MTM-FSS structure is controlled in real time by applying reverse bias voltage of the varactor diodes. The simulation results show the configurability of the proposed structure. Thus, the proposed MTM-FSS structure can be used for realizing reconfigurable components such as antennas and radomes in the GHz regime.

EM Performance Analysis of Airborne Radome Embedded with Active MTM-FSS Structure (Varactor Diode)

Reconfigurable metamaterials find potential applications in various domains including optics, microelectronics, and imaging where flexibility, miniaturization, and functionality are primary concerns. The phenomenon of reconfiguration has been achieved by various approaches. One common method is varying the geometrical design parameters of the metallic part or varying the dielectric properties of the substrate. It may not be possible to change both geometrical parameters and material properties for reconfiguration and tuning once the fabrication is completed except if

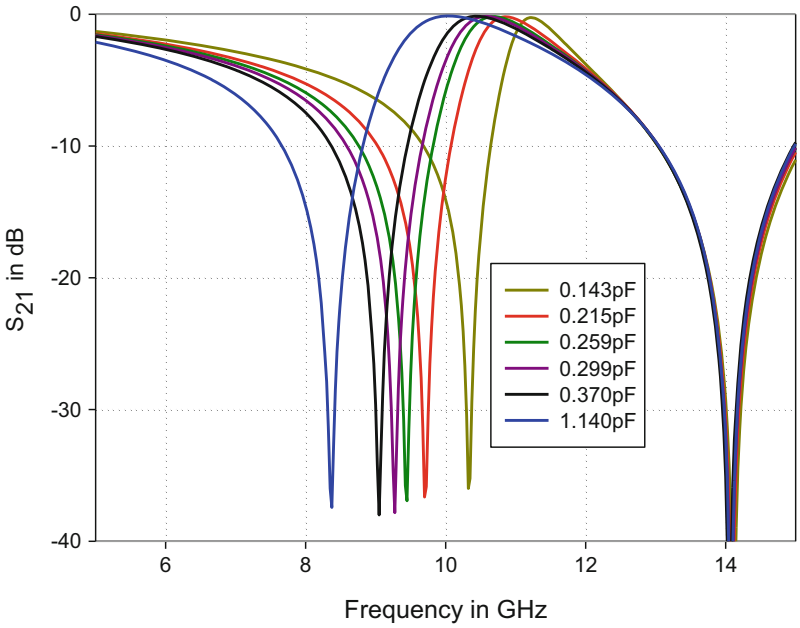


Fig. 45 Transmission characteristics of varactor-loaded MTM-FSS unit cell at different bias voltages for TM polarization

some sophisticated methods are used for real-time tuning. One of the most popular techniques for real-time tuning after fabrication is the simple integration of varactor diode into SRRs and various kinds of metamaterial resonant unit cells.

Real-time tuning is predominant when the operating frequency of radars needs to be changed if jammers are present for the current operating band, especially during electronic warfare. In such circumstances, metamaterial-based radome aids the tuning to desired frequency which is unattainable with standard dielectric materials. On this account, varactor diodes-loaded MTM- FSS structure has been embedded in the radome wall to analyze its tuning for airborne radome applications. Analysis has been made based on planar-radome wall configuration with active tunable meta-material. Further, the optimized planar configuration has been employed in the design of tangent ogive radome for airborne platforms. EM performance analysis has been carried out using 3D ray-based Geometrical Optics method in combination with aperture integration approach to compute various critical radome parameters such as copol power transmission characteristics, cross-pol power transmission characteristics, radiation pattern distortion, and boresight error w.r.t. to beam steering angle of *active electronically scanned array* (AESA).

Monolithic Radome with Hybrid MTM-FSS Structure

The proposed planar radome wall consists of a hybrid MTM-FSS layer embedded in the mid plane of monolithic radome wall of Quartz epoxy material ($\varepsilon_r = 3.28$,

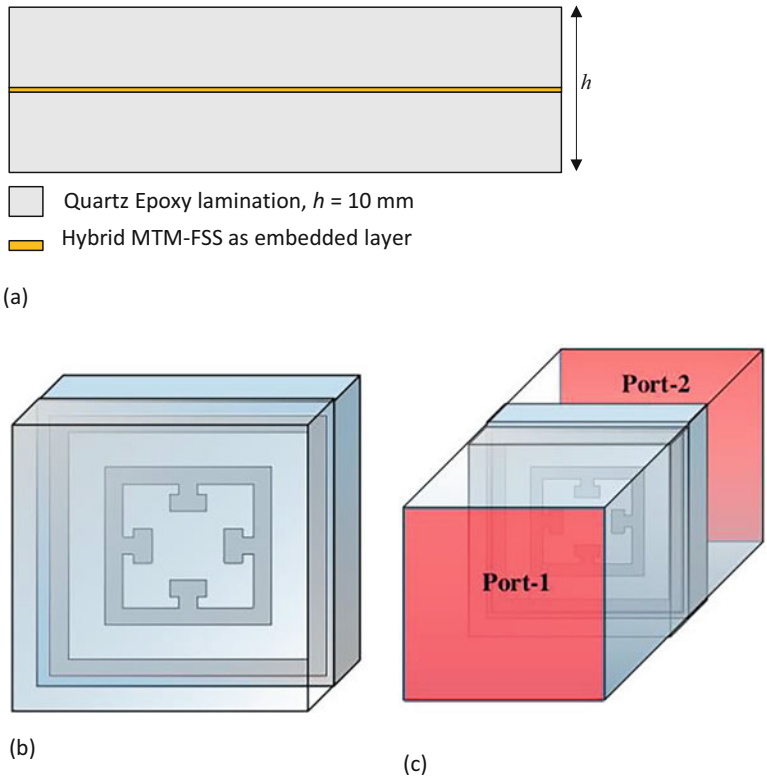
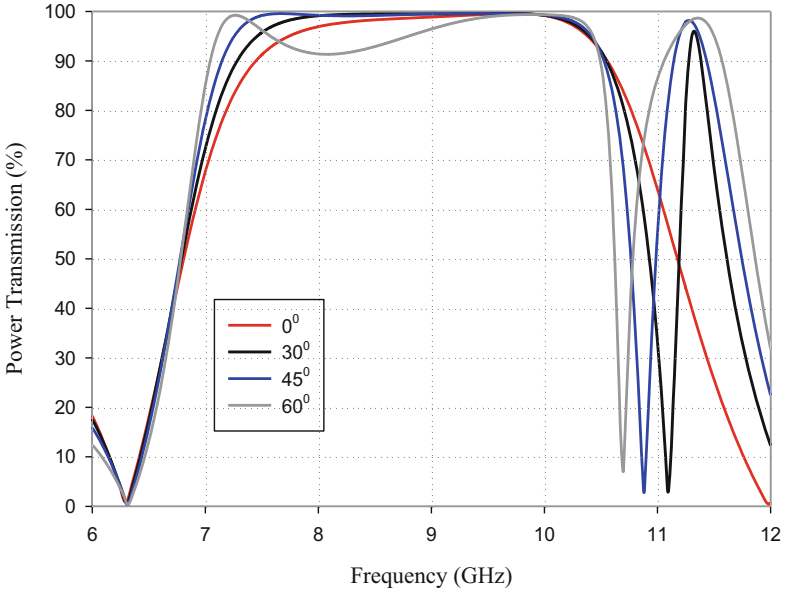


Fig. 46 Schematic of metamaterial FSS-embedded monolithic radome wall configuration; (**a**) side view; (**b**) 3D view; and (**c**) simulation setup of MTM-embedded planar radome with wave ports

$tan\delta = 0.001$) as shown in Fig. 46 (a, b). for EM simulation, the optimized dimension of unit cell of proposed radome is $9 \times 9 \times 10$ mm. The EM characteristics of the proposed configuration (Fig. 46c) has been analyzed with a low-loss material in terms of transmission, reflection, bandwidth, and sharp roll-off characteristics.

The transmission characteristics of the abovementioned configuration at various angles of incidence for both TE and TM polarizations, respectively, are shown in Fig. 47. For TE polarization, the quartz-epoxy-based monolithic configuration shows excellent power transmission performance. It provides 3 GHz (7.45–10.54 GHz) bandwidth w.r.t. 90% power transmission (0.4 dB loss), but there is an attenuation in transmission efficiency at the 8.2 GHz frequency for higherincidence angle (i.e., 60°) as evident from Fig. 47a.

For TM polarization (Fig. 47b), the proposed MTM-embedded planar radome also exhibits excellent band pass characteristics with a bandwidth of 3 GHz (7.46–10.56 GHz) w. r. t. 90% power transmission(−0.4 dB loss) at normal incidence. However, bandwidth is observed to reduce for oblique incidence (45° and 60°) due to the effective capacitance between the inter elements of the proposed configuration. It is further revealed that the proposed MTM-FSS structure exhibits



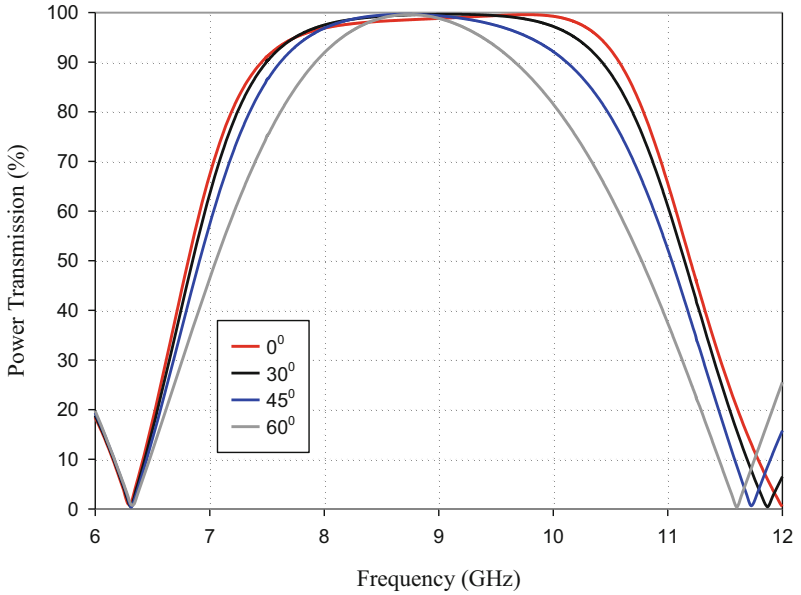


Fig. 47 Power transmission response of proposed radome configuration embedded with hybrid MTM-FSS structure at oblique angle of incidence for: (**a**) TE polarization (**b**) TM polarization

more than 1 GHz bandwidth w.r.t. 90% transmission at oblique incidence for TE and TM polarization. Thus, the planar radome wall configuration designed by embedding proposed MTM structure can be used for airborne radome applications.

Planar Monolithic Radome Embedded with Active MTM-FSS Structure (Varactor Diode)

To gain a physical insight into the real-time tuning of EM characteristics of radome wall configuration, varactor diode-loaded metamaterial FSS embedded between two quartz epoxy layers ($\varepsilon_r = 3.28$, $tan\delta = 0.001$) of unit cell as shown in Fig. 48 is analyzed. The optimized total thickness and periodicity of the active tuneable unit cell is 10 mm and 8 mm, respectively. The outer loop in the FSS structure resonates at lower frequency, and inner loop with T-shaped arms resonates at comparatively higher frequency, which results band pass characteristic over the desired frequency range.

Here, as diode is incorporated at the outer ring of the structure, lower resonance frequency of the structure can be tuned by changing the diode capacitance. The EM analysis of the proposed unit cell loaded with varactor diodes has been carried out using full-wave method-based CST Microwave Studio. Figures 49 and 50 illustrate the frequency response of power transmission and its active tuning at various incidence angles for both TE and TM polarization corresponding to different reverse

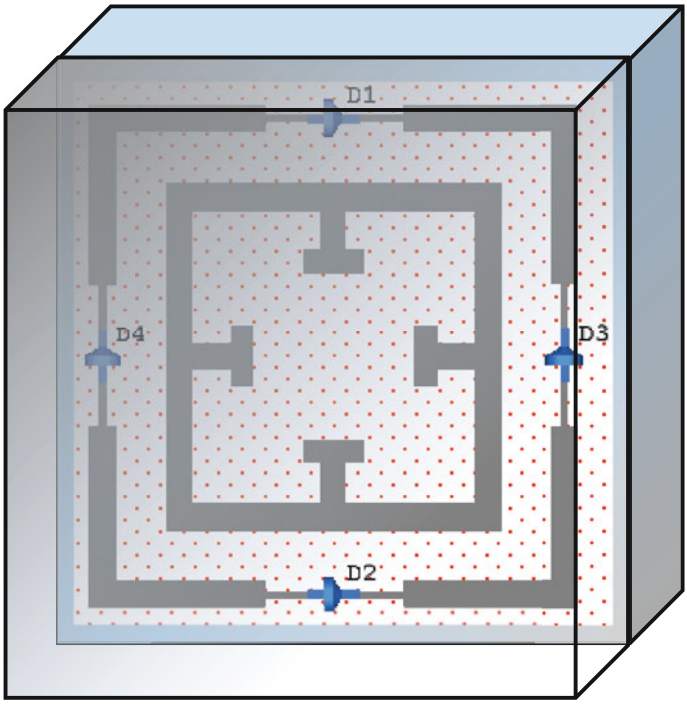


Fig. 48 Schematic of Quartz-epoxy-based varactor-loaded planar monolithic radome wall configuration

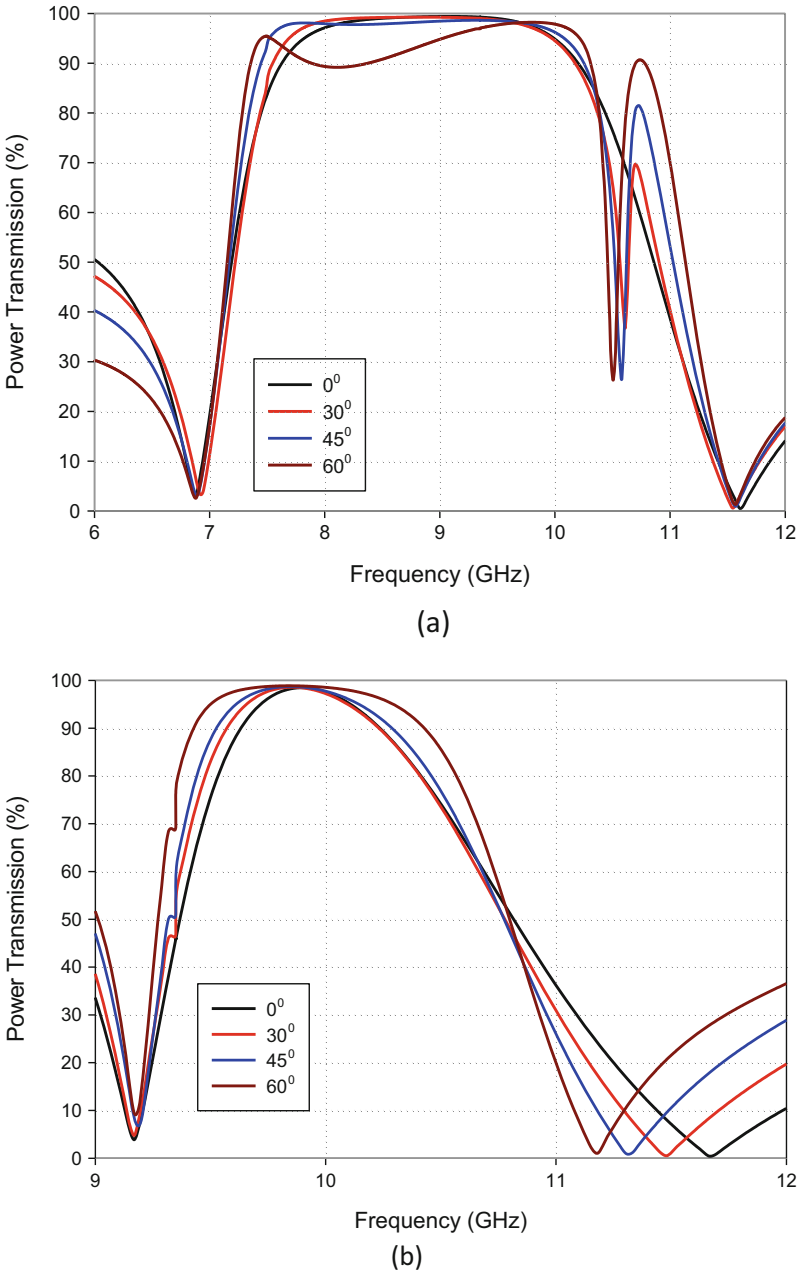


Fig. 49 Power transmission response of monolithic radome configuration embedded with varactor-loaded MTM-FSS structure for TE polarization at reverse-biased voltages: (**a**) 0 V and (**b**) 14.95 V

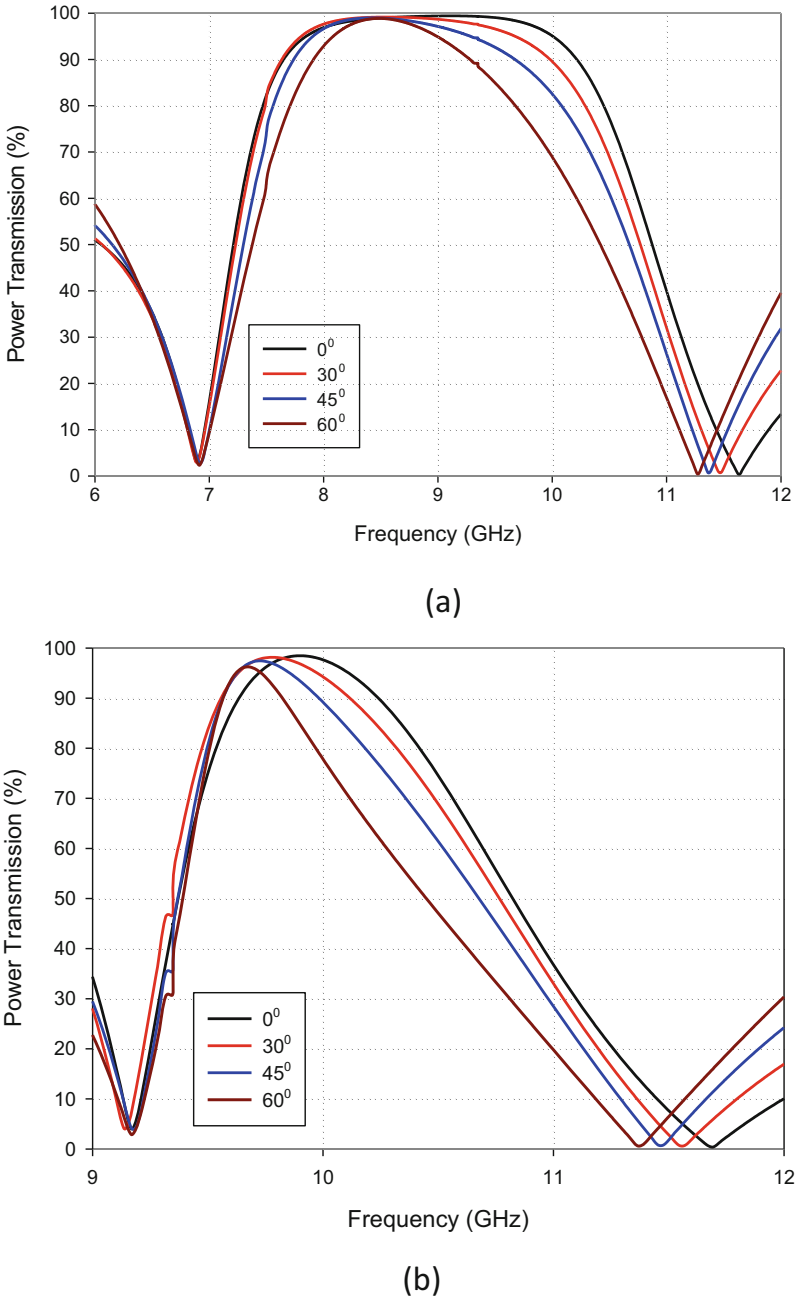


Fig. 50 Power transmission response of monolithic radome configuration embedded with varactor-loaded MTM-FSS structure for TM polarization at reverse bias voltages: (**a**) 0 V and (**b**) 14.95 V

bias voltages. For unbiased varactor (at 0 volts), the proposed MTM-FSS radome shows wideband bandpass characteristics over the frequency range 7.65–10.21 GHz (bandwidth = 2.56 GHz) w.r.t. 90% transmission characteristics efficiency at normal incidence. However, at higher angles of incidence, the power transmission efficiency degrades over the desired frequency band as shown in Fig. 49a.

By applying the reverse biased higher voltage (14.95 V), the center frequency of the radome is tuned from 9 GHz to 9.95. It is observed that lower resonances of the structure shift toward higher frequency side due to the varactor diode-capacitive effect at the outer ring of the MTM-FSS unit cell. Thus, the bandwidth of the radome wall configuration reduces from 2.56 GHz (at 0 volts) to 1 GHz (at 14.95 volts) as shown in Fig. 49b. For TM polarization, the proposed design shows more than 90% power transmission characteristics (i.e., −0.4 dB loss) over the frequency range, 7.65–10.21 GHz (bandwidth =2.56 GHz). For higher angles of incidence, when diode is biased at higher reverse voltage, there is degradation in bandwidth as depicted in Fig. 50.

The factors considered for the selection of a suitable EM design workflow of the radome wall configuration are radome material properties and thickness of each layer, dimensions of the embedded FSS unit cell, operating frequency, polarization of the radar/antenna system, etc. Range of incidence angle data is based on the location and tilt angle of antenna system within the radome and steering offsets. During the interaction of EM waves with radome wall, amplitude, and phase distortions, first-order effects must be accounted by the radome design technique. A suitable EM design methodology supports the computation of required EM performance parameters of proposed radome wall configuration (e.g., transmission, reflection efficiency, and insertion phase delay).

The presence of radome results in degradation of antenna's radiation characteristics due to phase and amplitude distortions when it passes through the radome wall. Transmission loss affects many parameters such as radar's range, system sensitivity, etc. Thus, 3D ray-tracing procedure based on *geometrical optics* method in combination with *aperture integration* technique is used to solve antenna-radome interaction problems (Nair and Jha 2007). The steps for EM performance analysis are the following: (i) altering the location of antenna-radiating slots to the radome coordinates, (ii) computing the space phase expressions for antenna far-field, and (iii) computation of EM performance parameters of radome as a function of antenna scan angle. The radome performance parameters such as copol power transmission, cross-pol power transmission, and boresight error (BSE) are computed considering the antenna in receiving mode.

In receiving mode approach (known as backward ray tracing), an external aperture is projected to the antenna's aperture plane through parallel lines. The copol and cross-pol voltage transmission coefficients are evaluated as function of radome material properties and angle of incidence at radome wall. Integrating the fields on aperture of the antenna gives the receiving port voltages, where sum and difference channel port voltages are calculated using Lorentz reciprocity theorem. Let E_t and H_t be the radiated electric and magnetic fields, respectively, from the slotted waveguide planar antenna array and, Er and H_r be the respective fields of the

plane wave illumination corresponding to reflected signals. Then, the port voltage is expressed by

$$V_{port} \propto \left[\int_s \left(\overset{\rho}{E}_t \times \overset{\rho}{H}_r \right) \cdot \hat{n} - \left(\overset{\rho}{E}_r \times \overset{\rho}{H}_t \right) \cdot \hat{n} \right] ds \tag{34}$$

Then

$$V_{port} = C \left[\int_s \left(\overset{\rho}{E}_t \times \overset{\rho}{H}_r \right) \cdot \hat{n} - \left(\overset{\rho}{E}_r \times \overset{\rho}{H}_t \right) \cdot \hat{n} \right] ds \tag{35}$$

where C represents a function of the current coefficient corresponding to the reflected fields. Equation (35) is modified further by vector operation, and return plane wave fields are replaced by induced currents J_r and M_r based on equivalence principle, and hence port voltage is taken expressed by

$$V_{port} = C \int_s \left[\overset{\rho}{E}_t \cdot \overset{\rho}{J}_r - \overset{\rho}{H}_t \cdot \overset{\rho}{M}_r \right] ds \tag{36}$$

The sum port voltage (Σ) and difference port voltages (ΔV_{AZ} and ΔV_{EL}) can be determined by eq. (36). Sum port voltage aids in the computation of copol and cross-pol power transmission parameter. The azimuth and elevation boresight errors are given by the following equations (Siwiak et al. 1979; Nair et al. 2015):

$$BSE_{AZ} = \frac{\Delta V_{AZ}}{K_{AZ}} \tag{37}$$

$$BSE_{EL} = \frac{\Delta V_{EL}}{K_{EL}} \tag{38}$$

where ΔV_{AZ} and ΔV_{EL} are the corresponding azimuth and elevation channel voltages for the given antenna scan angle. The term K indicates the mono-pulse antenna sensitivity, which is a function of difference channel port voltages and azimuth/elevation offsets.

In the transmitting mode, the far-field radiation pattern of the antenna in presence of radome is calculated by integrating the tangential field components on the external equivalent aperture. The equivalent aperture is assumed to be outside of the radome, and in this region EM waves propagate from antenna aperture through the radome wall in such a way that it is normal to the surface of radome. The far-field radiation pattern resulting from the antenna-radome system is given by

$$E_t = \sum_{m=1}^{M} \sum_{n=1}^{N} T_{mn} A_{mn} e^{-j\dot{k}(\mathrm{x}_m \sin\theta\cos\phi + y_n \sin\theta\sin\phi)}$$

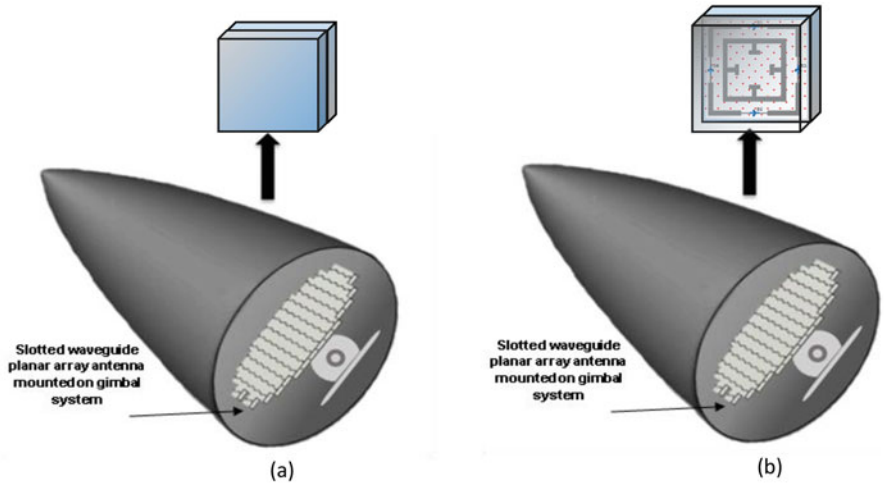


Fig. 51 Metamaterial-FSS-embedded tangent ogive radome; (**a**) without varactor diode, and (**b**) loaded with varactor diodes

where T_{mn} is the transmission coefficient corresponding to mn^{th} radiating slot on the antenna surface. The aperture distribution function of the mn^{th} radiating slot antenna is represented as A_{mn}, and (x_m, y_n) represents the location of antenna elements.

The transmission and reflection characteristics of proposed planar MTM-FSS based radome wall are computed with and without Varactor diode using full-wave method based simulation package, and prepared the Fresnel's transmission & reflection data table. These data are then fed into the program for the computation of radiation pattern, copol transmission characteristics, and boresight error (BSE). The previously proposed MTM-FSS structure has been utilized in the design of conformal tangent ogive radome wall configuration as shown in Fig. 51.

Table 7 illustrates the optimized dimensions of proposed MTM-FSS unit cell incorporated with/without varactor diode as shown in Fig. 52, which is optimized at the design frequency of radome for effective dielectric permittivity and thickness profile. A tangent ogive radome, having dimensions (height = 1000 mm; base diameter = 500 mm), enclosing an *active electronically scanned array onlenno* (AESA) (diameter = 300 mm) located at 300 mm from the base of the radome is considered. The antenna array is perpendicularly polarized (TE-pol) operating at the center frequency of 9 GHz. For the considered antenna system, the aperture distribution of the radiating elements is *Cosine*. The spacing between the slots of the planar array antenna is 16 mm and lattice geometry of radiating elements on the antenna surface is triangular. Radome wall thickness is optimized to be 10 mm at 9 GHz for better EM performance in terms of power transmission characteristics and boresight error (BSE).

The beam-steering range of AESA antenna is assumed to be $\pm 70^\circ$ in azimuth plane and $\pm 60^\circ$ in elevation plane. The performance parameters (like BSE) of

Table 7 Optimized design parameters of active MTM-FSS-embedded radome wall

Design parameters	Unit cell dimension without diode (mm)	Unit cell dimension with diode (mm)
Sl, Sw	9.00	8.00
Pl	0.50	0.35
Gl	1.00	0.58
G2	0.50	0.80
Tl	0.50	0.57
T2	1.00	1.20
Rl	8.00	5.70
R2	7.00	6.40
R3	4.96	4.54
R4	3.96	3.84
Tl	1.00	1
gl		0.7
g2		1.80
w		0.55

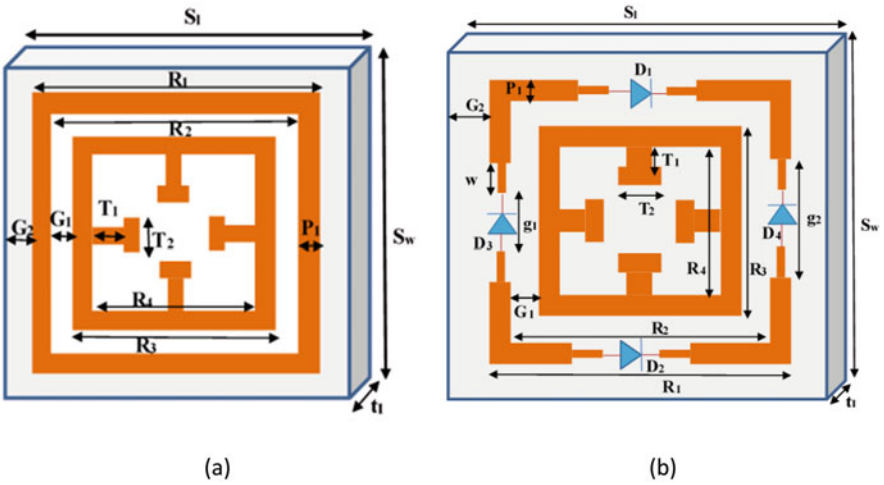
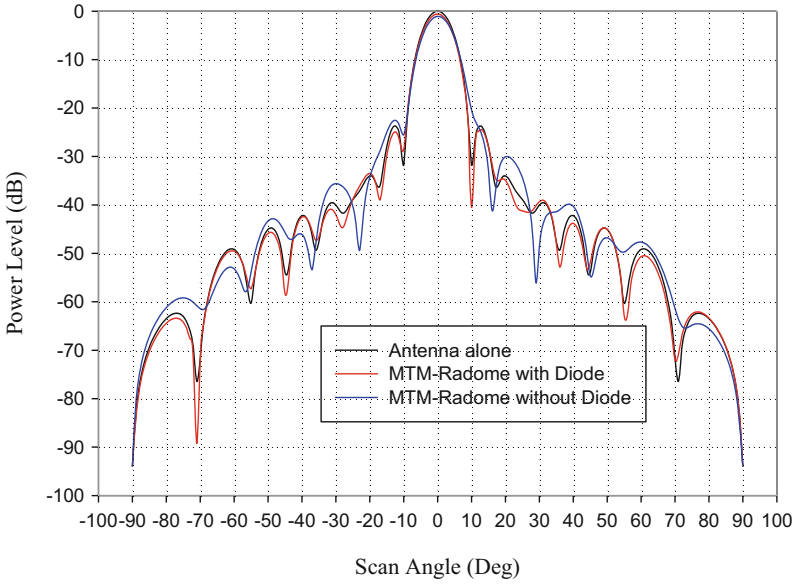


Fig. 52 MTM-FSS unit cell: (**a**) Without varactor diode, and (**b**) loaded with varactor diode

antenna-radome systems have been computed based on receiving mode technique, while radiation patterns have been computed based on transmitting mode approach.

Radiation Characteristics of Metamaterial FSS-Based Radome

Radiation pattern is computed in the presence of proposed MTM-FSS loaded with/without varactor diode-embedded radome at center frequency (9 GHz) for AESA orientation; azimuth (Az) angle = 0° and elevation (EI) angle = 0° shown in Fig. 53.



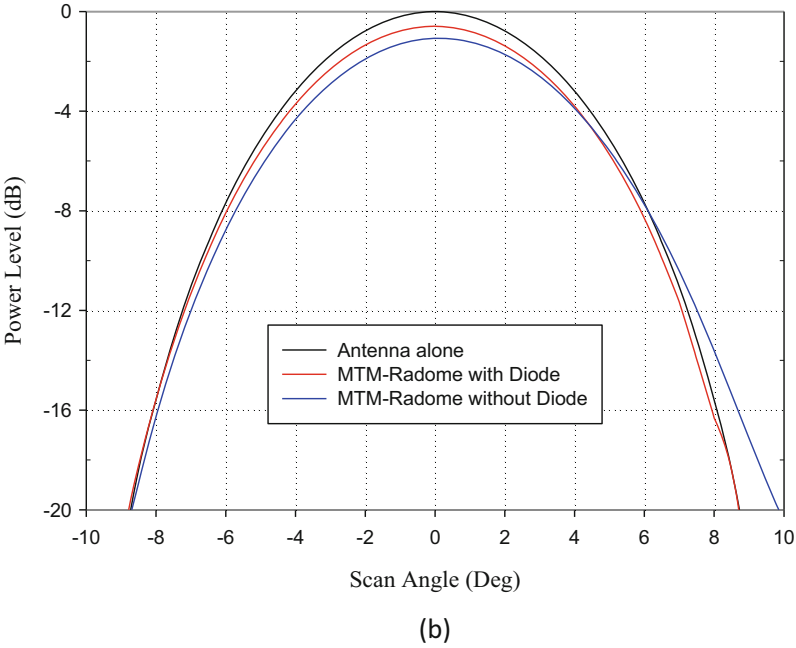


Fig. 53 (**a**) Radiation pattern of AESA antenna at central frequency, $f_0 = 9$, GHz for azimuth (Az) angle = 0°, and elevation (EI) angle = 0°; and (**b**) zoomed view of main lobe features of (**a**)

In addition, its radiation pattern is computed in presence of proposed radome wall incorporated with metamaterial unit cell loaded with and without varactor diode.

The power level of antenna main lobe is observed to be reduced to 0.59 dB down in presence of unbiased (at 0 volts) varactor-loaded MTM-FSS-embedded radome, while it is reduced to 1.068 dB down in presence of radome with MTM-FSS structure without varactor diode (Fig. 53b). Thus, the better power transmission is obtained by the varactor-loaded metamaterial radome as compared to without Varactor-loaded MTM-radome.

Further, the radiation characteristics of proposed MTM-FSS loaded with and without varactor diode-embedded radome are determined in presence of AESA antenna at higher band-edge frequency (9.72 GHz) for azimuth, (Az) angle $= 0°$, and elevation (EI) angle $= 0°$ as shown in Fig. 54. The power level of antenna main lobe is degraded to 0.46 dB down in presence of reverse biased (at 14.95 volts) varactor-loaded MTM-FSS-embedded radome, whereas it is reduced to 1.63 dB down in presence of radome with MTM-FSS structure only (Fig. 54b). Thus, the better power transmission (more than 90%) is depicted by the proposed varactor-loaded metamaterial radome as compared to without varactor-loaded MTM-radome.

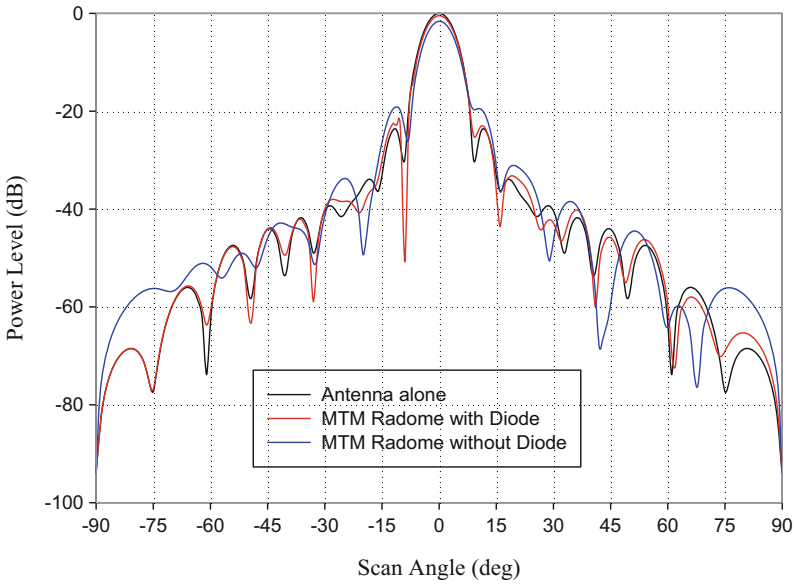
BSE Characteristics of Metamaterial FSS-Based Radome

Figure 55a depicts the azimuth BSE characteristics of the proposed radome embedded with MTM-FSS structure loaded with and without varactor diode corresponding to beam-steering angle of AESA for perpendicular polarization at center frequency (9 GHz). It is observed that BSE of MTM radome with unbiased (0 V) varactor diode is better than the proposed MTM-FSS without varactor diode radome. Figure 55b shows the azimuth BSE characteristics of the proposed radome embedded with MTM-FSS structure loaded with and without varactor diode corresponding to beam steering angle of AESA for perpendicular polarization at higher (9.72 GHz) band-edge frequency. lt is observed that up to 11° of beam steering, BSE of MTM radome with biased (14.95 V) varactor diode is better than the proposed MTM-FSS without varactor diode radome.

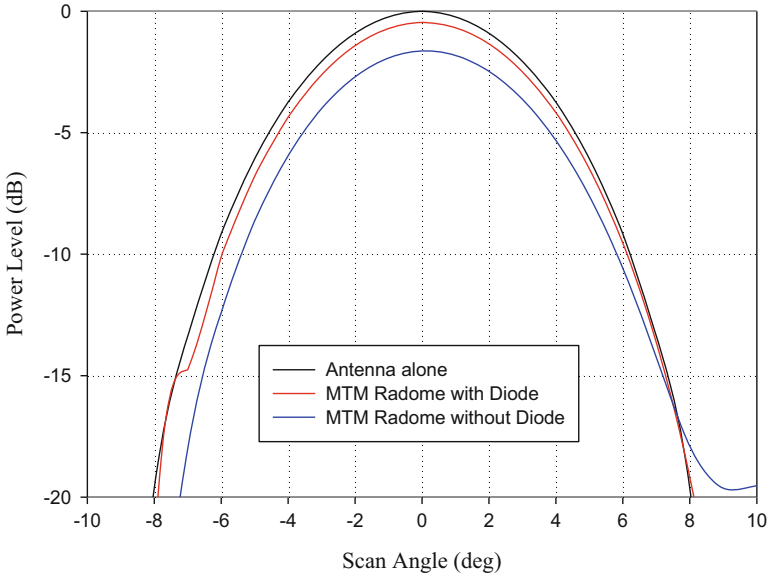
RCS Estimation

Due to the lack of computational resources, planar radomes with metamaterial FSS and PIN diodes of size 90 × 90 mm are modeled instead of ogive radome as shown in the Fig. 56, to perform the RCS analysis. The simulated bistatic RCS of a monolithic radome, metamaterial FSS-based radome without diodes, and metamaterial FSS-based radome with diodes are compared as shown in the Fig. 57. lt is observed that bistatic RCS of planar radome embedded with diode shows drastic RCS reduction to-22 dBsm at higher band-edge frequency, 9.72 GHz, as compared to without diode.

To summarize this work, the tangent ogive radome embedded with active MTM-element-based FSS exhibited superior EM performance characteristics as compared to MTM-element-based FSS radome with a significant RCS reduction. The proposed active metamaterial radome wall configuration showed excellent copol power transmission characteristics and minimal pattern degradation as compared to

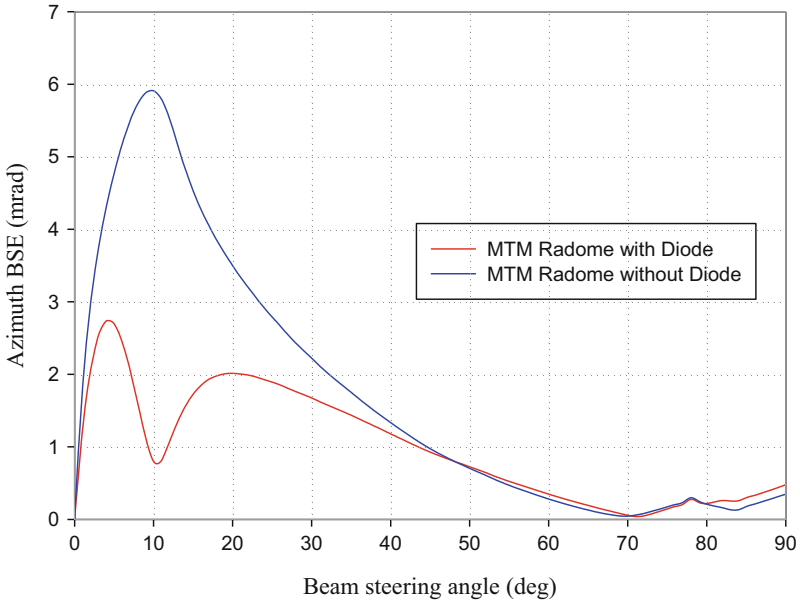


(a)



(b)

Fig. 54 (**a**) Radiation pattern of AESA antenna at frequency, $f_H = 9.72$, GHz for azimuth, (Az) angle = 0°, and elevation (EI) angle = 0°; and (**b**) zoomed view of main lobe features of (**a**)



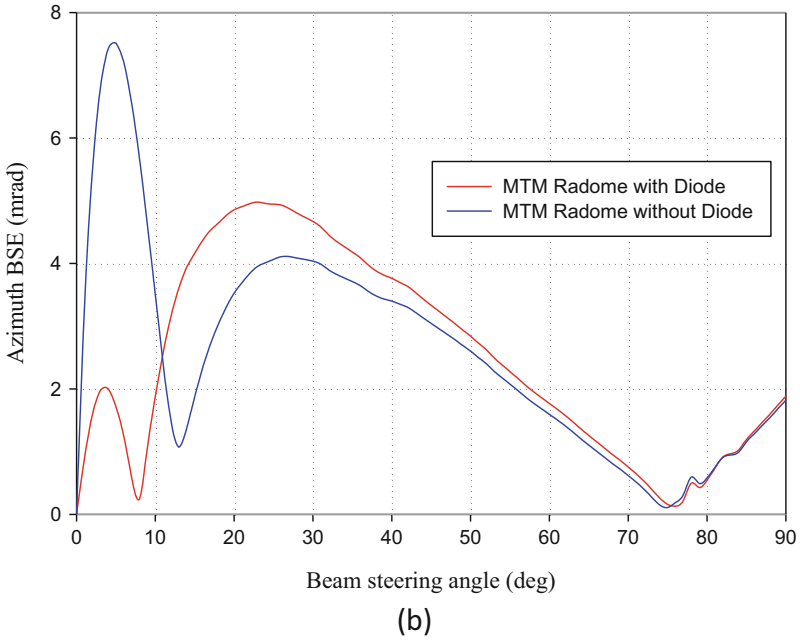


Fig. 55 Azimuth BSE characteristics versus beam-steering angle of tangent ogive radome embedded with and without varactor-loaded MTM-FSS: (**a**) at center frequency, 9 GHz, (**b**) at higher band-edge frequency, 9.72 GHz

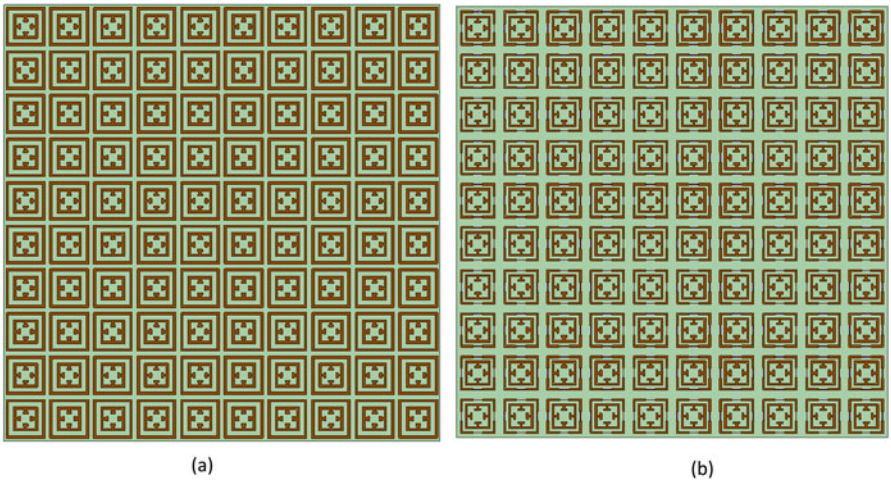


Fig. 56 Planar radome modeled for RCS estimation: (**a**) MTM radome without diode, (**b**) MTM radome with diode

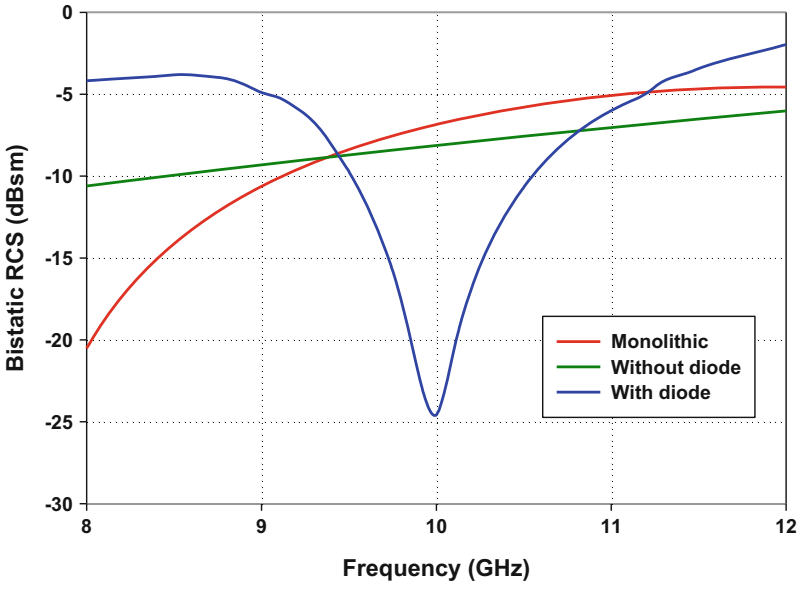


Fig. 57 Bistatic RCS of planar monolithic radome, MTM radome without diode, and MTM radome with diode in boresight direction (Az = 0, Ez = 0)

conventional MTM radome. Further, the proposed configuration has been observed to report azimuth/elevation BSE for TE polarized wave within permissible values as airborne radar requirements. lt is envisioned that the proposed MTM-FSS loaded

with varactor diode embedded in radome wall configuration can be employed for airborne platforms. However, hardware realization of such active streamlined airborne radome would be a challenging task.

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Metamaterial-based High-Performance Radar Absorbing Structure

2

R. S. Nilotpal and Somak Bhattacharyya

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Abstract

In this chapter, metamaterial-based absorbers have been discussed. Initially, the chapter deals with the evolution of metamaterial absorbers by introducing the design limitations of conventional absorbers. The reason for unity absorption in a metamaterial absorber has been explained based on different absorption mechanisms and different types of losses associated with the structure. A simulative study on the role of dielectric loss tangent on the absorptivity response has been discussed in detail. The chapter presents the design of various single-band, multi-band, and broad-band metamaterial absorbers in the GHz (S, C, and X), THz, and IR (Infrared) frequency bands. The experimental and testing methods have been discussed stepwise. The chapter also demonstrates the absorption mechanism based on multiple reflection interference theory in a detailed way. The ambiguity found in the resonance theory of absorption (based on electric and magnetic resonances) has been well explained through multiple interference theory. The last section of the chapter presents the mathematical modelling of the metamaterial absorbers. All the characteristics such as absorption, polarization insensitivity, and surface current orientation have been validated mathematically in light of electric and magnetic fields.

Keywords

Metamaterial absorbers · Metasurface · Absorption · Transverse electric · Transverse magnetic · Magnetic resonance · Electric resonance

Introduction to Metamaterial

The property of any naturally available materials mainly depends on the crystalline structure and the type of bond existing inside the structure (Callister and Rethwisch 2010). For example, both diamond and graphite are made entirely out of carbon, but their electrical and mechanical properties completely differ due to their crystalline arrangements. In another way, it can be said that the arrangement of atoms inside a material plays a crucial role in determining their electric and magnetic properties. The work on the exploration of artificial materials to manipulate the electromagnetic waves began in the early nineteenth century. The idea was to design an artificial material made up of very small metallic elements (called as meta-atoms), such that by artificially manipulating the shape and size of the meta-atoms, the electromagnetic property can be artificially manipulated. In this direction, a lot of research work was initiated in the twentieth century; however, the success came at the beginning of the twenty-first century (2000) when John Pendry practically demonstrated the simultaneous existence of negative permittivity and permeability value for an artificial material over a certain frequency range (Pendry 2000). This property of a material having both negative permittivity and permeability cannot be readily found in nature. For these artificial structures, the term metamaterial was first given by

Walser in the year 2001 (Walser 2001). Initially, most of the researchers had a misconception that metamaterial structures always have negative permittivity and permeability value. However, the truth is that the metamaterial provides the flexibility to achieve any value of permittivity and permeability merely by optimizing the metallic patterns (meta-atoms). In this way, metamaterial provides an artificial way to design a high-performance lens (Neu et al. 2010; Haxha et al. 2018), antennas (Hashmi et al. 2014; Liu et al. 2015; Zheng et al. 2018; Samantaray et al. 2019; Bai et al. 2020; Samantaray and Bhattacharyya 2020; Ghosh et al. 2020), filters (Nilotpal and Upadhyay 2016; Mishra et al. 2017; Shen et al. 2018, Bhattacharyya et al. 2017), polarization converters (Nilotpal et al. 2019; Saikia et al. 2017; R. Nandi, Nilotpal, and Bhattacharyya 2019; Yadav et al. 2018; Bhattacharyya et al. 2017), absorbers (Bhattacharyya et al. 2014; Dhillon et al. 2019; Nilotpal and Chakrabarti 2019; Hannan et al. 2020; Ghosh et al. 2020), plasmonic sensors (Liu et al. 2010; Wu et al. 2016; Kayal et al. 2020), and infrared photodetectors (Zhang et al. 2013; Jing et al. 2016) over the broad frequency band ranging from microwave to visible.

Electromagnetic metamaterials are composite arrays of artificially designed subwavelength size metallic elements that can be artificially optimized to tune the electromagnetic property of the medium such as permittivity, permeability, refractive index, and impedance (Nilotpal and Chakrabarti 2018). The densely arranged periodic metallic structures allow the wave to be engineered in a much better way, thus providing a better spatial resolution. Like the case of a bulk material where the property of the material mainly depends on the arrangements of atoms inside the structure, the property of a metamaterial can be determined by the arrangement of subwavelength size metallic structures, also known as meta-atoms or unit-cells.

The whole metamaterial structure is characterized by the effective medium parameters, i.e., effective permittivity $\varepsilon(\omega)$ and effective permeability $\mu(\omega)$ (Chen et al. 2016). It is due to the fact that the electromagnetic wave does not sense the presence of the individual unit-cells (meta-atoms) due to their subwavelength dimension. Hence the characterization of the metamaterial structure is done as a whole bulk structure rather as an individual unit cell. Also, according to the homogenization principle, the length of the unit cell should be very small as compared to the guided wavelength, i.e., $l << \lambda_g$, where λ_g is the guided wavelength. Usually, for a medium to be considered homogenous, the minimum requirement is ($l < \frac{\lambda_g}{4}$). The homogeneity limit ensures that whenever the wave interacts with the metamaterial structure, the phenomena like reflection and refraction should dominate over scattering and diffraction (Caloz and Itoh 2006).

Introduction to Metasurface

Though metamaterial and metasurface belong to the same class of artificially engineered structures, there is a slight difference between them. Metamaterial can be considered as a bulk array of resonating metallic structures (having a finite length, width and thickness). In contrast, a metasurface can be considered as a planar

two-dimensional array of resonating metallic structures. One can say that metasurfaces are the two-dimensional analogue of metamaterial where the thickness is considered as ultra-thin in nature lying in the sub-wavelength regime. Most of the functionalities which can be realized by a metasurface can also be realized by a bulk metamaterial. However, despite the fact that there are three different reasons in favor of use of metasurface in the place of metamaterial (Glybovski et al. 2016).

1. It is tough to implement the bulk metamaterial structures at high frequencies.
2. The bulk metamaterial structure offers the propagation of electromagnetic waves for a substantial distance, which led to a high dielectric loss in the structure. The low absorption in the metasurface allows the waves to be engineered in a more profound way.
3. The metasurfaces are very thin in nature which allows them to integrate easily with other electronic systems.

Conventional Electromagnetic Wave Absorbers

Electromagnetic wave absorbers can be used in the application of solar harvesting (Areed et al. 2016; Mullah and Sabah 2016; Bagmanci et al. 2017, 2019), sensors (Ozer et al. 2016; Saadeldin et al. 2019; Vafapour 2019), and wireless communication (Chen et al. 2014; Wang et al. 2016). For absorbing electromagnetic waves, the first absorber was designed by Winfield Salisbury in the early 1940s (Salisbury 1952). The Salisbury screen consists of three layers; (1) the first layer is a thin resistive screen, (2) the middle layer is a low loss dielectric (usually air is considered) having thickness equal to an odd multiple of quarter wavelength ($n\frac{\lambda_g}{4}$), and (3) the bottom layer which is a continuous metallic layer. The operating principle of the Salisbury screen mainly depends on the destructive interference theory, as explained through Fig. 1. The wave incident on the top of the resistive screen is

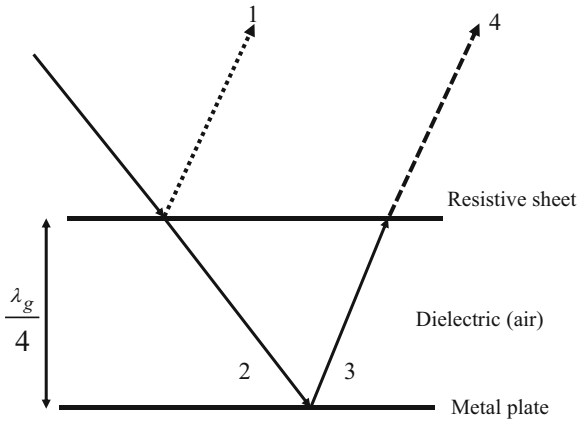

Fig. 1 Schematic of the Salisbury screen

reflected back as wave 1 and transmitted inside the dielectric layer as wave 2. Wave 2 gets reflected completely by the bottom metal sheet as wave 3 and again received back as wave 4. Inside the dielectric layer, the wave travels a total distance of $\frac{\lambda_g}{2}$ in one complete cycle of incidence and reflection. Thus, as compared to wave 1, wave 4 experiences a total phase change of 180° leading to the destructive interference between the two waves. This makes sense that the wave which is incident on the structure does not come out. In other words, the wave gets absorbed.

The principal operation of the Salisbury screen can also be understood by transmission line theory. According to the transmission line theory, the quarter wavelength size of the dielectric layer transfers the short circuit at the bottom layer to an open circuit at the resistive layer (Lee et al. 2016). The effective impedance of the structure is the impedance of the top resistive sheet. If the resistive sheet impedance is that of the free space, a complete impedance matching can occur. Suppose, instead of a quarter-wavelength size dielectric layer, we consider the dielectric layer of half-wavelength, in that case the short circuit at the bottom reappears at the top and hence there will be a complete reflection and no absorption (Cheng 1983). Nevertheless, Salisbury screen has two major limitations, i.e., thickness and the operating band. The quarter-wavelength size of the dielectric makes the structure thick for low-frequency operation. Also, the condition of destructive interference can only be satisfied for a fixed frequency. Hence the absorption bandwidth is very narrow.

The limitation of narrow bandwidth in a Salisbury screen was later on corrected by Jaumann absorbers (Knott and Lunden 1995), and Dallenbach absorbers (Dallenbach and Kleinsteuber 1938), but that also on the cost of extra thickness due to stacking of layers.

Circuit Analog Absorbers

The circuit Analog (CA) absorbers offer a broader bandwidth response than the Salisbury screen or Jaumaan absorbers. Therefore, the problem of stacking can be solved by the use of CA absorbers (Munk 2000). The design structure of a CA absorber is very similar to the Salisbury screen except the top layer's design. In a CA absorber, the top layer consists of a periodic arrangement of conducting metal patches instead of a resistive sheet. The shape of the top conducting patch could be either a cut wire type, cross wire type, or a split ring resonator type. The length of the metallic patch corresponds to an inductance, whereas the gap between two patches corresponds to a capacitance. Hence the CA absorber can be represented as basic circuit elements, i.e., R, L, and C. The role of the middle layer in a CA absorber is quite similar to the role of the middle layer in Salisbury screen. In CA absorber, the size of the middle layer is approximately quarter wavelength of the resonance frequency of the RLC circuit of the FSS. Therefore, the problem of thickness which was there in Salisbury screen still remains in CA absorbers.

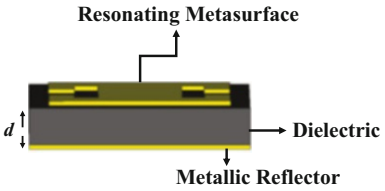

Fig. 2 Side view of the unit cell of the metamaterial absorber

Metasurface Based Absorbers

Metamaterials can be used to artificially manipulate the effective medium parameters, i.e., dielectric permittivity and permeability of the medium (Bhattacharyya et al. 2015). This provides the flexibility to efficiently manipulate and control the absorption of the electromagnetic wave as compared to the naturally available absorbing material. Mostly a metamaterial absorber is a three-layer structure, as shown in Fig. 2. The top layer consists of periodically arranged metallic patches, middle layer is a dielectric, and the bottom layer consists of a continuous metal sheet. The metallic patches at the top layer are very similar to that of the CA absorber. However, the thickness of the middle layer in a metasurface absorber is very small as compared to the thickness of the middle layer in a CA absorber. To realize unity absorption, there should be zero reflection and zero transmission simultaneously from the structure. The zero reflection can be obtained by optimizing the top metallic patches whereas the presence of continuous metallic plate at the bottom surface works like a metal backing, which blocks all the wave to transmit out from the bottom.

The middle dielectric layer provides a space for the wave to travel inside and get absorbed. To decrease the thickness of the dielectric layer, it is always desirable to select a dielectric having a high dielectric constant (ε), so that the optical path length (ηd) does not get affected. Here, η represents refractive index. Also, along with a high dielectric constant value, the dielectric should have a small loss tangent value, which is represented as an imaginary part of the dielectric constant. The presence of an imaginary dielectric constant signifies the dielectric loss in the medium. Henceforth, the small value of loss tangent is one of the reasons for absorption inside the dielectric layer.

Types of Absorption in a Metamaterial Absorber

The two main reasons for absorption in a metamaterial absorber are (1) dielectric loss and (2) ohmic loss. The dissipation of energy due to the dielectric layer is counted as a dielectric loss. This loss mainly depends on the loss tangent of the dielectric layer. A study of absorptivity response for different loss tangent values for a typical metamaterial absorber operating at 7.45 GHz is shown in Fig. 3. For zero loss tangent (lossless dielectric), the absorptivity falls down to 40% at the absorption

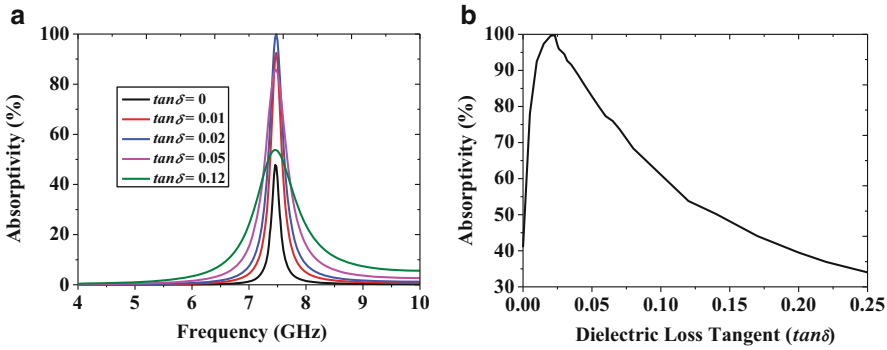

Fig. 3 (**a**) Absorptivity response and (**b**) absorptivity variation with dielectric loss tangent of the FR-4 substrate for a metamaterial absorber operating at 7.45 GHz

frequency 7.45 GHz, as shown in Fig. 3(a). Also, from Fig. 3(b) it can be seen that initially, the absorptivity value increases with an increase in the loss tangent which reaches a maximum at the loss tangent of 0.023, after which a gradual decrease in the absorptivity response can be observed. The initial increase in the absorptivity response is due to the fact that with an increment in the loss tangent value, the dipole present in the dielectric undergoes enough friction to orient themselves in the direction of the electric field. However, with a further increase in the loss tangent value, the applied electric field fails to orient the dipoles in the direction of the electric field which is responsible for decrease in the absorptivity response (Hu et al. 2010).

Ohmic loss is the loss in the structure due to the resistance offered by the metallic patches. The loss due to the dielectric and metal gives rise to heating in the structure, also called Joule's heating. In the metamaterial absorber both the losses take part in the absorption mechanism. Nevertheless, the dominance of any one of the loss mechanisms primarily depends on the frequency of operation.

In the lower frequency range (GHz), the metals usually act as a good conductor. Hence, the ohmic loss is very small in the lower frequency range. The loss in the metamaterial absorber at GHz frequency range is primarily due to the dielectric loss. However, as the frequency increases, ohmic loss dominates over the dielectric loss. This is because metal becomes very lossy at terahertz/infrared frequencies, due to their interband transitions. Therefore, at terahertz/infrared frequencies metallic loss dominates over dielectric loss.

Absorption Mechanism in a Metasurface Absorber

This section, primarily focus on understanding the mechanism to enhance absorption in the metasurface absorber. The two main conditions necessary for absorption in a metamaterial absorber are (1) impedance matching condition (2) excitation of electromagnetic resonances.

Impedance Matching Condition

Whenever an electromagnetic wave strikes the top of the metamaterial, either the wave gets scattered, or refracted, or reflected, or transmitted, or absorbed. Assuming the fact that the surface roughness is very small as compared to the wavelength of the incident wave, there is no chance of scattering from the surface. Hence the incident wave may be reflected, transmitted or absorbed. According to Kirchhoff's law, the relation between reflectance $R(\omega)$, transmittance $T(\omega)$, and absorbance $A(\omega)$ must follow Eq. 1.

$$A(\omega) = 1 - R(\omega) - T(\omega) \tag{1}$$

The amount of wave reflected and transmitted from the interface of two media can be solved by applying Maxwell's equation under suitable boundary and field continuity conditions. When a plane wave travelling in a homogeneous medium encounters a different medium, either the wave get reflected back or transmitted.

Suppose a plane wave impinges at an angle θ_i on a media interface that separate medium 1 (i.e., air) (μ_1, ε_1) with medium 2 (μ_2, ε_2). In that case the reflectance (rim) from the interface (r_{tm}) for both transverse magnetic (TM) and the reflectance (r_{te}) from the interface for transverse electric (TE) polarized incident waves are given by Fresnel's equation, as shown in Eqs. 2 and 3, respectively (Cheng 1983).

$$r_{tm} = \frac{\sqrt{\mu_r \varepsilon_r - \sin^2\theta} - \varepsilon_r \cos\theta}{\sqrt{\mu_r \varepsilon_r - \sin^2\theta} + \varepsilon_r \cos\theta} \tag{2}$$

$$r_{te} = \frac{\mu_r \cos\theta - \sqrt{\mu_r \varepsilon_r - \sin^2\theta}}{\mu_r \cos\theta + \sqrt{\mu_r \varepsilon_r - \sin^2\theta}} \tag{3}$$

Here, μ_r and ε_r are the relative permittivity and permeability of the second media. Consider the case when medium 1 is air and medium 2 is a dielectric, and assume that the wave is impinging normally on the media interface, then reflectivity (r) equation can be modified, as shown in Eq. 4.

$$r = \frac{1 - \sqrt{\mu_r \varepsilon_r}}{1 + \sqrt{\mu_r \varepsilon_r}} \tag{4}$$

In order to understand the relation between electric and magnetic fields, Eq. 5 can be derived in terms of impedance where the characteristics impedance of the medium is the ratio between electric and magnetic fields, as shown in Eq. 5.

$$r = \frac{Z_2 - Z_1}{Z_2 + Z_1} \tag{5}$$

In Eq. 5, Z_1 is the characteristic impedance of medium 1 (air) which is given as $Z_1 = \sqrt{\frac{\mu_1}{\varepsilon_1}}$ and Z_2 is the characteristics impedance of medium 2 (absorber), which is calculated as $Z_2 = \sqrt{\frac{\mu_2}{\varepsilon_2}}$.

The reflection coefficient derived from the Fresnel's equation is complex in nature. Therefore, the reflectance $R(\omega)$ can be calculated as

$$R(\omega) = r \times r^*$$

From the equation of $R(\omega)$, one can understand that if the metallic patches on the top layer are optimized in such a way that the characteristic impedance of medium 2 (absorber) becomes equal to the characteristic impedance of medium 1, i.e., air, then there will be no reflection from the structure and hence $R(\omega) = 0$. In such condition, the metamaterial absorber shows unity absorption. This can also be understood from Eq. 1, which suggests that for unity absorption both transmittance and reflectance should equal to zero. Matching impedance of medium 2 with the impedance of medium 1 satisfies the condition, i.e., $R(\omega) = 0$. Also, due to the presence of metal at the bottom there is no transmission from the structure, hence $T(\omega) = 0$. Thus, according to Eq. 1, the absorption $A(\omega)$ becomes equal to unity. In order to achieve a matched condition, the real part of effective permittivity and permeability should be equal to each other ($\mu_r = \varepsilon_r$). There is no natural material that can meet this requirement; hence this can only be done artificially by engineering the metallic pattern of the structure in order to tune the ratio of E and H. Hence, the term "meta" appears here which means beyond. Here, the absorption has been achieved by tailoring the geometrical dimensions and shape of the pattern.

Absorption Based on Electromagnetic Resonances

The absorption mechanism is explained on the basis of simultaneous excitation of electric and magnetic resonances at the frequency of absorption. When a time varying electromagnetic field is incident on the top arm of the ERR resonator, it induces a current (conduction current) in the ERR. The current induced in the top arm of the ERR will create a mirror image in the bottom metallic strip, as shown in Fig. 4.

These two antiparallel surface currents along with the displacement current in the dielectric layer (can be treated as a capacitive gap) form a circulating current loop.

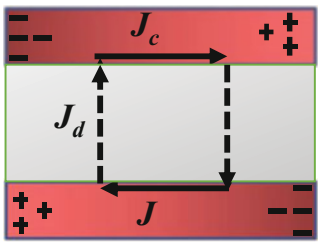

Fig. 4 Antiparallel surface currents forming a circulating current loop in a metasurface absorber

A Lorentz like magnetic response can be seen when due to the circulating current loop a time varying magnetic field induces with the magnetic dipoles.

Simulation of a Metasurface Absorber in CST Microwave Studio

The design and simulation of an entire array or a periodic structure array is a time- as well as memory-consuming process. The use of unit-cell periodic boundary condition in any electromagnetic solver allows a fast but no less accurate simulation response of an infinitely large periodic structures. The major advantage of Microwave CST simulation software is it automatically sets all the initial conditions required for the structure to realize an infinite one.

The basic modelling steps in CST Microwave studio are given below:

1. Choose an application area from the set of areas provided by CST and then select FSS, Metamaterial-Unit Cell from the workflows.
2. Select the recommended solvers for the workflows.
3. Set the frequency of operation and accordingly assign the dimension.
4. Draw the geometry.
5. Define the material parameters and define the boundary condition.
6. Define the mesh and run.
7. Post processing: Plot the 1-D result for S-parameters and 3-D result for electromagnetic fields.

The first step is to design the structure in CST for which a number of geometry options are provided such as rectangle, box, circle, cylindrical. If there are number of individual structures they can be added or subtracted by the Boolean operation. The structure is covered from an air boundary from all its side. The structure is covered from an air box surrounding from all its side. The main reason to provide a boundary is to restrict the computational area of the solver. The large boundary requires more time to simulate but provide more accurate results. A 3D model of an array of tri-layer metamaterial structure in CST Microwave studio is shown in Fig. 5.

The default Floquet port settings excite two orthogonal modes of plane waves, viz., TE (0,0) and TM (0,0) modes. However, other higher order modes can be excited from the port properties section. The S-parameters are represented in terms of co- and cross-pol coupling between the modes. For an example: co-polarization reflection of mode 1 at Floquet port 1 (Zmax) is represented as SZmax (1), Zmax (1), and the cross-polarization reflection between mode 1 and mode 2 is represented as SZmax (1), Zmax (2).

Metasurface Absorber Operating in Different Frequency Ranges

Metamaterial absorbers have gained significant attention due to their light wave and subwavelength profiles. The ease of artificial manipulation of effective parameters has increased the diversification in the application. A metamaterial absorber can be

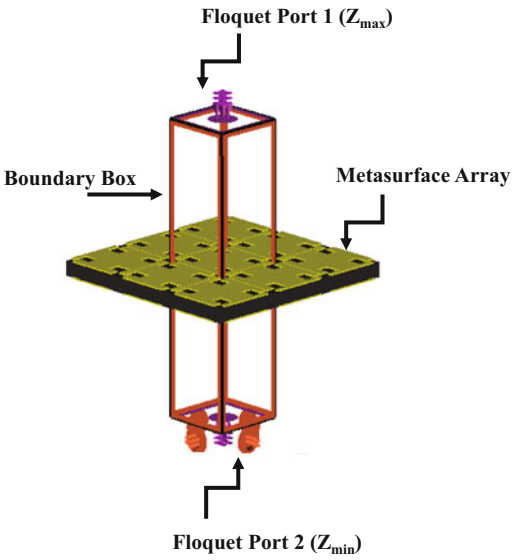

Fig. 5 3D model of an array of a tri-layer metamaterial structure in CST Microwave studio

used for the application ranging from microwave to visible frequency regions. The operating frequency of a metamaterial absorber can be scaled up (high frequencies) or down (low frequencies) by merely changing the dimensions of the metamaterial absorber to operate them at GHz, THz, infrared, or optical ranges. Apart from scaling the dimensions, the other factors which should be taken care is the property of the material (metal, dielectric) in that frequency regions.

Metasurface Absorber Operating in GHz Frequency Range

The first metamaterial-based absorber in the microwave frequency band was demonstrated by Landy et al. in the year 2008 (Landy et al. 2008). The unit-cell of the absorber consists of two metallic plates separated by a dielectric layer. The top layer is an electric ring resonator (ERR), whereas the bottom layer is a rectangular metallic strip as shown in Fig. 6.

The ERR present at the top layer gets coupled strongly to the incident electric field and gives rise to enhancement in the electric field and dispersion in the dielectric permittivity. The time varying electric field will induce electric current in the ERR. The current in the ERR induces a mirror current in the bottom metallic strip. This antiparallel electric current and the displacement current in the dielectric layer will give rise to strong magnetic resonance in the structure. The presence of a metallic strip at the bottom suppresses any transmission from the structure. Thus, the overall phenomenon of matching impedance and electromagnetic resonance gives rise to strong absorption of 96% (at 11.5 GHz) in the structure. Experimentally, the absorptivity was found to be 88% which the authors claimed that the discrepancies between the simulated and fabricated results might be due to the fabrication error.

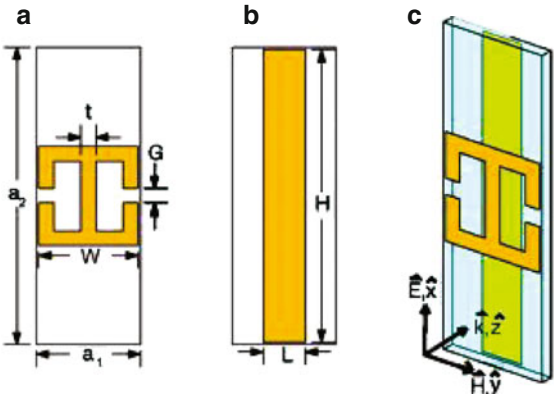

Fig. 6 (**a**) Top layer of the metamaterial structure consisting of an ERR (**b**) bottom layer consisting of a metallic patch and perspective view of the metamaterial structure. (Reproduced from Landy et al. (2008))

A dual band metamaterial absorber was reported by Li et al. in the year 2010 (Li et al. 2010). The structure consisted of an ERR on the front layer and a cut-metallic patch on the back layer. In between ERR and metallic patch, a FR-4 substrate has been taken. The simulated and fabricated results show that the structure has two resonating points at 11.15 and 16.01 GHz. At both the frequencies, the structure maintains an absorption of over 97% at low-frequency peak and 99% at high-frequency one. The absorption mechanism is based on impedance matching condition and antiparallel surface current formation. A triple band metamaterial absorber was theoretically and experimentally investigated by Shen et al. (Shen et al. 2011). The structure is a three-layered structure where the top layer consisted of an array of three nested copper closed ring resonators, responsible for the electric response to the incident EM wave and the bottom layer consists of a copper plane, responsible for zero transmission from the structure. The structure has an absorptivity of 99%, 96%, and 98% at frequency 4.02 GHz, 6.75 GHz, and 9.24 GHz, respectively.

A few years later, a single-band metamaterial absorber working in the C band was reported in (Bhattacharyya et al. 2012). The top view of the unit cell along with the electromagnetic field direction is shown in Fig. 7(a). The top layer consists of a resonating metallic pattern whereas the bottom layer consists of a continuous metallic plate that suppresses any transmission from the structure. For the metallic components copper has been used, whereas for the dielectric FR-4 has been used whose dielectric constant is 4.25 and loss tangent is 0.025. The thickness of the metallic plate has been optimized to 0.035 mm. The dimensions of the structure are further optimized as $a = 5$ mm, $d = 3.6$ mm, $w = 0.4$ mm, and $g = 0.2$ mm. It can be seen that the structure exhibits an absorptivity of more than 95% at frequency 7.46 GHz, as shown in Fig. 7(b).

The absorption mechanism of the structure can be explained on the basis of the electric field distribution on the top metallic plates and the surface current orientation on the top and bottom metallic plates as shown in Figs. 8(a)–(c) respectively.

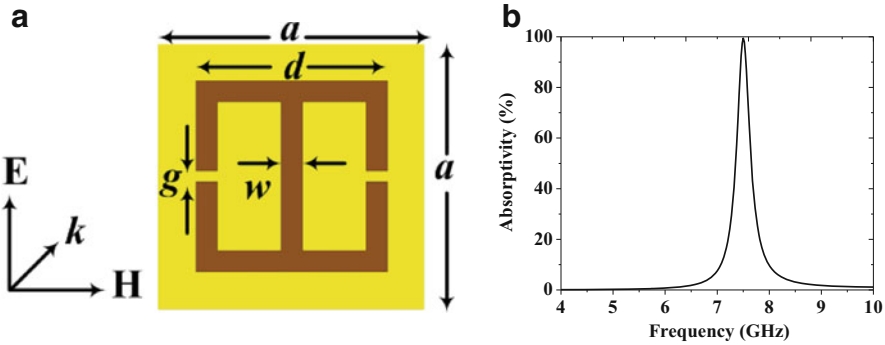

Fig. 7 (**a**) Top view of the metamaterial absorber along with the (**b**) absorptivity response. (Reproduced from Bhattacharyya et al. (2012))

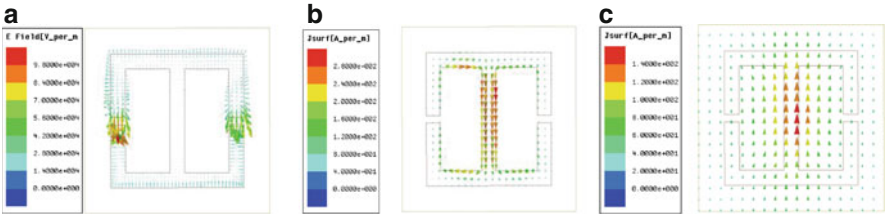

Fig. 8 (**a**) Electric field distribution on the top metallic plate and surface current orientation on the (**b**) top and (**c**) bottom metallic plates at the frequency of absorption, i.e., 7.45 GHz. (Reproduced from (Bhattacharyya et al. 2012))

The metallic structure on the top acts as an electric resonator driven by the electric field of the incident electromagnetic wave. The incident electromagnetic wave excites the free electron present in the metal, leading to the oscillation of the electrons. Depending on the shape and size of the metallic patches the electron resonates at a particular frequency forming an electric dipole at the metallic edges. It can be seen from Fig. 8(a) that at the edges, electric field is very strong, validating the formation of opposite charges (electric dipoles). Also, it can be seen from Fig. 8(b) and Fig. 8(c) that the directions of surface current at the top and bottom metallic plates are opposite to each other which gives rise to magnetic resonance in the structure. The simultaneous excitation of electric and magnetic resonances together at the frequency of operation is responsible for absorption in the structure.

Multiband Metamaterial Absorbers

To constitute a dual-band absorbing structure, the top layer must have a metallic pattern that can either be of two different shapes or two different dimensions where each shape or dimension corresponds to a unique frequency. Keeping this in mind the work presented in Bhattacharyya et al. (2014) has scaled up the metallic pattern of the unit cell shown in Fig. 7(a) by two different factors. The complete unit cell of the dual-band absorber is shown in Fig. 9(a). The unit cell of the dual-band absorber

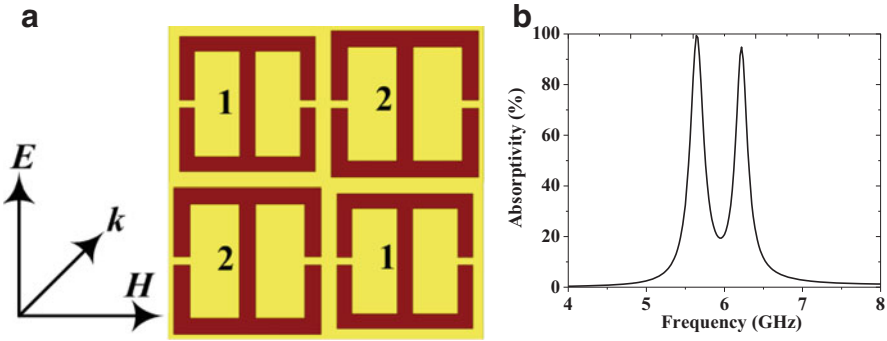

Fig. 9 (**a**) Design of the unit cell of a dual band metamaterial absorber and the (**b**) absorptivity response. (Reproduced from Bhattacharyya et al. (2014))

consists of sub-cell 1 and sub-cell 2 which are placed diagonally. The sub-cell 1 is scaled up by a factor of 1.2 whereas the sub-cell 2 is scaled up by a factor of 1.3. Due to different dimensions of the sub-cells 1 and 2, the structure exhibits absorption in the C band at two different frequencies, i.e., 5.64 GHz and 6.22 GHz, as shown in Fig. 9(b). The surface current orientation at the two frequencies is shown in Fig. 10. It can be seen that at both frequencies the surface currents are antiparallel to each other; thus, satisfying the condition of magnetic resonance in the structure.

The work done in Bhattacharyya et al. (2013) has presented a triple-band polarization independent ultra-thin metasurface absorber consisting of three concentric closed ring resonators (CRR) in the unit cell. The structure exhibits three absorption bands near to the middle of the FCC defined radar spectrum, i.e., at 5.5 GHz, 9.52 GHz, and 13.80 GHz with an absorptivity of 94.1%, 99.6%, and 99.4%, respectively. The top view of the unit cell along with the absorptivity response is shown in Figs. 11(a) and (b), respectively. The presence of three concentric rings of different radii is responsible for absorption at three different frequencies.

The direction of the surface currents at the top and bottom metallic plates and electric field distribution at the top metallic plate are shown in Fig. 12. It is well depicted from the circulating current curve that they are antiparallel to each other at all the three frequencies. The circulating current loops are created within the structure which is supported by the magnetic field, resulting in magnetic excitation. The interaction of electric field with the top metallic component excites the opposite charges at the two edges of the metal. This gives rise to electric resonance. The simultaneous occurrence of electric and magnetic resonances together creates strong absorption in the structure at three discrete frequencies.

The proposed CRR structure is fabricated with dimension of 250 mm $\times$ 250 mm $\times$ 1 mm. The prototype of the fabricated sample is shown in Fig. 13(a), and the enlarged portion is shown in Fig. 13(b). For the CRR pattern on the commercially available FR-4 substrate photolithography process has been performed to generate CRR pattern experimentally using printed circuit board (PCB) technology.

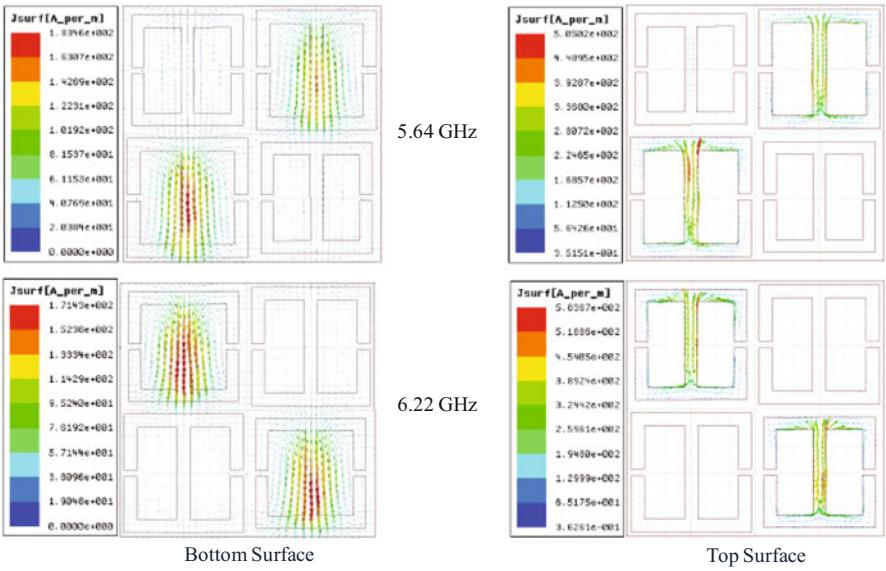

Fig. 10 Surface currents at the top and bottom metallic surfaces at 5.64 GHz and 6.22 GHz. (Reproduced from Bhattacharyya et al. (2014))

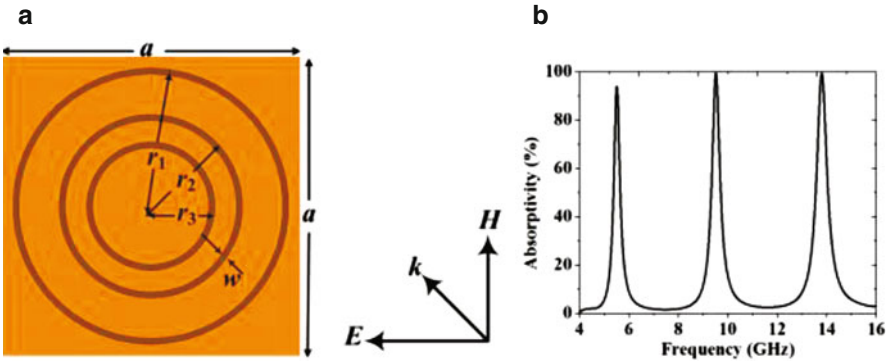

Fig. 11 (**a**) Top view of the concentric CRR structure and (**b**) absorptivity response of the triple-band metamaterial structure. (Reproduced from Bhattacharyya et al. (2013))

All the testing has been done in the anechoic chamber using a pair of identical horn antennas that are connected to a Vector Network Analyzer (VNA), as shown in Fig. 13(c). One of the antennas working as a transmitting antenna is used to launch the electromagnetic wave on the sample while the other one is used to receive the electromagnetic wave reflected from the sample. The response of the sample is measured and recorded in a VNA. The absorptivity response has been computed and shown in Fig. 13(d).

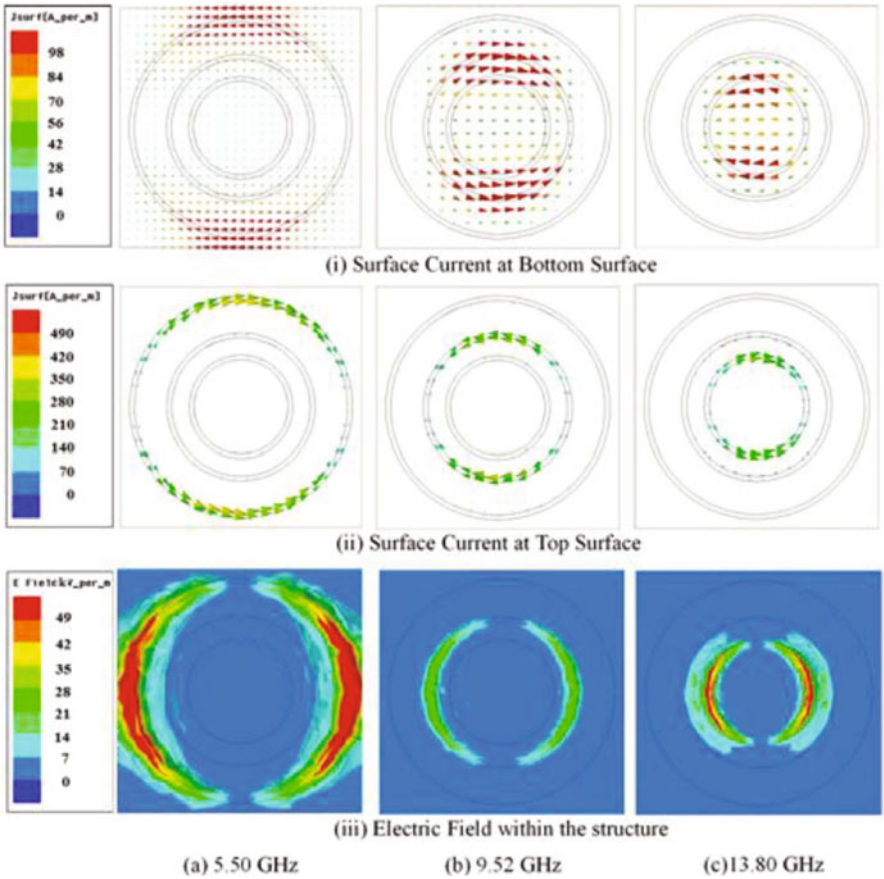

Fig. 12 Surface current distributions at the bottom and top surfaces along with the electric field distribution of the proposed structure at the frequencies of absorption (**a**) 5.50 GHz (**b**) 9.52 GHz and (**c**)13.80 GH. (Reproduced from Bhattacharyya et al. (2013))

The work done in Bhattacharya et al. (2015) has presented a penta-band polarization-insensitive metamaterial absorber for airborne radar applications. The proposed structure exhibits five absorption bands at 3.4 GHz, 8.34 GHz, 9.46 GHz, 14.4 GHz, and 16.62 GHz with an absorptivity of 98.6%, 96.6%, 90.1%, 97.8%, and 93.1%. The top view of the unit cell along with the absorptivity response is shown in Figs. 14(a) and (b), respectively. For all metallic components copper has been considered whereas, for the dielectric, FR-4 having dielectric constant of 4.25 and loss tangent of 0.02 has been chosen.

In order to check the robustness of the design, the absorber is examined under different polarization angles. Due to the fourfold symmetry the structure is examined under polarization angle variation up to 45°, for the case when the wave is incident normal to the structure. It is well depicted from Fig. 15(a) that due to the fourfold

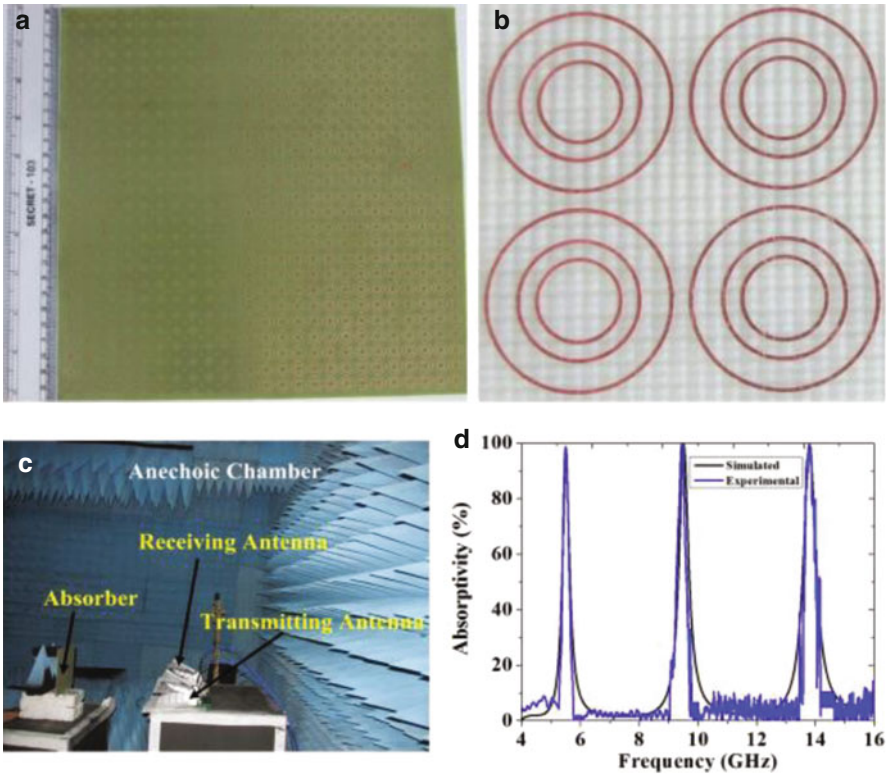

Fig. 13 (**a**) Prototype of the dual band absorber (**b**) enlarged view (**c**) experimental set up and (**d**) Comparison between simulated and experimental results. (Reproduced from Bhattacharyya et al. (2013))

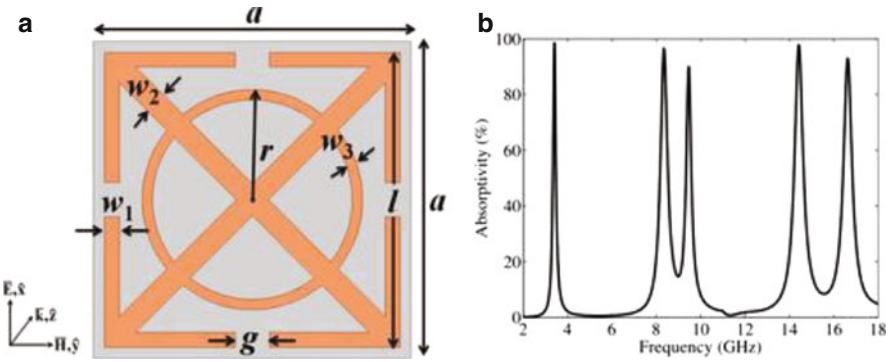

Fig. 14 (**a**) Top view of the unit cell of the penta-band absorber (**b**) absorptivity response of the structure. (Reproduced from Bhattacharya et al. (2015))

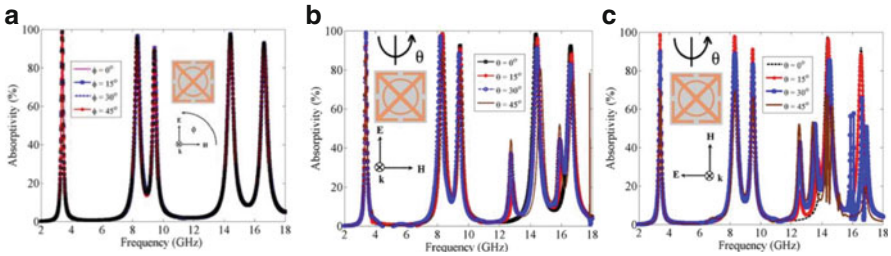

Fig. 15 Simulated absorptivity response for (**a**) different polarization angle under normal incidence and oblique incidence response for (**b**) TE and (**c**) TM polarization. (Reproduced from Bhattacharya et al. (2015))

symmetry the structure is polarization insensitive in nature. Also, the structure is studied under oblique incidences for both polarization (TE and TM) angle variation as shown in Fig. 15(b) and (c), respectively. In TE polarization the direction of electric field remains constant whereas the direction of magnetic field varies to make the wave incident obliquely on the top of the structure. In TM polarization, the direction of magnetic field remains constant whereas the electric field varies to make the wave incident obliquely on the top of the surface. It can be seen from Fig. 15(b)–(c) that all the five bands maintain a high absorptivity (>80%) up to 45° incident angle for both TE and TM polarizations incidence.

Broadband Metamaterial Absorbers

An optically transparent broadband absorber has been designed in Sheokand et al. (2017). The top layer of the structure consists of a resistive film of indium-tin-oxide and the bottom layer is a continuous resistive sheet. The top and bottom layers are separated by an air spacer. The top and side views of the structure are shown in Figs. 16(a) and (b), respectively. The ITO is uniformly deposited on a sheet of Polyethylene Terephthalate (PET) having a thickness of 0.175 mm. The top resistive film is in the shape of a slot-type cross dipole. Figure 16(c) shows that the structure exhibits a low reflection coefficient (below −10 dB) over a wide band of frequency ranging from 6.05 to 14.66 GHz, while the transmission coefficient is also low. The low reflection combined with the low transmission leads to high absorptivity value in the entire frequency range as observed from Fig. 16(d).

To explain the absorption mechanism of the structure, the surface current distribution on the top and bottom layers of the structure is studied at the peak absorption frequency, viz., 8.38 GHz, as shown in Figs. 17(a) and (b), respectively. It can be seen that at the peak frequency, the surface current distribution on the top and bottom layers are antiparallel to each other, signifying the formation of magnetic resonance in the structure. Besides, the current density is more on the top resistive surface, indicating that the top surface is the main contributor of broadband absorption.

To check the robustness of the structure, the designed absorber is also analyzed under oblique incidences as shown in Figs. 17(c) and (d), respectively. It can be seen that up to large variation in the polarization angle the structure can be used a

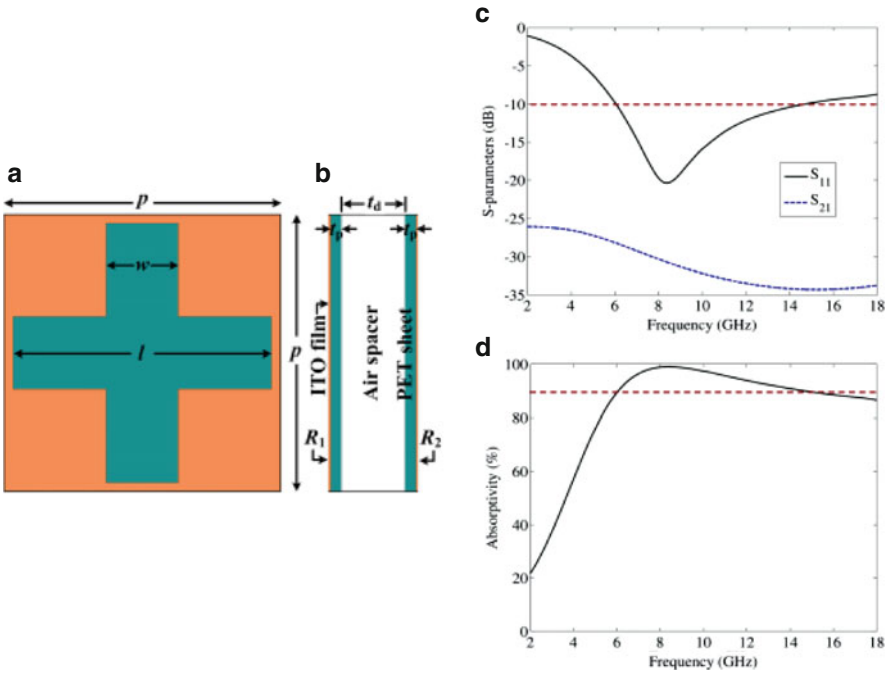

Fig. 16 (**a**) Top view and (**b**) side view of the unit-cell (**c**) Reflection and transmission coefficient response (**d**) Absorptivity response of the broadband absorber structure. (Reproduced from Sheokand et al. (2017))

metasurface absorber. Under TM polarization, the absorption band shifts to a higher frequency range for large incident angle, whereas for response gradually decreases for increasing angle of TE polarization.

Measurement Set-up of the Metasurface Absorber Structure

A dual-band metasurface structure has been fabricated with dimension 240 mm × 240 mm as illustrated in Fig. 18(a) along with the enlarged view in Fig. 18(b). The measurement of the reflection coefficient of the sample has been carried out in an anechoic chamber as depicted in Fig. 18(c). The sample has been designed on the top of the FR4 dielectric layer whereas the other side of the layer is completely metal-backed. The sample has been illuminated by a horn antenna while another identical horn antenna receives the reflection from the sample. The pair of the horn antennas has been connected to a Vector Network Analyzer (VNA).

Initially, the absorber sample has been removed and the background reflection has been measured as shown in Fig. 19(a). Thereafter, a larger metallic copper sheet with 1 m × 1 m dimension has been illuminated to measure the reflection from the sample. The large size of the sample ensures that the proper illumination of the sample by the electromagnetic beam launched from the antenna. Next, a sample identical to the dimension of the absorber prototype (240 mm × 240 mm) has been

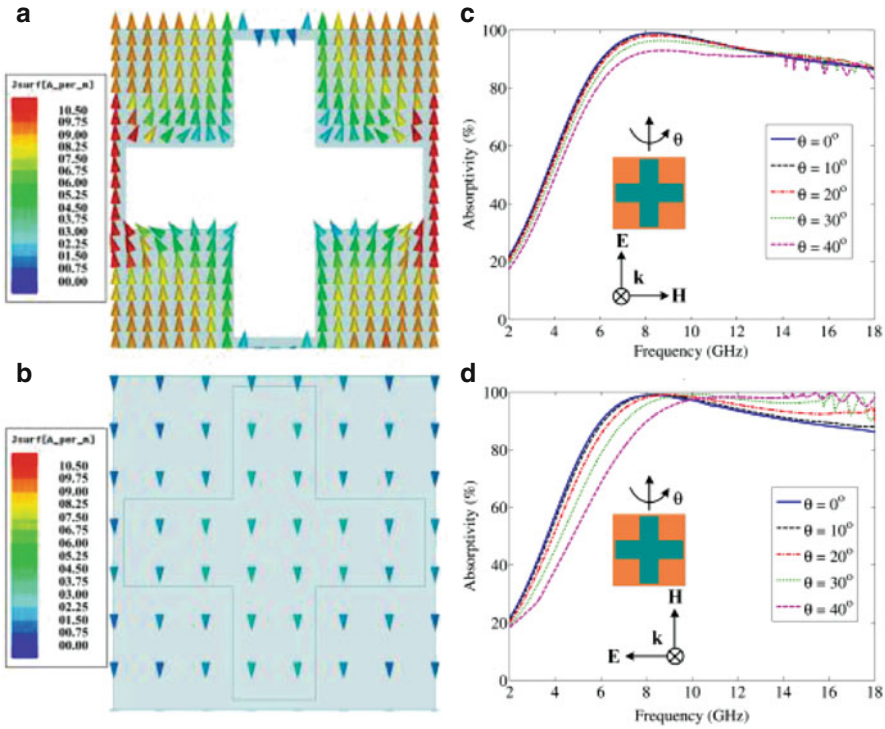

Fig. 17 Surface current distribution of the structure at (**a**) top and (**b**) bottom plates at 8.38 GHz. (**c**) Simulated absorptivity response of the structure under oblique incidence for TE polarization and (**b**) TM polarization. (Reproduced from Sheokand et al. (2017))

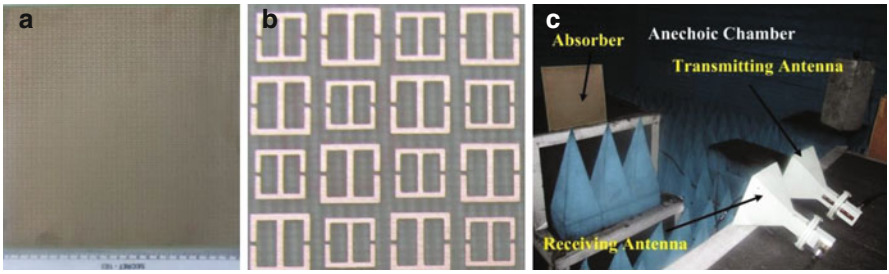

Fig. 18 (**a**) Prototype of the fabricated dual-band metasurface with (**b**) enlarged view and (**c**) experimental setup within the anechoic chamber. (Reproduced from Bhattacharyya et al. (2014))

illuminated by the electromagnetic wave launched. The comparison of the reflection coefficients from both the larger and smaller metallic sheets provide nearly identical responses as indicated in Fig. 19(a). Then, the absorber sample has been illuminated with the beam and the reflection coefficient of the sample has been measured. The two reflection coefficients have been compared as provided in Fig. 19(b). The

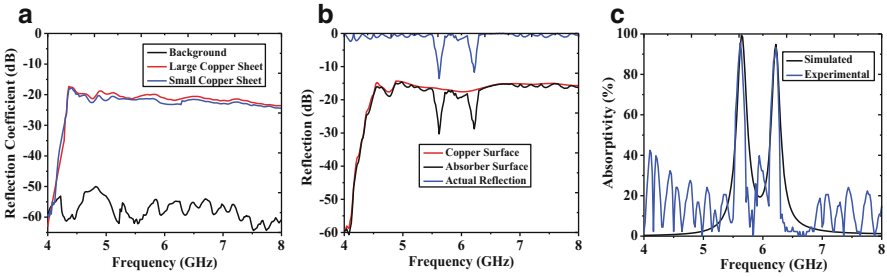

Fig. 19 (**a**) Reflection coefficient of the sample from a large and small copper sheet with background response; (**b**) reflection coefficient of the absorber sample and the copper sample for actual reflection from the absorber; and (**c**) comparison of simulated and measured absorptivity responses of the dual band absorber prototype. (Reproduced from Bhattacharyya et al. (2014))

difference between the two provides the actual reflection coefficient from the absorber sample. This, in turn, generates the absorptivity response as shown in Fig. 19(c). The measured absorptivity response has been compared with the simulated absorptivity response and consequently have been found to be in good agreement as seen from Fig. 19(c).

Recent Advancement in the Metamaterial Absorber Operating in the GHz Frequency Regime

Recently, in an era of artificial intelligence the design of a driver less cars are one of the most vibrant areas of research. In the design of such cars the most important part is the design of the radars. Automotive radars are used to detect the speed and range of the object running in the vicinity of the cars. Hence the most important sensors in the design of a driver-assistance system is the 24 GHz automotive radar which consists of a transmitter and a receiver. The transmitter sends the radio waves which when hit any object in the vicinity bounce back to the receiver. While designing such radars, the main conditions to be taken care are: (1) the metamaterial absorber must have a high absorptivity at 24 GHz and (2) the absorber should be polarization insensitive and must have a stable absorbance curve for a large angle of incidence. (3) In addition, the absorber must have a low profile, a flexible design, and can be fabricated as a sheet for proper installation.

A metamaterial-based absorber for automotive radar application has been designed in Tak et al. (2017). The absorber is polarization insensitive, flexible, low cost, while maintaining a high absorbance and wide-angle incidence. For the top metallic component an electric LC resonator has been taken. The absorber exhibits a high absorptivity of more than 97% at the frequency 24 GHz.

A flexible and conformable broad-band metamaterial absorber with wide angle and polarization stability for radar application has been proposed in Chen et al. (2018). The lumped resistor is mounted on the top of flexible substrate made up of silicone rubber and backed by a copper metal plate. The simulated result shows that the structure exhibits a large absorption of more than 90% over a bandwidth from

7.6 GHz to 18.3 GHz which covers both X band and Ku band. The main advantage of such design lies in the flexibility of the substrate which can be attached to any structure of different shapes without any break. The design structure has been simulated as well as fabricated for planar as well as conformable cases where the result shows that they are in good agreement with each other.

In Winson et al. (2019), a radar absorber consisting of non-periodic metallic patch coated with a layer of graphene has been discussed. The structure yields a good absorption of 99.93% over a broadband width of 1–60 GHz. Due to the limitation in the fabrication facility, a small unit cell of the structure has been fabricated and measured. The structure shows a good agreement between fabricated and measured results. To change the Fermi energy level of the graphene layer, the electric potential has to be changed. The structure can be used for the applications of stealth technology due to broadband absorption, low thickness, and low weight.

A tunable single-band metamaterial absorber has been presented in Ghosh and Srivastava (2016). The configuration is based on periodically arranged circular rings having PIN diodes. The PIN diodes are connected in such a way that by switching the diode ON to OFF, the absorption frequency can be shifted from S band to C band. The switching of the PIN diode has been done by controlling the bias voltage of the PIN diode. When the PIN diode is in ON state, the absorption frequency is in S band (3.41 GHz) which shifted to C band (5.35 GHz) when the PIN diode is in OFF state. Moreover, due to the fourfold symmetry of the absorber structure, the design is polarization insensitive as well as have stability for a wide angle of incidence.

Recently graphene-based metamaterial designs have attracted immense growing interest due to its excellent electrical, mechanical, thermal, and optical properties. A broad-band and tunable radar absorber based on graphene capacitor integrated with resistive frequency selective surface has been proposed in Song et al. (2020). A schematic model of the proposed broad-band and tunable absorber is shown in Fig. 20.

By applying an electrostatic bias, the effective resistance of the graphene sheet can be changed, through which the reflectivity from the structure can be changed. The simulative as well as the experimental result shows that the structure can tune its reflectivity ranging from 2.9 GHz to 15.9 GHz under normal incidence condition when the graphene resistance varies from 120 to 700 Ω/sq. The tuning range of the reflectivity is from -4.6 to -15.3 dB. The physical mechanism of the tunable absorber is discussed by studying the electric field distribution and power loss density. Additionally, the structure offers stability over wide angle of incidence and is polarization insensitive.

Recently, the work done in Choudhary et al. (2020) has proposed a dual-band metamaterial absorber for radar applications in S and X band. The design has used a closed ring resonator on a low-profile substrate. The absorber exhibits more than 90% absorptivity at the frequency 3 GHz and 9 GHz. It has been proposed that the absorber can be used to improve the performance of a radar system for short range applications when positioned between transmitter and receiver antennas. Through which the error in estimated target detection can be reduced and clutter reduction can be improved.

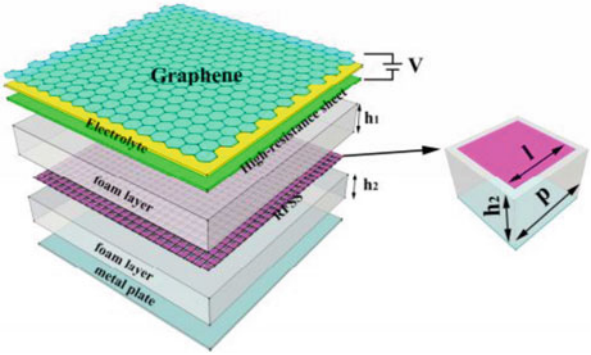

Fig. 20 Design of a broadband and tunable graphene base metamaterial absorber. (Reproduced from Song et al. (2020))

Metasurface Absorbers Operating in the THz and IR Frequency Regimes

It has been demonstrated that metamaterial has potential application in THz domain and various optoelectronics devices in this frequency range have been reported including an absorber. Metamaterial absorbers can work efficiently over a large spectral range extending from THz to Visible frequency range by scaling down the dimension and by utilizing nano-micro fabrication techniques. In this chapter we define THz as the frequency range ranging from 1 THz to 10 THz, also known as THz-Gap. The frequency range above it is termed as infrared frequency having various sub-divisions such as Far infrared, Long-wavelength infrared, Mid-wavelength infrared.

Metamaterial Absorbers Operating in THz Frequency Range

The first metamaterial device at terahertz was introduced by Yen et.al (Yen et al. 2004) who experimentally demonstrated THz metamaterial using a planar layer of double SRR layer as shown in Fig. 21.

Recently, efforts have been made on the design of broadband metamaterial absorbers operating in the THz region. This can be done by either combining two identical geometries of different dimensions or by combining two different geometries in a single unit cell. Also, it can be done by adapting multi-layer structures for multi-resonant responses.

In Shen et al. (2012), they designed a triband metasurface absorber with three different peaks at 0.50 THz, 1.03 THz, and 1.71 THz, while exhibiting an absorption of 96.4%, 96.3%, and 96.7%, respectively. The absorption mechanism in the structure has been explained based on the multiple-reflection interference theory. Also, it is found that the structure is insensitive to the variation of incident angle up to 40°.

A dual-band metamaterial absorber consisting of two identical square patch on top of a dielectric layer and a continuous metallic plate has been demonstrated in

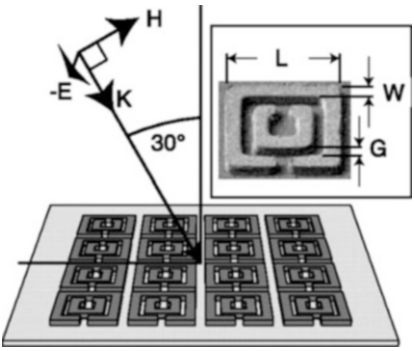

Fig. 21 Magnetically-coupled SRRs with different geometries excited by a 30° off-normal incident with an ellipsometer. (Reproduced from Yen et al. (2004))

Wang et al. (2020). The structure exhibits two peaks labelled as A and B with near unit absorption (100% absorption) are obtained, where peak A is due to the localized resonance of the two square patches having a line width of 0.2571 THz and a quality factor of 6.9156 whereas peak B results from the hybrid coupling of the localized resonance of the two square patches and surface lattice resonance has a narrow line-width of 0.0083 THz and large quality factor of 296.2771. Narrow line-width and large quality factor have important prospects in sensing application.

A graphene-based metasurface absorber generating a wide absorption band in the THz gap has been reported in Ghosh et al. (2020). A layer of graphene is applied on both the surface of silicon dioxide (SiO_2) substrate to generate a wide absorption band within the entire THz gap. The top graphene pattern acts as a matching layer between air and SiO_2 interface under a certain chemical potential value (μ). The bottom of the SiO_2 substrate is covered entirely with the graphene layer with a very high chemical potential value such that the entire layer acts a metallic reflector. The top view, perspective view, and side view of the structure is shown in Fig. 22(a). The whole unit cell is a three-layered structure, where the graphene layer has a modified-fractal type pattern on top of a 2.5 μm thick amorphous silicon dioxide (SiO_2) substrate ($\varepsilon_r = 3.9$, $\tan\partial = 0.0006$). The absorptivity response of the structure is shown in Fig. 22(d) whether it can be seen that the structure maintains a high absorptivity response covering the entire THz band gap. The work also compares the absorptivity response when the bottom layer is gold to the response when the bottom layer is graphene, and it can be seen from Fig. 22(e) that the presence of graphene as a metal improves the overall absorptivity response of the whole structure.

Metamaterial Absorbers Operating in the Infrared (IR) Frequency Range

A long wavelength infrared metamaterial absorber presented in Nilotpal et al. (2018) uses an electric field driven LC resonator. The structure exhibits a high absorption value of 96.2% at 30 THz. The structure is ultrathin (~λ/30) in nature and maintains a

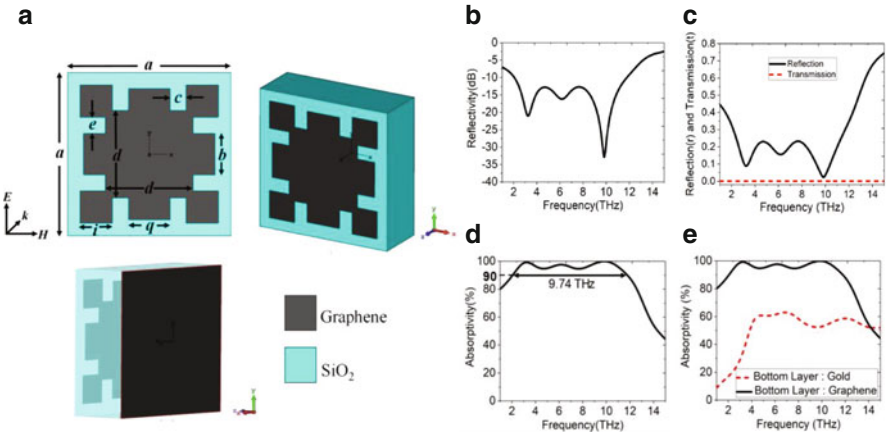

Fig. 22 (**a**) Top view, 3-D perspective and back view of the wideband graphene-based metasurface absorber in lower MIR region ($a = 6.7$ μm, $b = 1.8$ μm, $d = 3.8$ μm, $c = 0.9$ μm, $e = 0.9$ μm, $i = 1.4$ μm, $q = 1.8$ μm); frequency responses of (**b**) reflectivity (dB), (**c**) transmission (t) co-efficient, (**d**) absorptivity (%), and (**e**) comparison of absorptivity between bottom gold backed metasurface and bottom graphene backed metasurface. (Reproduced from Ghosh et al. (2020))

high absorptivity response for a wide angle of incidence under both TE and TM polarizations. The simulation is done in HFSS simulation software which is further verified in COMSOL electromagnetic solver. The top view of the metamaterial absorber along with the absorption curve has been shown in Figs. 23(a) and (b) respectively. For the dielectric ZnSe has been taken whereas for the metallic component gold has been taken. The dielectric property of ZnSe in the frequency of interest has been taken from Hawkins and Hunneman (2004), whereas the conductivity of metal has been modelled using the Drude model (Ordal et al. 1983). The thickness of the metallic component is 0.035 μm, which is slightly greater than the skin depth at the frequency of absorption (10 μm) (Qiang et al. 2004). The geometrical dimensions of the unit cell, as shown in Fig. 23(a), are optimized as $a = 1.5\ \mu m$, $b = 1.2\ \mu m$, $c = 0.1\ \mu m$, $d = 0.1\ \mu m$, $e = 0.2\ \mu m$.

A quad-band ultrathin polarization insensitive metamaterial absorber for infrared applications have been reported in Nilotpal and Chakrabarti (2019). The top and side view of the unit cell of the quad-band absorber along with the direction vector of electric field, magnetic field, and wave vector is shown in Fig. 24. The unit cell of the quad-band absorber consists of dipole of unequal length. The structure exhibits absorptions at 18.4 THz, 33.4 THz, 51.4 THz, and 75.2 THz with respective absorptivity of 99.2%, 99.4%, 92.2%, and 98.2%, as shown in Fig. 24(b). The structure is ultrathin $\frac{\lambda}{33}$ in nature and also compact in periodicity $\frac{\lambda}{8}$. The permittivity of gold has been modelled using the drude model ($\varepsilon(\omega) = 1 - \frac{{\omega_p}^2}{\omega(\omega+i\gamma)}$), where the plasma frequency (ω_p)and damping frequencies (γ) are 2176 THz and 5.9 THz, respectively (at the frequency of absorption).

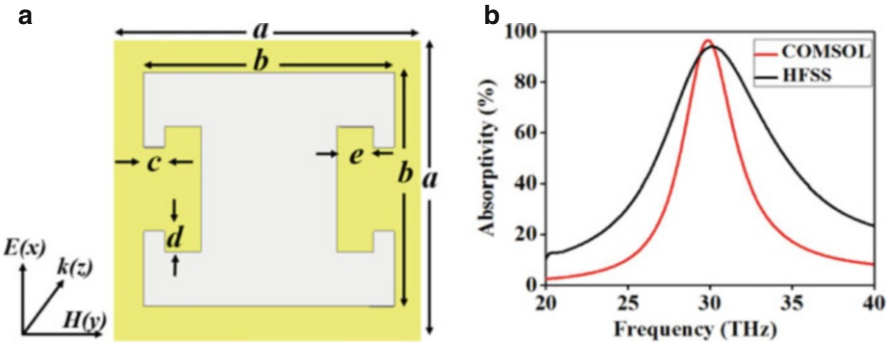

Fig. 23 Schematic of the (**a**) top view of the metamaterial absorber and (**b**) absorptivity response. (Reproduced from Nilotpal et al. (2018))

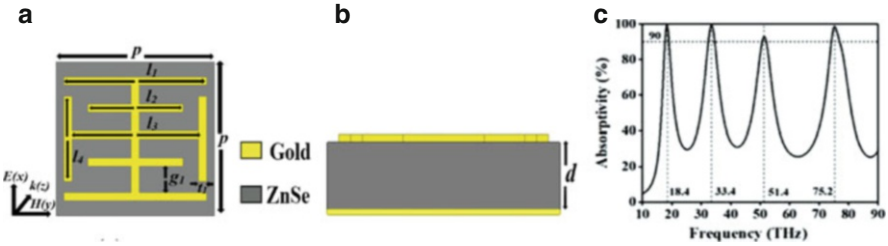

Fig. 24 Schematic of the (**a**) top view (**b**) side view and (**c**) absorptivity response of the quad band metamaterial absorber. (Reproduced from Nilotpal and Chakrabarti (2019))

The absorption mechanism of the absorber can be understood by the curve of electric and magnetic fields at the quad frequency bands as shown in Figs. 25(a) and (b), respectively. The opposite distribution of electric and magnetic fields at each dipole signifies the accumulation of opposite charges which corresponds to the formation of electric and magnetic resonances. Also, from the surface current distribution at the top and bottom metallic plates as shown in Figs. 25(c) and (d), respectively, it can be seen that the formation of antiparallel surface current results in the formation of magnetic dipoles which strongly interact with the incident magnetic field to give rise to magnetic resonance. The interaction of electric field with the top metallic component give rise to electric resonance. The simultaneous occurrence of these two resonances give rise to absorption in the structure. From the distribution of electromagnetic field and surface currents at the top and bottom metallic surfaces, it can be observed that each absorbing band is obtained due to different section of dipoles.

An ultra-thin $\frac{\lambda}{24}$ broad-band metamaterial absorber using array of split ring resonators and crossed slot structure is demonstrated in (Nilotpal and Chakrabarti 2018). The top view of the metasurface along with the electromagnetic field direction is shown in Fig. 26. The top metasurface layer consists of three split rings having unequal length along with a crossed shaped slot, as shown in Fig. 26(a). The structure exhibits more than 90% absorptivity in the frequency range 42 THz to

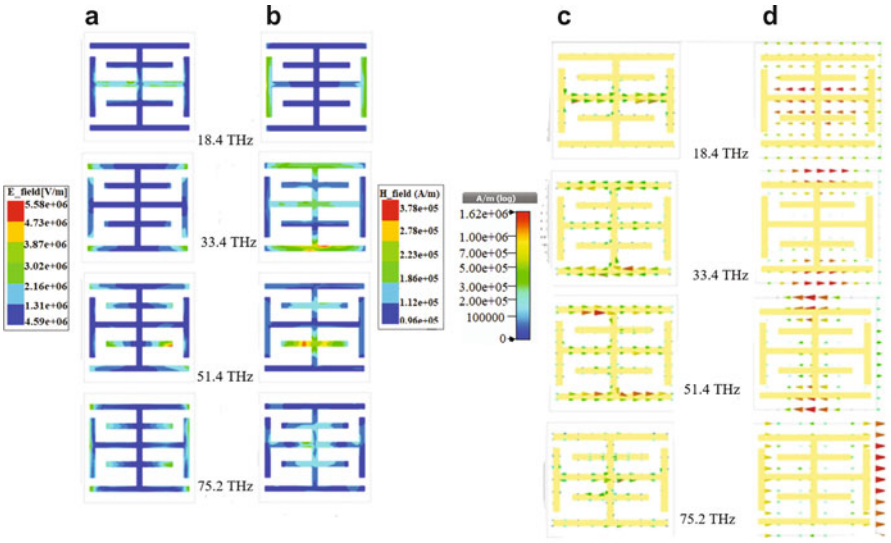

Fig. 25 (**a**) Electric field, (**b**) magnetic field distribution within the structure, (**c**) surface current orientation at the top, and (**b**) bottom metallic plate at 18.4 THz, 33.4 THz, 51.4 THz and 75.2 THz. (Reproduced from (Nilotpal and Chakrabarti (2019))

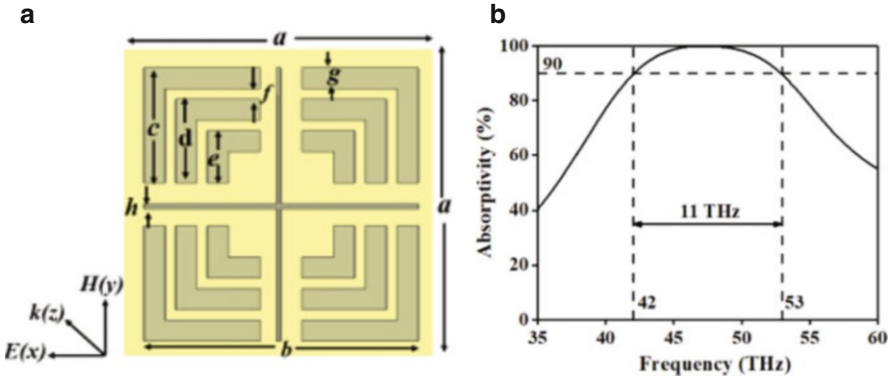

Fig. 26 Schematic of the (**a**) top view and (**b**) absorptivity response of the broad-band metamaterial absorber. (Reproduced from (Nilotpal and Chakrabarti 2018))

53 THz. It is observed from Fig. 26(b) that the structure exhibits more than 90% absorption between the frequency range 42 THz to 53 THz and a maximum absorption of 99.5% between frequency range 46 THz to 47.6 THz. The fourfold design makes the structure polarization insensitive. The three split layers are so dimensionally optimized that their resonance frequencies are in very close proximity to each other. Therefore, the structure shows a broad-band response.

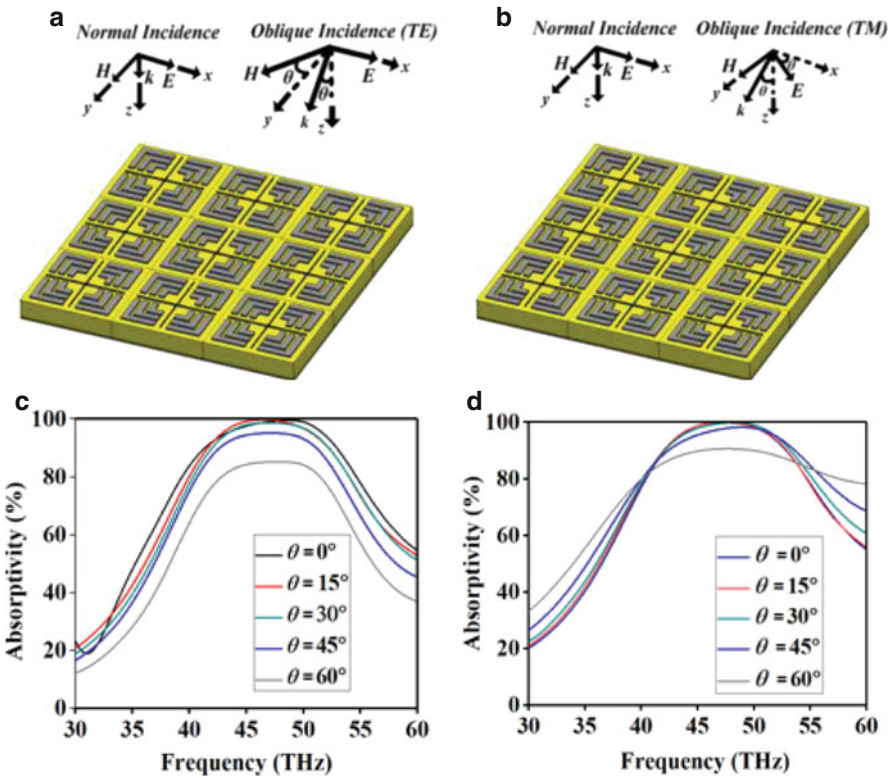

Fig. 27 Setup of the oblique incidence response under (**a**) TE and (**b**) TM polarization and the simulated absorptivity response of the structure under (**a**) TE and (**b**) TM polarization. (Reproduced from (Nilotpal and Chakrabarti 2018))

The structure is studied under oblique incidences for TE as well as TM polarization. The setup of the proposed structure under TE polarization is shown in Fig. 27(a) whereas the setup of the proposed under TM polarization is shown in Fig. 27(b). The absorptivity response of the structure under TE polarized incidence is shown in Fig. 27(c) and the absorptivity response of the structure under TM polarized incidence is shown in Fig. 27(d). From both responses, it can be seen that the structure maintains a high absorptivity value up to 60° incidence. The structure is also studied for polarization angle variation under normal incidence, and it has been found that the structure is polarization insensitive owing to its fourfold symmetry as shown in Fig. 28.

Absorption Based on Multiple Reflections and Interference

The theory of absorption which evolves initially was based on single effective medium theory where the whole metamaterial structure was considered as a single entity having an effective medium parameter, i.e., ε and μ. However, the

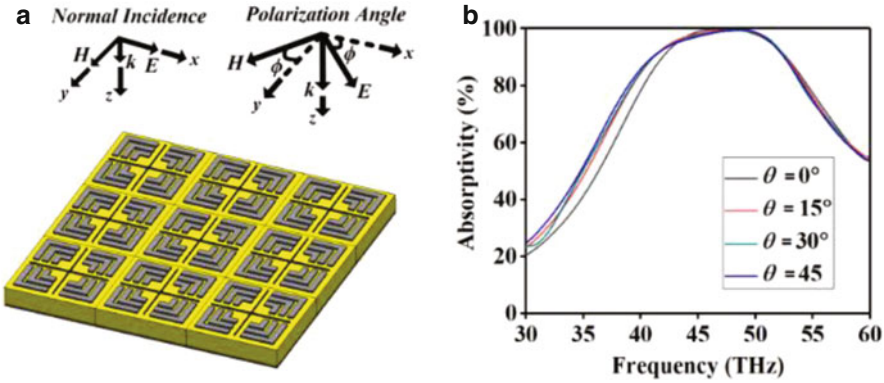

Fig. 28 (**a**) Set up and (**b**) absorptivity response for different polarization angles under normal incidence. (Reproduced from (Nilotpal and Chakrabarti 2018))

work presented in Nilotpal et al. (2020a) has demonstrated that the theory based on single effective medium theory was questionable when instead of illuminating the structure from front, the structure is illuminated from the bottom. In this case, there is complete reflection instead of absorption, despite the fact that the characteristics impedance of the whole structure remains the same. This asymmetry in the absorption cannot be explained by single effective medium theory. Further, it was found that when the thickness of the dielectric layer was increased by an integral multiple of half-wavelength then the absorption characteristics of the metamaterial absorber repeats. Ambiguity between orientation of surface current and absorption was found when the increment was for an odd multiple of half wavelength. In this case, absorption was found despite the fact that the surface currents were parallel to each other. This totally contradicts the theory of single effective medium theory where the presence of antiparallel surface current was compulsory. Hence the theory of absorption based on multiple reflection interference evolved later. In this theory, absorption mechanism is based on three-layer model which treats all the three layers separately. The overall reflection and absorption from the structure is the multiplication of reflection due to each layer.

Validation of Theoretical Model

In order to explain this theory a narrowband metamaterial absorber has been shown in Fig. 29. Though all the calculation is done in the terahertz range, the same theory is equally applicable for GHz range also. An initial work (in GHz frequency) in this direction has also been done in Shen et al. (2012); Wanghuang et al. (2013); Ghosh et al. (2015).

The top and side view of the metamaterial absorber along with the absorption curve has been shown in Fig. 29. For the dielectric material, ZnSe has been taken whereas for the metallic component gold has been taken. The geometrical

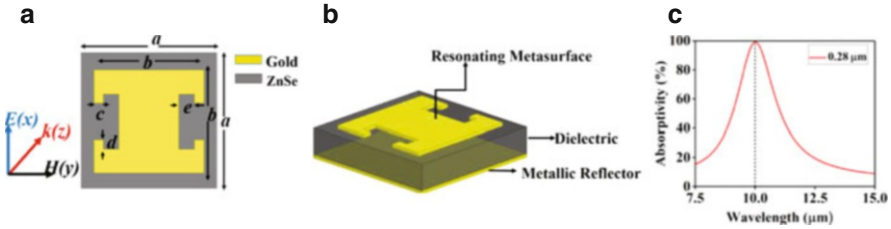

Fig. 29 (**a**) Top view (**b**) side view and (**c**) absorptivity response of the metamaterial absorber. (Reproduced from Nilotpal et al. (2020a))

dimensions of the meta-atom, as shown in Fig. 29(a) are optimized as $a = 1.7\ \mu m$, $b = 1.4\ \mu m$, $c = 0.1\ \mu m$, $d = 0.1\ \mu m$, $e = 0.2\ \mu m$.

In order to understand the multiple reflections inside the structure a schematic showing multiple reflections and transmissions at each layer of the metamaterial absorber is shown in Fig. 30(a).

When the wave is incident on the top metallic layer, a part of the wave is reflected back whereas a part of the wave enters into the dielectric and then undergoes multiple reflections inside. The overall reflection from the structure is the multiplication of reflections due to each layer, as given in Eq. 6 (Jeon et al. 2016).

$$r_{structure} = \frac{R_{12} + \tau R_{23} e^{2/\beta}}{1 - R_{21} R_{23} e^{2i\beta}} \tag{6}$$

Where R_{12}, R_{21} and T_{12}, T_{21} and are the reflection and transmission coefficients of the top metallic layer when the wave is incident from the top side (i.e., air) and bottom side (i.e., dielectric), respectively. R3 is the reflection coefficient of the bottom metallic layer. τ is equal to $T_{21}T_{12} - R_{12}R_{21}$ and $\beta = \eta \times k \times t_d$ is the total phase change of the wave inside the dielectric cavity where η, k, and t_d are the refractive index, propagation vector, and dielectric thickness of the metamaterial structure. In order to get a zero reflection, both amplitude as well as phase condition of Eqs. 7 and 8 has to be satisfied simultaneously.

$$R_{12} = \mid \tau R_{23} \mid \tag{7}$$

$$\theta = \phi(R_{12}) - \phi(R_{23}) - \phi(\tau) - 2\beta = (2n+1)\pi, \quad \text{where} \quad \mid n \mid = 0, 1, 2 \ldots \tag{8}$$

Here θ *is the phase difference between the coefficients written in the numerator part of Eq. 2. In Eq. 2,* $\phi(R_{12})$ *is the phase of the reflection coefficient* R_{12}*,* $\phi(\tau)$ *is the phase associated with the constant* τ*,* $\phi(R_{23})$ *is the phase of the reflection coefficient* R_{23} *and* 2β *is the total phase change experienced by the wave when it propagates inside the cavity.* It can be seen from Figs. 30(b) and (c) that both these equations are satisfied at the frequency of absorption (10 μm). This signifies that all the wave which incident on the top metallic layer enters into the middle layer without any reflection. Now, the wave which has entered into the dielectric layer reflects multiple

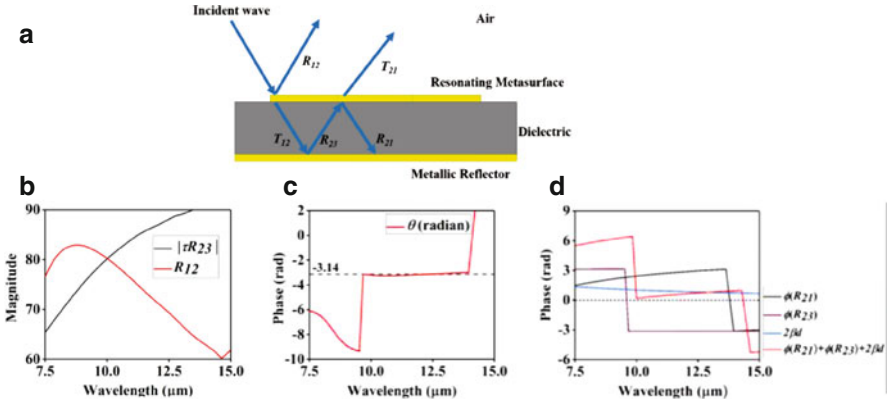

Fig. 30 (**a**) Multiple reflections and transmissions at each layer, (**b**) amplitude condition, (**c**) phase condition necessary for zero reflection, and (**d**) Fabry Perot resonance for absorption at 10 μm. (Reproduced from Nilotpal et al. (2020a))

times between the top and bottom metallic layer. If the multiple reflected waves satisfy Eq. 9, then the waves will be in phase with each other and thus they interfere constructively which led to the formation of Fabry Perot resonance responsible for absorption in the structure. It can be seen from Fig. 30(d) that the condition of Fabry Perot resonance is satisfied at the absorption frequency.

$$\varphi = \phi(R_{21}) + \phi(R_{23}) + 2\beta = 2n\pi, \quad |n| = 0, 1, 2.. \tag{9}$$

Proper analysis of Eqs. (8) and (9) suggests that both the phase conditions repeat when there is an increment of $2n\pi$ in β. This is equivalent to say that the condition will repeat when there is an increment of $n\frac{\lambda_d}{2}$ in the cavity thickness, where λ_d corresponds to the wavelength in the dielectric. An increment of n decides the order of the cavity and also order of Fabry Perot resonance mode. In order to validate this, the structure is simulated for first three order of cavity thickness, i.e., the 1st ($t_{d1} = 0.28$μm), 2nd ($t_{d2} = 0.28 + \frac{\lambda_d}{2} = 1.99$ μm), and 3rd ($t_{d3} = 0.28 + \lambda_d = 3.71$ μm). It is found that the absorption characteristics repeat itself with an increment of $n\frac{\lambda_d}{2}$ in the cavity thickness as shown in Fig. 31. A slight shift in the resonance frequency can be seen in the second and third order of cavity. This might be due to the change in the distance between top and bottom metallic plate.

The surface current distribution for the first three order of cavity thickness is evaluated and shown in Fig. 32. It can be seen that for the first and third order of cavity thickness the surface currents are antiparallel to each other. However, for the second order of cavity thickness the surface current is parallel to each other. Despite having parallel surface current, the second order of cavity exhibits unity absorption just like first and third order of cavity thickness.

In order to elucidate the reason, the electric field for the first three order of the dielectric cavity has been studied in Fig. 33. The electric field has been

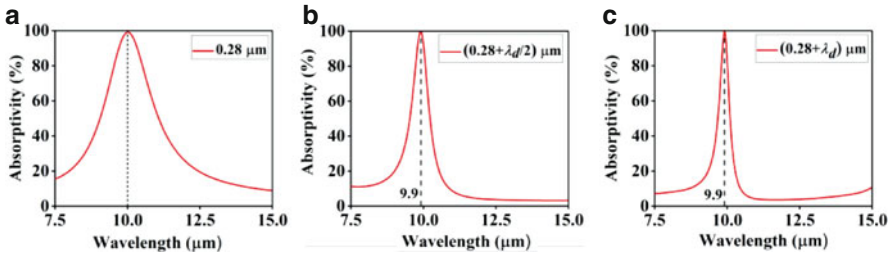

Fig. 31 Absorptivity response for the first three order of cavity thickness (**a**) $t_{d1} = 0.28$ μm (**b**) $t_{d2} = 1.99$ μm (**c**) $t_{d3} = 3.71$ μm. (Reproduced from Nilotpal et al. (2020a))

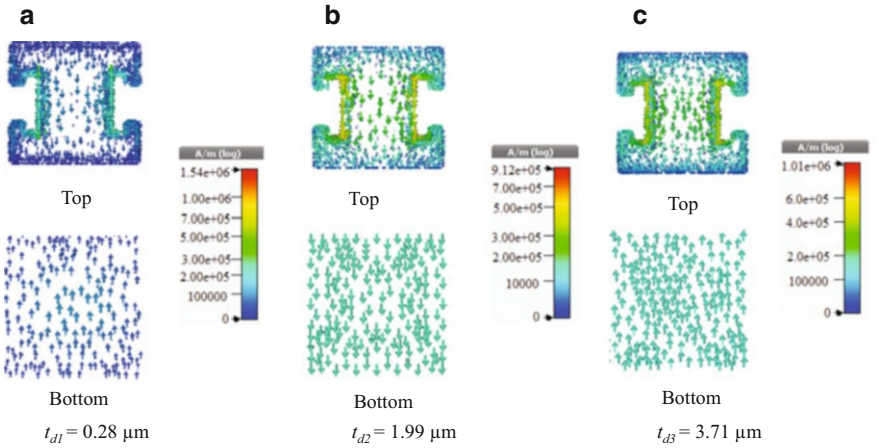

Fig. 32 Surface current distribution for the dielectric thickness (**a**) $t_{d1} = 0.28$ μm (**b**) $t_{d2} = 1.99$ μm (**c**) $t_{d3} = 3.71$ μm. (Reproduced from Nilotpal et al. (2020a))

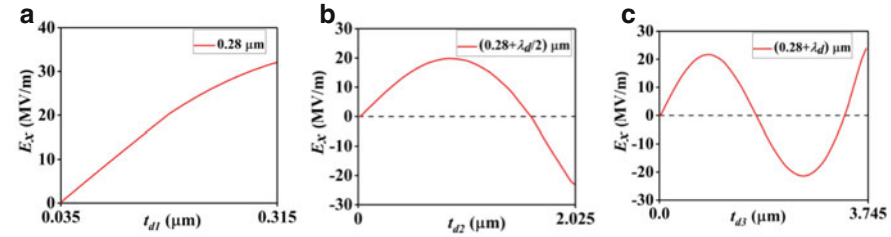

Fig. 33 Real part of E_x component along z direction for the first three order of cavity thickness (**a**) $t_{d1} = 0.28$ μm (**b**) $t_{d2} = 1.99$ μm (**c**) $t_{d3} = 3.71$ μm. (Reproduced from Nilotpal et al. (2020a))

evaluated from $z = 0\ \mu m$, i.e., bottom metallic reflector to $z = (0.035 + t_{dn})\ \mu m$, i.e., the top metallic layer for all the three order of cavity thickness as shown in Fig. 33.

Reason for Parallel and Antiparallel Surface Currents

The direction of the surface current at the top and bottom metallic plates are decided by the phase of the electric field at these two boundaries and also depend on the sign of effective dielectric permittivity, according to Eqs. 10 and 11, respectively. Directions of the surface currents at the top (J_{x-top}) and bottom metallic layers ($J_{x-bottom}$) are decided by the phase of the tangential electric field component E_x at the two layers and the sign of the effective permittivity of the top metasurface (ε_{top}) and bottom metasurface (ε_{bottom})

$$J_{x,top} = j\omega\varepsilon_{top}E_x \tag{10}$$

$$J_{x,bottom} = j\omega\varepsilon_{bottom}E_x \tag{11}$$

The effective permittivity value of the bottom layer is always negative, according to the drude model and the dielectric permittivity of the top metallic plate is always positive. Hence, for the occurrence of antiparallel surface current the electric field at the two extreme boundaries should be in phase, whereas for parallel surface currents the electric field at the two boundaries should be out of phase. It is well evident from Figs. 33(a) and (c) that for the first and third order of cavity thickness, electric field at the two boundaries are in phase and thus giving rise to antiparallel surface currents whereas the electric field for second order of cavity are out of phase as shown in Fig. 33(b), and thus corresponds to parallel surface currents.

Validation of Multiple Resonance from Time Domain Analysis

In this section, the multiple reflection phenomenon has been verified for normal as well as oblique incidence from time domain aspect. In order to do so, the S_{11} parameter from the frequency domain has been transformed to time domain using inverse chirp-Z transform technique (Frickey 2014). For the time domain analysis, the third order of cavity has been taken. The multiple reflection phenomenon under normal incidence condition is shown in Fig. 34(a). The distance between the incident port and top metallic layer is $h = 12\ \mu m$, so the incident wave is planar in nature. The first wave reflected back to port 1 is termed as R_{12}, whereas the second and third wave is termed as R_{13} and R_{14}, respectively. The time domain conversion of the S_{11} parameter for normal incidence is shown in Fig. 34(b). The time taken by each of the wave to received back to the port is shown in Fig. 34(b). The first peak obtained at 0.08 ps corresponds to the time occurrence of first reflected wave, i.e., R_{12}. Similarly, the peaks obtained at 0.141 ps and 0.20 ps correspond to the time occurrence of second and third reflected waves, i.e., R_{13} and R_{14} respectively.

The results obtained by the time domain analysis has also been validated numerically through equations numbered from 12, 13, 14. Here h is the distance from the incident port to the top of the metaurface structure, c is the speed of light in vacuum, and t_d is the dielectric thickness.

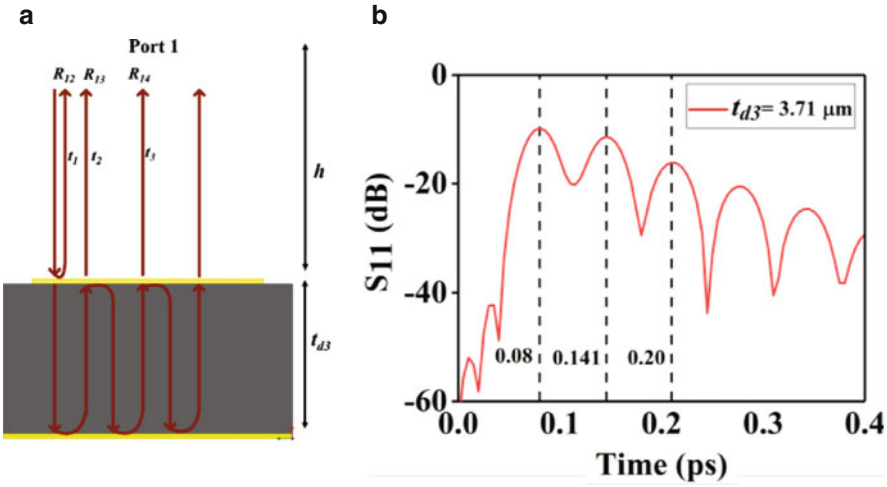

Fig. 34 Multiple reflection phenomenon inside the dielectric layer of the metamaterial absorber for normal incidence (b) time-domain conversion of the reflection coefficient of the metamaterial absorber for normal incidence. (Reproduced from Nilotpal et al. (2020a))

$$t_1 = \frac{2h}{c} \tag{12}$$

$$t_2 = t_1 + \frac{2t_{d3}\sqrt{\varepsilon_r}}{c} \tag{13}$$

$$t_3 = t_2 + \frac{4t_{d3}\sqrt{\varepsilon_r}}{c} \tag{14}$$

For $h = 12$ μm and $t_{d3} = 3.71$ μm, the time occurrences of first three reflected have been analytically obtained as $t_1 = 0.08$ ps, $t_2 = 0.138$ ps, and $t_3 = 0.196$ ps, respectively. Comparing this with the time obtained in Fig. 34(b), it has been found that the they are in well match with each other. Apart from normal incidence this analysis has also been done for oblique incidence as shown in Fig. 35(a). The result obtained is shown in Fig. 35(b). The occurrence of peak changes due to the change in the path length travelled by the wave. If a wave is incident at an angle θ_i then the path travelled by the electromagnetic wave will be modified to h sec θ_i, instead of h. Similarly, the path travelled by the wave inside the cavity will be modified to t_{d3} sec θ_t instead of t_{d3}. The occurrences of first three peaks at time t_1', t_2', and t_3' can be calculated from Eqs. (15, 16, and 17), respectively.

$$t_1' = \frac{2h \sec\theta_i}{c} \tag{15}$$

$$t_2' = t_1' + \frac{2t_{d3} \sec\theta_t \sqrt{\varepsilon_r}}{c} \tag{16}$$

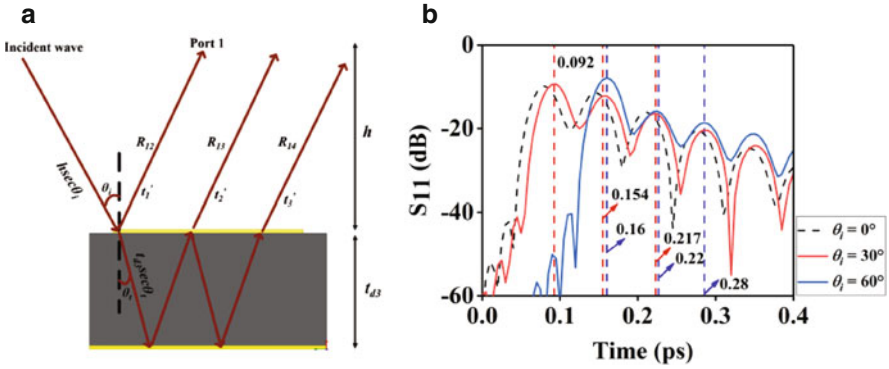

Fig. 35 (**a**) Multiple reflection phenomenon inside the dielectric layer of the metamaterial absorber for oblique incidence (**b**) time-domain conversion of the reflection coefficient peaks of metamaterial absorber for oblique incidences of 30° and 60°. (Reproduced from Nilotpal et al. (2020a))

$$t_3' = t_2' + \frac{4t_{d3}\sec\theta_t\sqrt{\varepsilon_r}}{c} \tag{17}$$

The S_{11} parameter for both the incidences i.e., 30° and 60° has been converted into time domain, as shown in Fig. 35(b). The time of occurrence of first three reflected wave for 30° incidence (at port 1) is 0.092 ps, 0.1574 ps and 0.217 ps respectively, whereas, for 60° the time of occurrence is 0.16 ps, 0.22 ps and 0.28 ps respectively. The time occurrence data obtained numerically using Eqs. 15, 16, and 17 are in well match with the data obtained analytically, as shown in Fig. 35(b).

Mathematical Interpretation of Metamaterial Absorber

The work presented in (Nilotpal et al. 2020b) has discussed the mathematical interpretation of wave propagation and absorption in a metamaterial absorber. The work is completely on the basis of multiple reflection interference phenomenon as discussed in the previous section. The wave propagation inside the dielectric cavity has been explained on the basis of electric and magnetic fields. The effect of change in the polarization angle on the standing wave formation and surface current orientation has also been discussed separately.

In order to describe the theory, a schematic showing the propagation of wave in the form of electric and magnetic field components are shown in Fig. 36. All the mathematical interpretation has been done considering the fact that a transverse electric (TE) electromagnetic field is incident on the top of the structure. However, the theory is equally applicable for a Transverse magnetic (TM) polarized wave also, the only change required will be in the direction vectors of electric and magnetic field components.

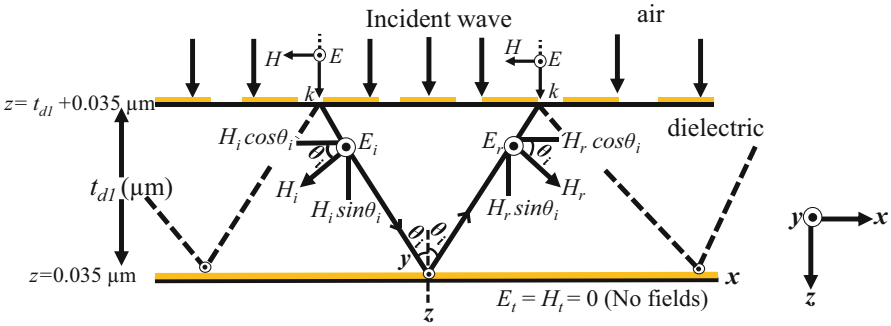

Fig. 36 The propagation of electric and magnetic fields inside the dielectric layer of the metamaterial absorber. (Reproduced from Nilotpal et al. (2020b))

Electric and Magnetic Fields Inside the Dielectric Cavity

Consider a TE polarized wave is incident on the top of the metallic plate of the metamaterial absorber. The presence of gaps in the structure diffracts the wave obliquely inside the dielectric cavity (Peng et al. 2012). Let the wave propagates in $+z$ direction and strikes the bottom metallic plate at an angle θ_i. Let the electric field is oriented in $+y$ direction, then the orientation of magnetic field can be decomposed into two components which are in x and z direction. The expression of the total electric and magnetic field incident inside the dielectric cavity can be derived as Eq. 18 and Eq. 19, respectively.

$$\overline{E}_i = E_{io} i^{-j}(\beta z \cos\theta_i + \beta x \sin\theta_i + \pi)\hat{y} \tag{18}$$

$$\overline{H}_i = \frac{E_{io}}{\eta}(\sin\theta_i \hat{z} - \cos\theta_i \hat{x}) e^{-j}(\beta z \cos\theta_i + \beta x \sin\theta_i + \pi) \tag{19}$$

Here, β and η represents the propagation constant and intrinsic impedance of the dielectric. An additional phase of π in the equation of incident electric field is to compensate the reflection from the bottom metallic layer. Accommodating the proper sign change for the reflected wave, the total electric and magnetic field for the reflected wave can be written as Eqs. 20 and 21. respectively.

$$\overline{E}_r = E_{ro} e^{-j}(-\beta z \cos\theta_i + \beta x \sin\theta_i)\hat{y} \tag{20}$$

$$\overline{H}_r = \frac{E_{ro}}{\eta}(\sin\theta_i \hat{z} + \cos\theta_i \hat{x}) e^{-j}(-\beta z \cos\theta_i + \beta x \sin\theta_i) \tag{21}$$

By applying maxwell's equation, the tangential component of electric field oriented along +y direction vanishes at the media interface due to the presence of metallic component at the boundary. By applying maxwell's boundary condition, it

can be found that $E_i = E_r$. Therefore, the net electric field at any point inside the dielectric cavity is given as Eq. 22.

$$E_y = -2jE_{io}e^{-j\beta x \sin\theta_i} \sin(\beta z \cos\theta_i) \tag{22}$$

Similarly, the net magnetic field inside the dielectric cavity can be obtained by adding the incident and reflected magnetic fields. Both the components of magnetic field are evaluated and shown in Eqs. 23 and 24, respectively.

$$H_x = -2\frac{E_{io}}{\eta}e^{-j\beta x \sin\theta_i} \cos\theta_i \cos(\beta z \cos\theta_i) \tag{23}$$

$$H_z = -2\frac{E_{io}}{\eta}e^{-j\beta x \sin\theta_i} \sin\theta_i \sin(\beta z \cos\theta_i) \tag{24}$$

From the expression of electric field, it can be seen that the electric field represents a travelling wave in $+x$ direction and a standing wave in z direction. Likewise, the magnetic field has a propagating nature in $+x$ direction and a standing wave nature in z direction. The expression of electric field and magnetic fields can be used to understand the power dissipation inside the dielectric cavity by finding the average pointing vector. From the expression of electric and magnetic fields it is evident that E_y and H_x are out of phase by 90°. Hence the power dissipation along z direction ($E_y \times H_x = -z$) is zero. On the contrary, the field components E_y and H_z are in phase with each other. Thus, signifies a power flow in $+x$ direction ($E_y \times H_z = x$). This clearly shows that the waves propagate in $+x$ direction while undergoing multiple reflection inside the dielectric cavity. But, with each set of incidence and reflection it loses all its power while forming standing wave in z direction.

To validate this theory, a narrow band metamaterial absorber operating at $10\mu m$ has been shown in Fig. 37. The thickness of the dielectric cavity is $t_{d1} = 4.48\ \mu m$, and the thickness of metallic layer is $0.035\ \mu m$. In order to validate the electromagnetic field expressions as derived above, the nature of electric field and magnetic field inside the dielectric cavity from $z = 0.035\ \mu m$ (bottom metallic reflector) to $z = (t_{d1} + 0.035)\mu m = 4.515\ \mu m$ (top metallic layer) has been evaluated and shown in Fig. 38.

Standing Wave Resonance Formation inside the Dielectric Cavity

It can be observed from Figs. 38(a) and (b), that the electric field represents a standing wave in z direction which varies with the term $\sin(\beta z)$, derived in Eq. 22. and magnetic field represents a standing wave in z direction which varies with the term $\cos(\beta z)$, derived in Eq. 23. While evaluating the electric and magnetic fields the term $\cos\theta_i$ has been ignored. This is due to the fact that the refractive index of ZnSe is more than twice as compared to the refractive index of air which makes the term $\cos\theta_i \approx 1$.

The evaluation of the curve as shown in Fig. 38 is for the case when the polarization angle is 0°. A study showing the variation of standing wave pattern

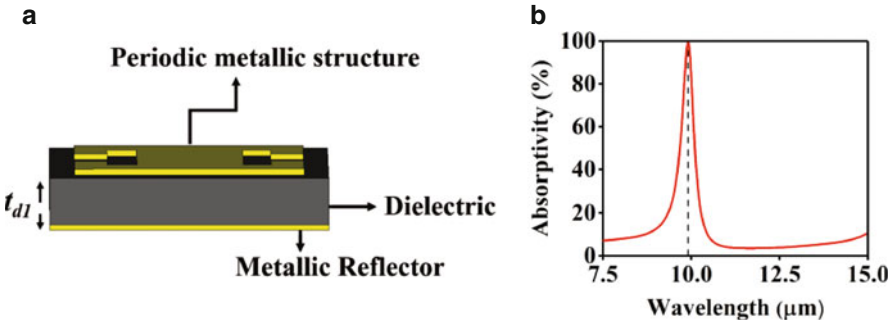

Fig. 37 (**a**) Schematic of the metamaterial absorber and (**b**) absorptivity curve. (Reproduced from Nilotpal et al. (2020b))

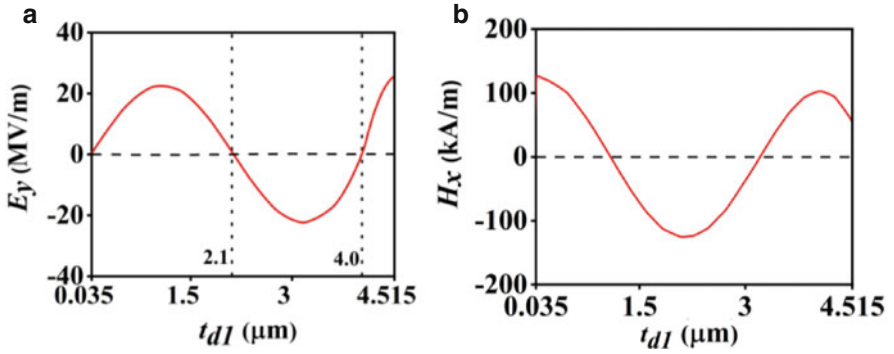

Fig. 38 Variation of real part of (**a**) E_y and (**b**) H_x component along z direction (from bottom metallic plate to top). (Reproduced from Nilotpal et al. (2020b))

with change in the polarization angle is done in Fig. 39. It can be seen that the standing wave pattern is maximum for $\phi = 0°$ which subsequently decreases when the polarization angle changes from 0° to 90°. The standing wave pattern ceases when the value of $\phi = 90°$. A further rise in the polarization angle increases the magnitude of the standing wave pattern which reaches a maximum (in negative phase) at $\phi = 180°$.

The above phenomenon can be explained in the light of polarization conversion also. When ϕ was 0° the electric field was in $+y$ direction, hence the E_y component of electric field has maximum magnitude of standing wave. With increase in ϕ value from 0° to 90°, the electric field changes its orientation from y to x as evident from a decrease in the standing wave pattern. At $\phi = 90°$, the electric field completely changes its orientation from y to x. Hence, there is no electric field in y direction, as evident from a zero standing wave pattern. For further increase in ϕ, value the electric field again shifts its orientation from x to $-y$. Hence a subsequent rise in standing wave pattern can be seen which reaches to a maximum (in negative phase) at $\phi = 180°$.

The above explanation is done for the TE wave incidence. However, the theory is well applicable for TM polarized wave too. To verify this, the effect of phase ϕ on

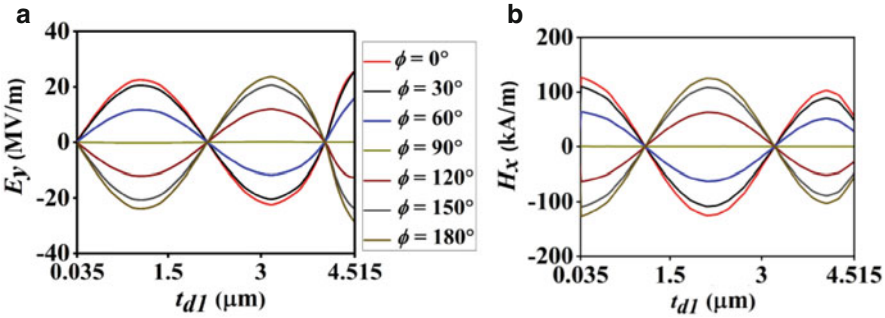

Fig. 39 Effect of polarization angles on the standing wave pattern for (**a**) E_y and (**b**) H_x component. (Reproduced from Nilotpal et al. (2020b))

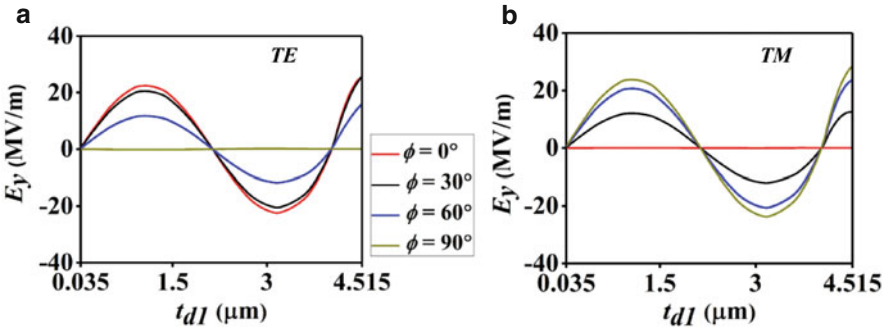

Fig. 40 Standing wave pattern for variation in polarization angle for (**a**) TE and (**b**) TM polarization. (Reproduced from Nilotpal et al. (2020b))

the standing wave pattern for both TE and TM polarized incident wave is shown in Figs. 40(a) and (b), respectively.

From Fig. 40, it can be verified that for TE polarized wave the electric field is in y direction, hence the E_y field pattern has maximum magnitude at $\phi = 0°$ which ceases for $\phi = 90°$. But, for the TM polarized incidence (electric field in x direction), the E_y component of electric field has a minimum magnitude for $\phi = 0°$ and maximum for $\phi = 90°$. Hence the theory which is applicable for TE polarized wave is equally applicable for TM polarized wave, while the only change required is in the direction of electric and magnetic fields vectors.

Study of Electric Field and Polarization Angle on the Orientation of Surface Currents

From Eqs. 10 and 11, it is already explained that the orientation of surface current primarily depends on the phase of the electric field. When electric field at the top and bottom metallic components are in phase then the orientation of surface currents at

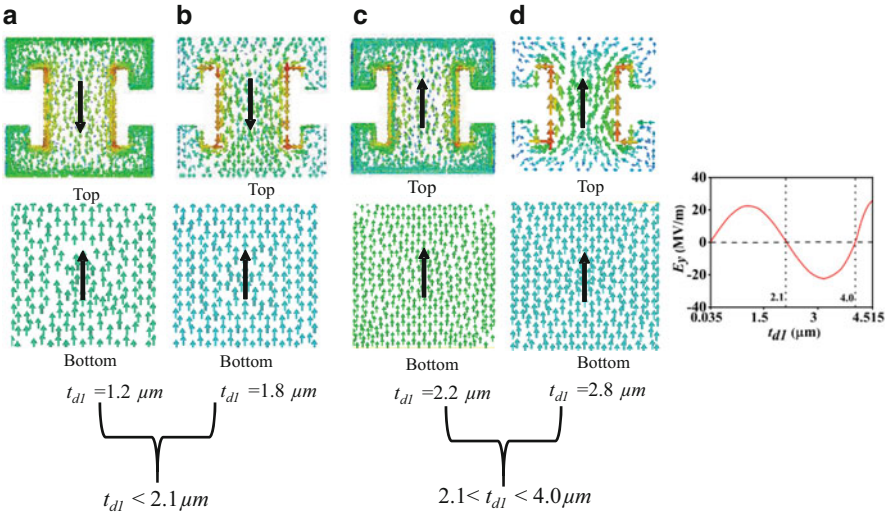

Fig. 41 Orientation of surface currents at the top and bottom metallic plates for the dielectric thickness (**a**) 1.2 μm (**b**) 1.8 μm (**c**) 2.2 μm and (**d**) 2.8 μm (Inset depicts the curve of Fig. 38(a)). (Reproduced from Nilotpal et al. (2020b))

the two metallic plates are antiparallel to each other. However, when the electric field at the top and bottom metallic plates are out of phase with each other, the parallel surface current can be obtained. To validate the point, the curve of electric field of Fig. 38 has been considered (also shown in the inset of Fig. 41) and the surface current for dielectric thickness; $t_{d1} = 1.2\mu m$, $t_{d1} = 1.8\mu m$, $t_{d1} = 2.2\mu m$, $t_{d1} = 2.8\mu m$ has been evaluated and illustrated in Fig. 41. It can be seen from the curve shown in the inset that for any structure having dielectric thickness $t_{d1} < 2.1\mu m$ the electric field at the top and bottom metallic plates are in phase and thus contribute to an antiparallel surface current as evaluated for $t_{d1} = 1.2\mu m$ and $t_{d1} = 1.8\mu m$, shown in Figs. 41(a) and (b), respectively. On the contrary, for the dielectric having thickness in the range $2.1\mu m < t_{d1} < 4.0\mu m$. The electric field at the top and bottom metallic sheets will be out of phase. Thus, contribute to a parallel surface current, as evaluated for the case $t_{d1} = 2.2\mu m$ and $t_{d1} = 2.8\mu m$ in Figs. 41(c) and (d), respectively.

The effect of change in the polarization angle on the orientation of surface current has been depicted in Fig. 42. From the curve of Fig. 42(a) it can be seen that the standing wave pattern for $\phi = 0^\circ$ and $\phi = 180^\circ$ are exactly equal in magnitude, but opposite in phase. Thus the effect of the standing wave patten on the surface current orientation will also be equal in magnitude but opposite in phase as shown as case 1 of Fig. 37 it can be seen that the surface current orientation for $\phi = 0^\circ$ is exactly equal and opposite to that of $\phi = 180^\circ$ considering the same physics the orientation of surface current for $\phi = 30^\circ$ and $\phi = 210^\circ$ are also equal in magnitude and opposite in orientation, shown as case 2 of Fig. 42.

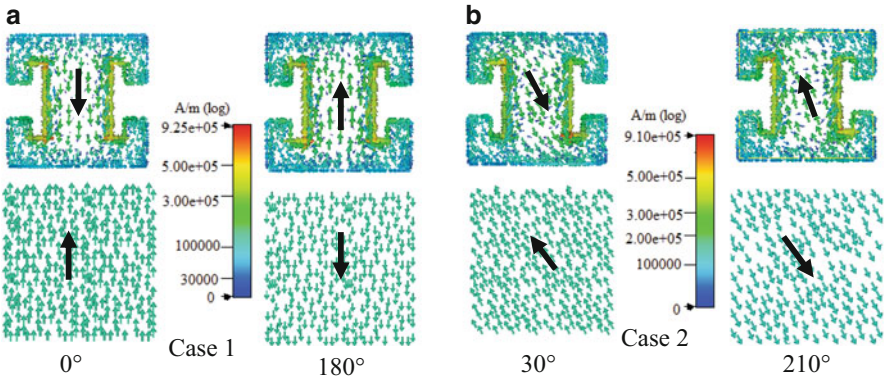

Fig. 42 Surface current orientation at the top and bottom metallic plate for (**a**) case 1: (0° and 180°) (**b**) case 2: (30° and 210°). (Reproduced from Nilotpal et al. (2020b))

Metasurface Absorbers Operating in the Optical Frequency Range

Unlike the metasurface absorber in GHz or THz ranges, the most challenging task in the realization of metasurface absorber in the visible ranges is the small size of the unit-cell. In this region the unit cell's size should be much smaller than the resonating wavelength to fulfill the sub-wavelength condition of metamaterial design. As in this frequency region, the size of the unit-cell is of the order of some hundred nanometer, therefore, the size of the array should be few tens on nm. The techniques which is available currently is sufficient to do the nanofabrication, however, on the cost of uniformity of individual unit cells.

The work done in Xu et al. (2019) has presented a dual band metasurface absorber where one of the absorption band is in the infrared frequency region while other lies in the visible frequency band. The absorption in the infrared frequency band attributes to the excitation of a propagating surface plasmon polariton and the absorption in the visible spectrum attributes to the localized magnetic polariton in the proposed nanostructure. For the fabrication process the authors have used a simple and cost-effective colloidal lithography nanofabrication method. The diameter of the aluminum nano disc can be varied to vary the position of absorption.

A broadband metasurface absorber for the frequency range 100 THz to 1000THz has been proposed in Patel et al. (2019). The top metasurface layer consists of a pyramid shape block made up of tungsten material. The response of the absorber has been compared for three different dielectric materials, viz., SiO_2, SiN, and GaAs. It has been found that the absorber having as SiO_2 substrate is giving the best result among the three. Analytically, the structure exhibits an average absorption of 79.67%, 93.7% and 88.4% respectively in all three regions, i.e., Infrared, Visible, and Ultraviolet. The structure provides a unique flexibility band which can be tuned between low-frequency band to the high-frequency band and consequently can be

useful for high-performance terahertz devices and can be used in the design of efficient solar cells.

Conclusion

This chapter is dedicated to metamaterial absorbers. Initially, the chapter deals with the evolution of metamaterial absorbers by introducing the design limitations of conventional absorbers. The absorption mechanism along with the role of each layer to achieve unity absorption has been explained in detail. A simulative study elucidating the role of loss tangent on the absorptivity response is presented in detail. The chapter presents the design of various singleband, multiband, and broadband metamaterial absorbers in the S, C, and X bands where the structure consists of an array of sub-wavelength unit cells. The mechanism of absorption has been explained on the basis of electric field distribution on the top metallic pattern and the orientation of surface currents on the top and bottom metallic patterns. The experimental set up along with the testing method has been discussed step by step. The chapter also discusses several metamaterial absorbers designs in the THz, Infrared regime, and Optical region.

Next, metamaterial absorption based on multiple reflection interference model has been discussed. The repetition of absorption characteristics with each increment of half wavelength in the dielectric thickness has been explained in the light of formation of higher order modes. Through higher order resonance, it has been shown that presence of antiparallel surface current is not the mandatory condition for absorption in a metamaterial structure. To validate the multiple reflection phenomenon, a time domain analysis of reflection coefficient phenomenon using inverse chirp-Z transform technique has been implemented. The last section of the chapter presents the mathematical modelling of the metamaterial absorbers, where all the characteristics of metamaterial absorber such as absorption, polarization insensitivity, surface current orientation has been explained mathematically in the light of electric and magnetic fields.

In summary, the chapter deals with various metamaterial design in the GHz, THz, Infrared, and Optical regime. The chapter discusses various types of absorption mechanism such as absorption based on electric and magnetic resonance and based on multiple reflection interference method. The chapter presents a very detailed study on mathematical modelling of the metamaterial absorber which can be helpful to understand absorption, wave prorogation.

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Graphene-Based Tunable Metamaterial-FSS RAS

3

Gaurav Pal Singh and Neha Sardana

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Abstract

Metamaterials (MMs) are periodic and/or non-periodic subwavelength structures with a unique response to electromagnetic (EM) waves compared to conventional materials. MMs possess negative permittivity and/or permeability; the combination of both negative is not found in conventional materials. The shape, dimensions, and materials used in the MM determine its ultimate working frequency and response. Owing to their peculiar properties, MMs have found new-age applications such as cloaking, perfect absorber, hyperlenses, modulators, holography, etc. Graphene is a remarkable material with distinctive properties, including excellent thermal and electrical conductivity, very high mechanical strength, along with having a zero bandgap. Further, one of the most important properties of graphene, which has enabled it to be widely used as/along with MMs, is its

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tunable response to EM waves via a change in conductivity when its chemical potential is varied by applying a bias or doping. Frequency selective surfaces (FSSs), or flat MMs, and MMs utilizing graphene, find applications for defense and security systems in infrared, terahertz, and microwave frequency range. The broad areas of application include stealth, communications, and imaging. Stealth applications aim to prevent radar detection by absorbing or dissipating the incoming EM waves using graphene MMs or FSSs [which act as radar absorbing surfaces (RASs)]. Another important stealth application, although a futuristic one, is cloaking. Fast, secure, and high volume data transfer is critical for any security system, and graphene MMs can improve upon the existing technologies, particularly in the terahertz frequency range. Graphene MMs can be used to develop lenses, mirrors, hyperlenses, modulators, etc., for infrared (thermal) and terahertz imaging systems. In the current chapter, the basic properties of graphene, the working of FSSs and RASs, and applications of graphene-based MMs in defense and security systems are discussed.

Keywords

Graphene · Metamaterial · FSS · RAS · Data encryption · Cloaking

Introduction

Nature has provided mankind with materials for a wide variety of applications, but some applications require properties to go beyond the limits of naturally available materials. Therefore, metamaterials (MMs) are being developed to fulfill these requirements with a combination of negative permittivity and/or permeability (Padilla et al. 2006). MMs are able to interact with electromagnetic (EM) waves in a unique way due to the presence of periodic or non-periodic structures which alter the EM response of the original constituent materials. The structures range from the order of nanometers to millimeters, depending upon the targeted working range of EM waves. There has been considerable development of materials in the visible region of the EM spectrum; however, there is a lack of availability of materials to control the far-infrared [terahertz (THz) range] and microwave frequencies (Federici and Moeller 2010). The THz frequency range is gaining importance in communication, security, sensing, and characterization. MMs are widely used in this range, although MMs have applications all over the EM spectrum. As a result, they find uses in cloaking (Rainwater et al. 2012), perfect absorbers (Landy et al. 2008), superlenses (Lipworth et al. 2015), modulators (Chen et al. 2009), antennas (Li et al. 2010), sensing (Xu et al. 2017), and various other areas. MMs, owing to their diverse and distinctive properties have been established in the defense sector since the 2000s (Smith et al. 2000). The keen interest in defense research and development has contributed immensely to the exploration of MMs (Derov et al. 2017). MMs in the defense sector are primarily focused on stealth, communication, and imaging. New and exciting applications such as anti-fog glasses, glasses to protect pilots from

visual impairment from laser strikes, transparent antennas, high-frequency communications, faster and more accurate depth perception, etc., are specifically aimed toward using MMs. The structure of the repeating unit cell in a periodic MM defines its overall properties irrespective of the bulk material properties. The common structures used are thin wires, split ring resonators (SRRs), and H-shaped structures (Monticone and Alù 2014). The characteristic dimension of these structures is usually much less the wavelength range in which the material is operated; therefore, the fabrication of MMs is highly challenging and tedious. The tunability of the electric and magnetic field response of the MMs is a highly desirable property that can increase the scope of its use. Some of the techniques used to achieve tunability for passive MMs include changing structure, material, and employing micromechanical systems (Vendik et al. 2012). The active MMs include ferroelectric, ferromagnetic, semiconductor, and phase change materials; these can be tuned using electrical, optical, or thermal approaches (Rout and Sonkusale 2017). However, graphene can achieve a wide frequency range of tunability, which is simpler and, therefore, more practical to attain than the aforementioned techniques. The chemical potential (Fermi energy) of graphene can be tuned by controlling its doping or varying the applied biasing voltage. The conductivity of graphene varies as its Fermi energy is tuned, thereby achieving a tunable response. Moreover, owing to its near-zero bandgap, graphene can absorb EM waves from ultraviolet (UV) to the THz range. The properties and preparation of graphene will be summarized in greater detail in the next section, followed by the working of frequency selective surfaces (FSSs) and radar absorbing surfaces (RASs). Thereafter, the chapter will focus on the advancements in technologies employing tunable graphene MMs, which are useful in defense and security applications.

Graphene

The discovery and understanding of graphene was an important step in the progress of science, which was validated when the Nobel Prize in Physics was awarded to Andre Geim in 2010 "for groundbreaking experiments regarding two-dimensional material graphene." Graphene is a monolayer of carbon atoms arranged as a hexagonal lattice. Graphene has very high tensile strength and remarkably high thermal and electrical conductivity despite being just an atom within and hence, is not visible to the naked human eye (Poothanari et al. 2019). Moreover, graphene flakes can be peeled off using a regular adhesive tape from a piece of graphite. Despite these extraordinary properties, graphene is still far from commercial applications due to the cost and complexity of producing large and defect-free graphene sheets (Choi and Lee 2016). The closest application of graphene for industries is graphene composites, which pertains to improving the properties of conventional materials by adding graphene. Furthermore, it is agreed that rather than finding applications particularly for graphene, the integration of graphene into commercial applications will be much faster if graphene improves upon the currently available technology (STOYE 2015). The remarkable properties of graphene can be exploited to improve

the existing MM technology. The distinctive property of graphene employed in MM technology is the tunability of its response to EM waves via the applied voltage. This section of the chapter aims to present the structure, basic electronic properties, and ultimately its tunable response.

Properties

Band Structure

The electronic states of a 3D crystalline solid are denoted by an infinite number of periodic wave functions that are infinitely spread over space but have eigen energies within a finite energy range (energy band). Graphene is a crystalline material with a hexagonal lattice consisting of carbon atoms with 2s and two 2p orbitals. The carbon atoms which are hybridized are connected via the σ band, which is contributed by 3/4 of the valence electrons. The 2p (p_z) orbital, which is perpendicular to the σ band, forms the π band, which is partially filled by the remaining valence electrons. The σ band is the main contributor to the strength of graphene, while the π band is responsible for the optical and electronic properties. To calculate the band structure of graphene, initially tight-binding approximation has been considered (Wallace 1947), which neglects the interaction between the p_z orbitals; further, the interaction between nearest neighbor carbon atoms is calculated ($\gamma_0 = 3.15$ eV). Thereafter, the Schrödinger equation is solved to get the expression for the energy of the π band (Wallace 1947):

$$E_{C,V}(k) = \pm\gamma_0\sqrt{1 + 4\cos\left(\frac{k_y a}{2}\right)\cos\left(\frac{\sqrt{3}k_x a}{2}\right) + 4\cos^2\left(\frac{k_y a}{2}\right)} \quad (1)$$

where + and – signs are for the energy of the π [conduction (E_C)] and π* [valence (E_V)] band, respectively. $k = (k_x, k_y)$, is the wavenumber of the wave function of an electronic state in the π band, and a is the lattice constant (0.246 nm). Here, the π band is regarded as fully filled, and π* band is empty; and the Fermi energy is located between the π and π* bands. The electronic states in Eq. (1) are represented in the Brillouin zone in the k space (reciprocal lattice), the 3D representation of which is shown in Fig. 1a. The Dirac point is where conduction and valence band touch each other; it is located at the K point (in a hexagonal lattice, it is a point of high symmetry at the middle of an edge joining two rectangular faces). The top and bottom of the valence and conduction band lie in the center (Γ) of the Brillouin zone. An approximate conical nature of the band structure is observed near the K point (Fig. 1a inset). When the origin is shifted from Γ to K, the Eq. (1) is approximated as:

$$E_{C,V}(k) = \pm\gamma_0 ak\frac{\sqrt{3}}{2} \quad (2)$$

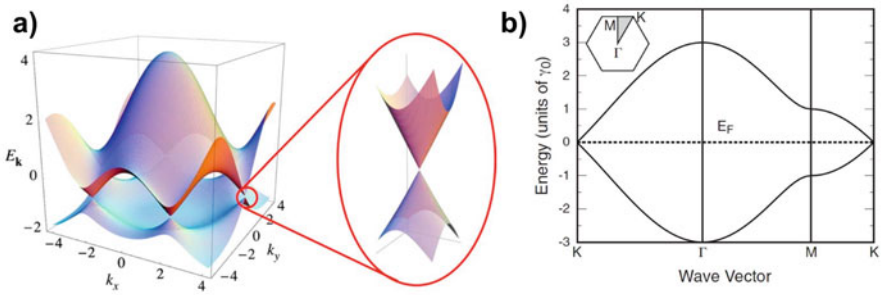

Fig. 1 (**a**) The 3D π electronic band structure of graphene in reciprocal space. The inset shows the magnified view of the structure near Dirac point. (Reprinted from Ref. (Castro Neto et al. 2009), © 2009, with permission from the American Physical Society). (**b**) 2D representation of the band structure along K/ Γ /M/K points in the reciprocal lattice. (Reprinted from Ref. (Ando 2005), © 2005, with permission from Physical Society of Japan)

where k is the norm of the wavenumber vector k whose origin is shifted. The linear relation of the energy is highly unusual for a 2D system, which is generally parabolic in nature. The density of states is derived from the Eq. (2) as:

$$D(E) = \frac{8|E|}{3\pi\gamma_0^2 a^3} \tag{3}$$

The density of states of graphene is proportional to the energy; furthermore, the bandgap is zero at the Fermi energy. Hence, the effective mass of graphene electrons/holes is zero, even at room temperature. This phenomenon is attributed to the movement of electrons in the periodic hexagonal lattice creating a new quasiparticle called Dirac fermion which is massless (Geim and Novoselov 2007). Due to their "massless" nature, electrons/holes in graphene have very high mobility, which can theoretically reach 2,00,000 cm^2/Vs (Lemme 2009). Therefore, the electrons can travel large distances at very high speeds across the graphene lattice with minimal scattering. Moreover, the temperature has little effect on mobility near room temperature (Tan et al. 2007).

The zero band gap of graphene is the source of its remarkable properties and makes it a revolutionary material in optical and electronic-based applications. However, the above expressions are derived considering monolayer graphene; as the number of layers is increased, its band structure and properties change. The bilayer graphene has zero bandgap for one layer, but for the other layer, the linear nature of the band structure is changed to parabolic (still symmetric), and a bandgap is created (Takai et al. 2019). When more layers are introduced, the parity of layers and the interaction of other orbitals (previously only the effect of p_z orbital was considered) play an essential role in determining the electronic band structure.

Conductivity

In some electronic and optical applications, it is necessary to introduce a bandgap in graphene. There are various methods of tuning the bandgap of graphene. The main techniques are applying a gate voltage and changing the chemical structure (doping). Modifying the gate voltage is a simple and effective method to obtain tunable properties. The conductivity of graphene will be studied in greater detail to understand the tunability of graphene. The main contributors to the surface conductivity in the THz to the far IR of monolayer graphene are the interband and intraband transitions, their contributions are derived from the Kubo formula (Hanson 2008):

$$\sigma_{intra} = -j\frac{e^2 k_B T}{\pi\hbar^2(\omega - j2\Gamma)}\left[\frac{\mu_C}{k_B T} + 2\ \ln\left(e^{-\frac{\mu_C}{k_B T}} + 1\right)\right] \tag{4}$$

$$\sigma_{inter} = -j\frac{e^2}{4\pi\hbar}\log\left[\frac{2|\mu_C| - (\omega - j2\Gamma)\hbar}{2|\mu_C| + (\omega - j2\Gamma)\hbar}\right] \tag{5}$$

where e is the electron charge, K_B is the Boltzmann constant, T is the temperature, ω is the angular frequency, ħ is the reduced Planck's constant, Γ is the scattering rate that takes into account the type of losses, and μ_C is the chemical potential (Fermi energy). In Eqs. (4) and (5), it is considered that graphene is isotropic, and there is no magnetism. The intraband term dominates for microwave and THz region at room temperature. The carrier density for monolayer graphene is expressed as (Hanson 2008):

$$n_s = \frac{2}{\pi\hbar^2 v_F^2}\int_0^\infty \varepsilon[f(\varepsilon) - f(\varepsilon + 2\mu_C)]d\varepsilon \tag{6}$$

where v_F is the Fermi velocity which is approximately 10^6 m/s, and ε is the permittivity. On the application of bias V_g, the Fermi energy or chemical potential of graphene can be altered, which is represented as (Gómez-Díaz and Perruisseau-Carrier 2013):

$$E_F = \mu_c \approx \hbar v_F\sqrt{\frac{\pi\varepsilon_r\varepsilon_0 V_g}{et_s}} \tag{7}$$

where t_s, ε_r, and ε_0 are the thickness of the substrate, permittivity of the substrate, and vacuum permittivity, respectively. Therefore, on the application of bias, the response (conductivity or resistivity) of graphene can be tuned (Fig. 2). This phenomenon is observed several times in the next section of the chapter, where the applications of tunable graphene are discussed. However, the equations are for monolayer graphene; as mentioned earlier, as the number of layers increases, the properties of graphene change as there is EM interaction between the layers. Further, the defects and purity of graphene immensely influence its properties. The number of layers in graphene and its defects can be deduced from the Raman plot of the sample.

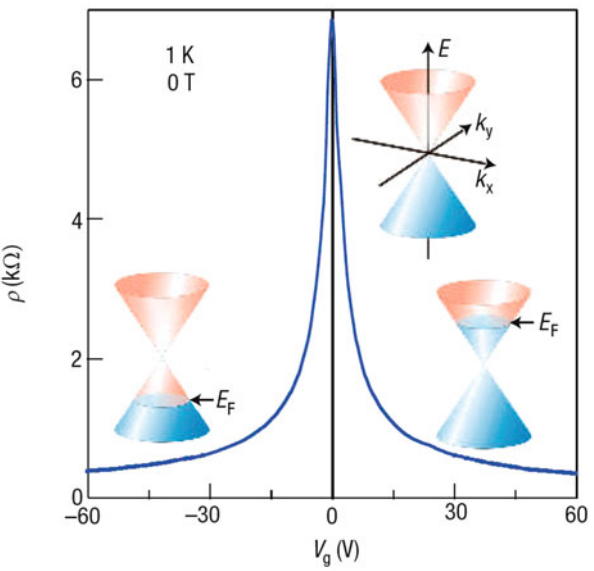

Fig. 2 Resistivity (ρ) response of monolayer graphene on the application of gate voltage (Vg). The insets show the change of the position of Fermi energy in the simplified energy band diagrams. (Reprinted from Ref. (Geim and Novoselov 2007), © 2007, with permission from Nature Publishing Group)

Preparation

The commercial applications of graphene are restricted by the challenges in the preparation of graphene. The discussion of the preparation of graphene will be limited to graphene sheets since they are the most common form used in MM-based devices. There are mainly three methods of graphene fabrication, chemical vapor deposition (CVD), cleavage (peeling), and exfoliation. CVD is a bottom-up approach, while cleavage and exfoliation are top-down methods, in which graphene is prepared from graphite and graphene oxide, respectively. CVD and cleavage are much more effective in producing graphene with minimal defects as compared to exfoliation; however, these processes are slower and are restricted to research laboratories. Although exfoliation introduces considerable defects, it is used to produce graphene sheets on a commercial scale. The method of graphene production influences the properties obtained, as each method of fabrication produces its own unique defects in the structure (Takai et al. 2019). Exfoliation uses graphene oxide, which is produced by severe oxidation of graphite, followed by exfoliation using thermal or chemical methods and then reduction. CVD uses various hydrocarbon gases (ethylene, methane, etc.) to produce high purity graphene on substrates such as platinum, copper, nickel, metal and silicon carbides, etc., under high vacuum conditions. The graphene prepared on these substrates can be transferred to a desired substrate such as silicon, SiO_2, Polyethylene terephthalate (PET), etc., without the introduction of any significant defects. However, CVD fabrication requires optimizing process parameters, including choice, concentration, and flow rate of precursor, substrate, heating and cooling rate, and temperatures. The method of preparation

using cleavage can be done via mechanical means or in solution. Highly crystalline graphite, which is available in the form of highly oriented pyrolytic graphite (HOPG) and kish graphite, is used to extract graphene as sheets/flakes. As stated earlier, graphene flakes can be obtained just by using adhesive tape repeatedly until the desired thickness of less than 100 nm (Ohashi et al. 1997). However, sophisticated techniques such as silicon cantilever used in atomic force microscopy (AFM) to transfer thin flakes from HOPG onto SiO_2 substrates are generally used to get high purity graphene sheets (Zhang et al. 2005). The cleavage process can also be done in solution; there are two methods, surfactant-free and surfactant-assisted. Various solvents are used to disperse graphite flakes; however, the process must be optimized to get the desired number of layers with minimal defects. Flakes down to the thickness of 1 nm can be obtained via this process (ANTISARI et al. 2006). Single-layer graphene shows rippling of about 1 nm to obtain thermodynamic stability (Meyer et al. 2007). Graphene is doped with heteroatoms to suppress the rippling (Wang et al. 2014). Among all the processes, CVD is the most suited to produce thin, large area, and defect-free graphene sheets, with a higher yield than the cleavage process. Further, it can be combined with the lithographic process, which is commonly used in fabricating electronic circuits.

Frequency Selective Surfaces

Frequency selective surface (FSS) or metasurface is a surface that has periodic and commonly subwavelength structures arranged on top of a substrate. The FSS generally have some resonant frequency at which either the EM wave will be entirely or partially transmitted or reflected. The requirements of a good FSS are broadband range, effectiveness at various incidence angles, and dual polarization. Moreover, the fabrication of an FSS is much simpler than constructing a 3D MM. The thickness of an FSS is much less than the wavelength of the EM waves which are incident upon it; however, it is larger than the skin depth of the material used. The EM wave passing through the metasurface can be modified in both amplitude and phase. To simplify the study of the EM response of FSS, the basic FSS configurations can be represented as low pass, high pass filter, band stop, and bandpass filters; and the structure can be represented as inductance and capacitive circuits (Fig. 3). The structures can be envisioned to create resistance and inductance, while the gaps between them can be considered to generate capacitance (Ullah 2012). The characteristic response of these surfaces can be changed by altering the geometry of the structures and the gap between them (Munk 2000), and also the materials used. Therefore, varied responses can be created by creating complex geometries and using materials with unique properties. This approach is called equivalent circuit modeling. Similar to the 3D MMs, tuning the response of FSSs by varying the geometry is an uneconomical task because creating structures of the order of nanometers to micrometers accurately and over long distances is very tedious and complex. The relatively unique and still accessible properties of graphene can help

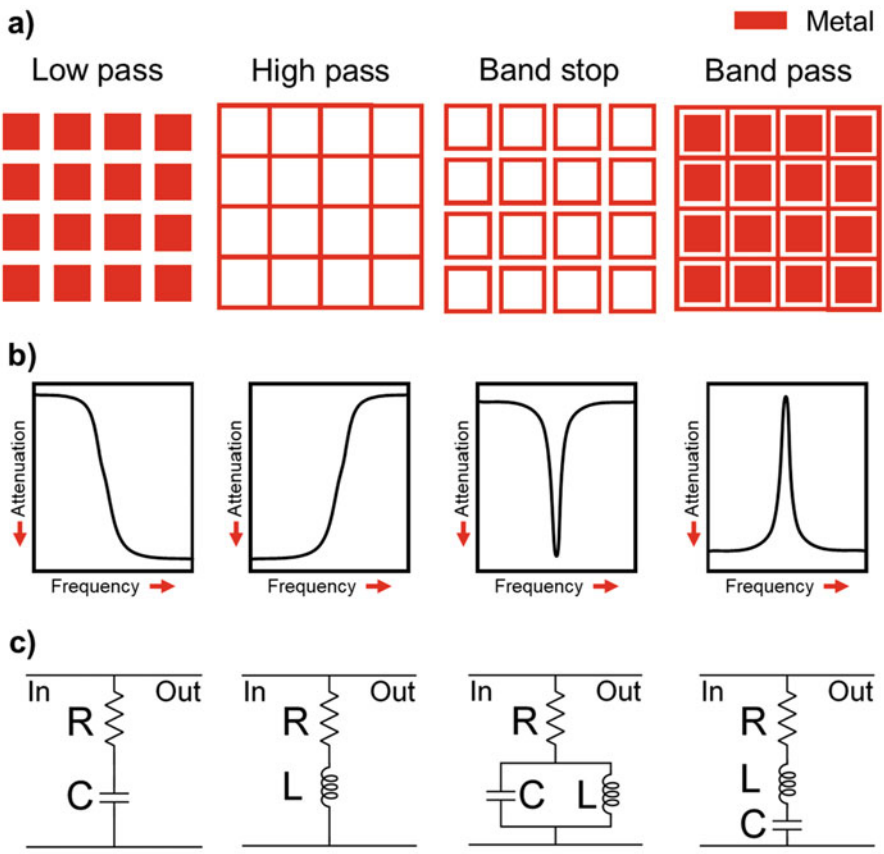

Fig. 3 (**a**) Low pass, high pass, band stop, and band pass FSS filters with the (**b**) frequency responses and (**c**) equivalent circuits

these FSSs and MMs to tune the response of the structures by simply varying the gate voltage. Therefore, without creating new structures or using new materials, the response of the structures can be tuned.

Radar Absorbing Surfaces

The radar has been an essential component of defense technology since World War II; modern military technology generally accompanies radar cross-section (RCS) in its performance specifications. The response of the vehicle/weapon to EM waves is an essential criterion in determining its success over the battlefield (Sadhu 1995;

Knott et al. 2004). The vehicle/weapon could be destroyed even before reaching its intended target if the enemy is easily able to track it via radars, which can track the position as well as the speed of the target. Stealth is of vital importance in military technology; this led to the U.S. military stealth aircrafts such as Lockheed A-12 and F-19 being kept a secret from public records (although A-12 was made public in 1964 (O'Leary and Schulzinger 1991)) to prevent the replicating of technology. Radar emits EM waves that reflect off the target and are captured by the antennas; usually, the same antenna is used to emit and detect the EM waves. The antennas are positioned such that they have two axes of rotation, which enables the radar to accurately capture the position of the target. Arrays are often used as antennas to capture the signal from multiple directions without physically moving the antenna, enabling faster scan rate radars. The frequency used depends upon the range and application of the radar system. The radars for defense applications are generally in the frequency range of microwaves and radio waves. Other applications of enemy detection include optical, acoustic, and IR (thermal) imaging. IR imaging is crucial in low light and applications where optical systems can fail. However, IR waves are highly prone to scattering by airborne particles. However, THz-based systems owing to their higher frequency, have a large range in rainy and foggy weather as compared to the generally used IR systems (Brown 2003; Federici and Moeller 2010). Radar signal can be reduced practically by mainly two methods, reflecting the signal away from the surface of the object by using specific geometries or by using radar absorbing surface (RAS). The former is usually geared toward suppressing the signal at some angle but might cause enhancement at others. RAS dissipates the EM waves within the material and its structure; therefore, all of the signals will be absorbed by the material, and none will be left to be detected by the radar. However, RASs are complex to construct, maintain, and have added weight and volume.

Applications of Graphene-Based Metamaterials in Defense and Security Systems

Graphene-based MMs and FSSs find a wide range of applications in defense and security systems. FSS can be simply considered a subgroup of MM as there is no particular thickness defined where an FSS can be considered a MM. Applications of graphene-based MMs are spread over a broadband range of frequencies, from the ultraviolet (UV) to the radio frequency region. However, defense and security applications mainly range from the infrared (IR) to the microwave region, and these regions have different units that are generally used to represent the wavelength/frequency of EM waves. The applications discussed in this section discussing IR region, THz region, and microwave region will be mentioning the wavelength/frequency in μm [1 μm = (Speed of light $\times$ 10^6) Hz], THz (1 THz = 10^{12} Hz), and GHz (1 GHz = 10^9 Hz), respectively. The applications of graphene-based MMs in defense and security systems are divided into three broad categories of stealth, communication, and imaging.

Stealth

There are various MMs developed to be perfect absorbers used in applications such as photodetection, anti-reflective coatings, sensors, etc. As mentioned in the previous section RCS is a significant area of interest for the defense sector. The MM-based perfect absorbers can improve the efficiency of the absorbers as well as decrease the size and weight of the absorbing components owning to their sub-wavelength structures. The losses in a material when interacting with EM waves are defined by the imaginary part of permittivity and permeability. The permittivity and permeability of MMs and FSSs can be controlled. Furthermore, control over the response can be achieved by using the tunable properties of graphene. The structure of the MMs can cause the incident EM waves to be trapped in the regions containing cavities and the energy to be absorbed via the destructive interference. The absorption should be over as broad range of frequencies as possible to get the maximum probability of protection from the enemy's radar.

Absorbers

IR Range Applications

The absorbance over the IR range is a vital defense application because thermal detection technology has significantly improved and is widely accessible. Lin et al. (2019) developed a broadband absorber in the visible to near IR spectrum (0.3 μm to 2.5 μm) using stacks of alternating layers of graphene and a dielectric (total thickness 90 nm) separated by a layer of air. The overall structure acted as a grating for the incoming EM waves, which leads to coupling, and hence, the high absorption. The direct stacking of graphene layers on top of each other leads to the formation of graphite which has very different optical properties compared to graphene, therefore, the multilayer structure was used. The average response from incident angles from 0° to 60° showed more than 80% absorption from 0.3 μm to 1 μm; however, it decreased to about 60% as the wavelength increased to 2.5 μm. As a result of the design of the FSS, the absorber was polarization independent. The experimental results of absorptivity were in close agreement with the simulation for both the TE and TM modes. The absorber was not tunable due to the complexity involved in applying voltage. However, the absorption was very high as compared to existing technologies. Yao et al. (2021) incorporated tunability to the FSS absorber by using perpendicular and parallel layers of graphene nanoribbons sandwiched between Al_2O_3 and by applying a bias. The modulation achieved was more than 50% upon application of a voltage of 14 V. Bandwidth of 3.55 μm, which was more than 1.7 times that of monolayer graphene was measured. The small relaxation time of graphene was regarded as the main contributor to the broadband absorption. The simulation was performed on finite element method (FEM) based software COMSOL, and validated with experimental results. The structure was deemed to polarization angle dependent. The structure was nanopatterned and graphene was grown using CVD. Similarly, Safaei et al. (2017) used nanoimprinting for patterning an array of CVD grown graphene to develop and absorb in the mid-IR region. The

patterned monolayer graphene was used as an absorber FSS over a broadband range of 8–12 μm. The structure was tunable up to 2 μm. The experimental and simulation results were in close agreement with each other.

THz Range Applications

As explained earlier in the RAS section, the THz frequency region is gaining importance in stealth applications. Further, the lack of response to this region from conventional materials led to MMs, specifically graphene MMs, to be highly prevalent in literature as THz absorbers. He et al. (2016) developed a tunable absorber in the THz range, consisting of the following structure from top to bottom, graphene, SiO_2, silicon, polyimide, and copper ground plate. As the μ_C was increased from 0.1 eV to 1.0 eV, the reflection peak was shifted toward higher frequencies. Three types of simple unit cells were analyzed, circular, rectangular, and cross-shaped. All the shapes were enclosed in a 24 μm^2 area, and periodicity was kept constant at 32 μm. The response of all the periodic structures was analyzed in CST Microwave Studio. The circular and rectangular structure showed a single peak in the frequency range up to 3.5 THz. However, in the cross-shaped structure, two peaks were observed because of the asymmetry of the structure. The absorption modulation depth was the highest for the cross geometry; the main reason was deemed to be the smaller area of the structure. Moving toward more complicated patterns, Lie et al. (2018b) simulated a broadband tunable absorber with single-layer graphene rings, which were connected to each other on a metallic perfect electric conductor (PEC) ground plate separated by a dielectric layer. The absorber was polarization independent at normal incidence due to the inherent symmetry of the unit cell. A large bandwidth, from 2.25 THz to 3.2 THz, of greater than 90% absorption was achieved for the single ring and dual ring (one smaller ring inside the other), respectively. As μ_C was varied from 0.4 eV to 1.0 eV, the absorption and spectral range increased, and the peak moved toward higher frequency for the single ring structure. Jiang et al. (2020) incorporated three concentric layer split ring pattern of monolayer graphene on a SiO_2 layer, placed on top of a PEC metallic mirror. The ring of each layer was of different size and connected to a gold gate, and each layer could be tuned separately. The response of the structure was simulated using a finite difference time domain (FDTD) solution. Absorption of frequencies ranging from 0.6 THz to 1.8 THz was observed. The absorption peak was tuned by varying the μ_C of graphene. The response of the structure was angle and polarization insensitive due to the symmetry of the structure. A wide tunability range was achieved, which was confirmed by varying the gate voltage from 0.7 eV to 1.0 eV for all rings. The absorption peak was blue shifted as a result of the applied voltage. The bias voltage of the two rings was kept constant at 1 eV, and for the smallest ring, it was varied from 0.7 eV to 1 eV, which led to a more continuous absorption. There was an intersection of peaks at some combination of the bias applied, which drastically increased the absorption peak. In the final design, the absorption rate was continuous and above 70%. Wu et al. (2019) used elliptical graphene with a gradient edge as the periodic structure to simulate a tunable absorber. The absorption rate increased from 0% to 100% when the μ_C was changed from 0.0 to 1.0 eV. Further, the absorption bandwidth increased

from 20% to 90.32% as the μ_C was varied. The absorber was polarization independent and displayed good absorption for a wide range of incident angles (0°–60°). At the edges of the absorption band for μ_C 0.8 eV, 1.7 THz and 4.3 THz, the electric field concentration was on the vertex of the focused main diagonal and off-diagonal, respectively. The mechanism of the absorption was explained using the asymmetric Fabry-Perot cavity (Liu et al. 2009). It was concluded that under normal angle of incidence, the periodic structures could be neglected, however at oblique angles, the absorption has to be calculated by the interference theory. Mishra et al. (2018) used two types of unit cells, one was a single loop square and the other was a double loop square. The designed FSS consisted of metallic structures upon a SiO_2 layer, which had a graphene layer and dielectric substrate placed on top of a metal plate which acts as a PEC metallic mirror. The structure was easier to fabricate, since, the graphene monolayer did not have to be patterned. Moreover, metallic structures are much easier to fabricate than graphene. The transmission line theory and equivalent circuit modeling were used to design the FSS for easy and accurate analysis. The structure was designed for minimal reflectivity by calculating the reflection coefficient. At −10 dB, a broadband absorption range of 0.79–1.47 THz was achieved for RCS reduction. The μ_C of graphene was varied from 0.15 eV to 0.55 eV to tune the response of the FSS. As the μ_C was increased, the minimum reflection moved to higher frequencies. Lengths of squares, strip width, gap between strips and periodicity, dielectric thickness was varied. It was observed that there was an increase in the absorption bandwidth when the double square configuration was used instead of single. Simulation results were used to validate the equivalent circuit modeling. The optimal absorber had μ_C 0.35 eV and a bandwidth of 0.68 THz.

The aforementioned absorbers were designed for a broadband absorbance; however, some specific applications require dual-band or multi-band absorbance. Yan et al. (2020) selected a monolayer dumbbell-shaped graphene placed on a silicon substrate and a grounding PEC gold plate as the unit cell of the absorber (Fig. 4a). The design was aimed to create a narrow bandwidth and dual-band absorption. The absorption peaks were located at 0.2242 THz and 0.5302 THz, and the absorption was more than 99.6% for both peaks. Good performance was observed up to 80° angle of incidence under TM polarization; therefore, the absorber was shown to have wide-angle usability. Simulations were used to study the absorber, and the results were validated with the equivalent circuit modeling. The performance of the absorber was improved by optimizing the geometry of the FSS. The μ_C was varied from 0.2 eV to 1.6 eV, the absorption peaks were slightly blue shifted. The width of the peak absorption first decreases and then increases as the μ_C was varied. At μ_C 0.2 eV, the two peaks disappeared and a wide-band absorption of greater than 90% was observed (Fig. 4b). Multimode tunability of graphene MM in the THz range was achieved by Hu et al. (2018). The unit cell structure consisted of a gold graphene pattern that had a gold ring which was cut by two parallel graphene ribbons. The structure was placed on an Al_2O_3 substrate. The unit cell structure was followed by SiO_2, graphene, Al_2O_3, graphene, SiO_2, and finally, a grounding gold sheet. The response of the patterned graphene and the following graphene layer was tuned separately. The response of the MM was analyzed via circuit modeling as well as

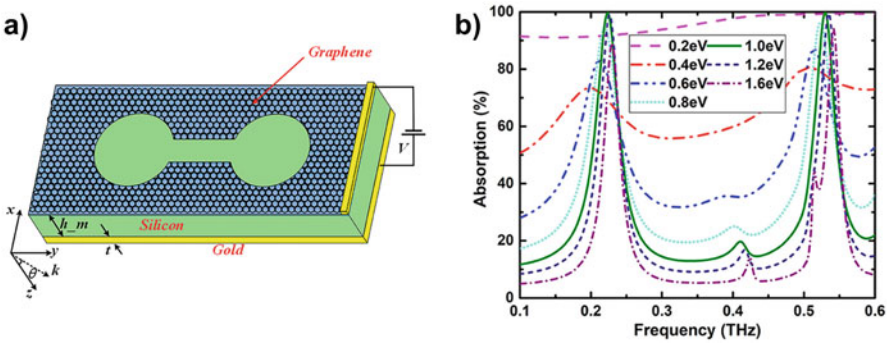

Fig. 4 (**a**) Unit cell of the THz perfect absorber, which includes monolayer graphene on a silicon substrate and a gold grounding plate. The radius of the dumbbell is 55 μm, h_m = 50 μm, and the periodicity in the x and y direction is 200 μm and 450 μm, respectively. (**b**) The absorption response with varying μ_C. (Reprinted from Ref. (Yan et al. 2020), © 2020, with permission from Frontiers in Physics)

simulations. There were three absorption peaks with absorption greater than 99%, namely 2.4 THz, 4.2 THz, and 7.2 THz, at various combinations of μ_C for the two graphene layers. The absorption peaks could be tuned from 1 THz to 8 THz. There were three absorbance modes, and each of the absorbance modes showed a decrease in absorbance as the angle of incidence was increased. The absorbance was greater than 80% for all the three modes for incident angles less than 45°. The MM could be considered dual-mode because at least two modes could exist together without changing the structure.

Microwave Range Applications

Microwave frequency region is essential for radar detection of aircrafts, ships, and other vehicles. Microwave radars are commonly used in commercial as well as defense applications. Hence, the development of surfaces capable of absorbing microwaves has been significant for a long time. Graphene MMs can provide lightweight tunable absorbers that cannot be developed using traditional materials. Balci et al. (2015) developed a microwave absorber using two large sheets of graphene electrodes sandwiched between a Polyvinyl chloride (PVC) support and an electrolyte. Therefore, the structure acted like a graphene capacitor which could be tuned electronically to absorb radar signals. On application of voltage between the graphene layers, the electrolyte gets polarized, and high mobility charge carriers are formed on the graphene layers, which are responsible for the response to the microwaves. The reflection from the electrolyte layer is very low due to the low mobility of ions. Graphene was grown using CVD on ultra-smooth and thin copper layers. The area of the prototype developed was 20 cm^2, and the device was flexible. TE polarized microwaves at a frequency of 10.5 GHz and power of 15 mW were incident on the absorber. The reflected and transmitted waves were measured using microwave detectors. Only 1.8% of the incident wave was reflected when there was

no bias applied. As ±3.5 V bias was applied, due to the change in μ_C, the reflection increased up to 20%. A similar response was observed for absorption, the minima 21% was at 0 V, and it increased to 45% on application of ±3.5 V bias. At 0 V bias, the resistance of the graphene layer was highest, which was also confirmed by the conductivity model; as bias was applied, the resistance decreased drastically. A suppression rate of 50 dB could be achieved at a bias of less than 5 V. Huang et al. (2014) developed a RAS using the copper structures, which were in the shape of the Hilbert curve (Fig. 5a). Graphene was deposited via CVD and was placed

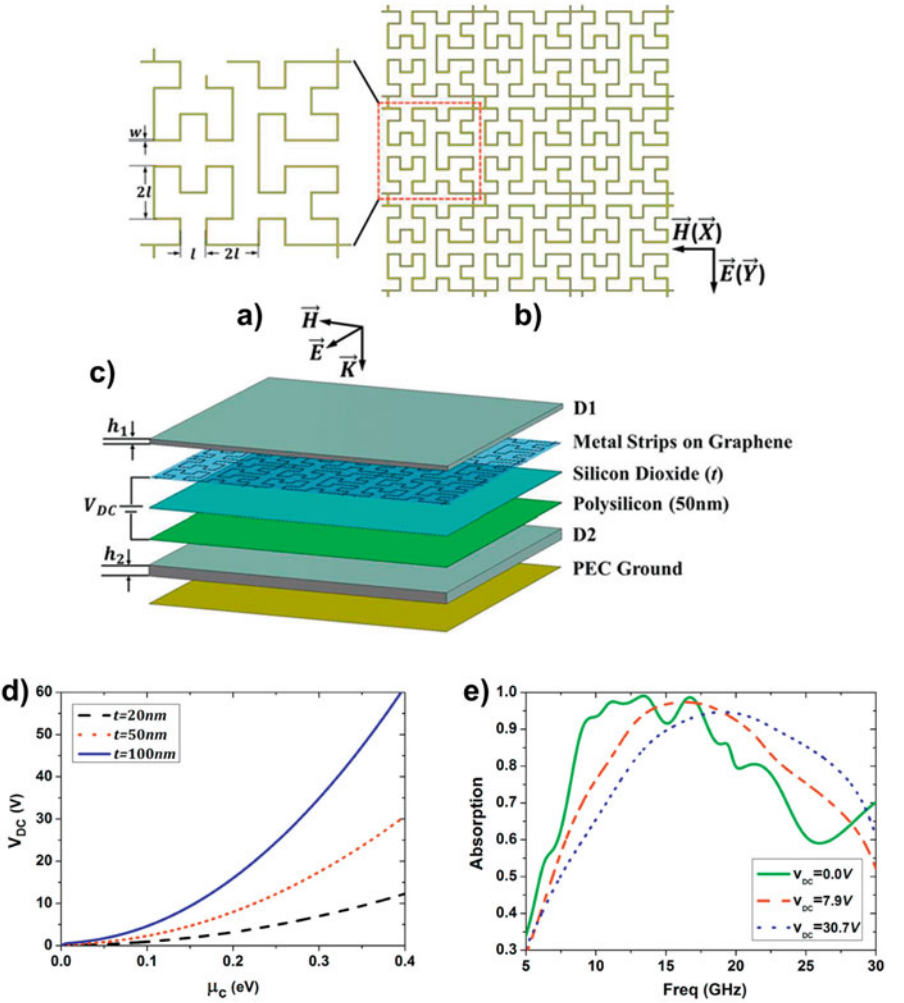

Fig. 5 (**a**) Hilbert curve unit cell and (**b**) array. (**c**) The layers of the unit cell of the FSS in which D1 and D2 are dielectric layers. (**d**) Variation of chemical potential (μ_C) and applied biasing voltage (V_{DC}). (**e**) Absorption response of the FSS to V_{DC} when t = 50 nm. (Reprinted from Ref. (Huang et al. 2014), © 2014, with permission from AIP Publishing)

underneath the Hilbert curve. Further, to obtain the tunability of the device, SiO_2 and polysilicon layers were placed under graphene, and the structure was grounded by a PEC layer (Fig. 5c). For −10 dB, the absorption bandwidth observed was from 8.9 GHz to 18.1 GHz at 0.0 eV μ_C. As μ_C was varied from 0.0 to 0.4 eV, the peak frequency could be tuned from 13.5 GHz to 19.0 GHz (Fig. 5e). The metallic structures caused multiple narrow-band resonances, while the tunability and further broadband range were provided due to graphene. Moreover, the graphene sheet increases the dissipation of the incident EM wave. The response of graphene was considered frequency independent in the microwave region. The polysilicon layer improved the angular range of the device, the RAS was simulated in CST Microwave Studio to optimize the dimensions. It was confirmed that the broadband range of the absorber was mainly due to the graphene layer and not by the two dielectric layers. The performance of the absorber varied when the angle of incidence was changed, absorption of greater than 90% was achieved from 0° to 20°. Further, till 50°, the absorption was about 90°, however, increasing the angle by more than 70° caused the absorption to go below 80%.

Cloaking

Cloaking is one other application of MMs where conventional techniques cannot be used. Although there was a massive hype regarding cloaking in the optical wavelengths when MM fabrication gained widespread attention across research communities, they are still in the development stage. Cloaking at the radio frequencies was effectively demonstrated using metals. However, as the applications of higher frequencies came to light, the same metals could not be used for cloaking due to their depreciated conductivity at higher frequencies and finite skin depth. Electronically induced transparency (EIT) is an effective method to cloak objects, wherein the EM waves bend around the object to be cloaked due to the presence of certain materials and/or structures. Conventional quantum EIT has extreme requirements, including stable optical pump and cryogenic temperatures. Hence, these structures have limited practical usage. Graphene MMs can deliver the required properties with less stringent conditions for fabrication. Moreover, the size of graphene-based MM devices can be much lower. Further EIT has several applications such as slow light control, biochemical sensing, filtering, switching, and communications.

IR Range Applications

Farhat et al. (2013) theoretically explained the cloaking of 3D objects in the far IR range using graphene layers that covered the object intended to be cloaked. The cloaking frequencies were tuned electrically by varying the μ_C of graphene by applying a bias voltage. Simulations were performed to confirm the cloaking over near and far-field regions using scattering cancellation technique (Alù and Engheta 2005); the dipole moment of the object was suppressed by the negative local polarizability due to negative permittivity. Further, the surface plasmon polaritons excited on graphene also helped to improve the cloaking. When using a single layer of graphene around the object, the variation of μ_C from 0.5 eV to 1.2 eV caused the dip of the scattering efficiency shift from 1.5 μm to 0.6 μm for close to zero, but

positive complex permittivity of graphene. For negative permittivity, a tunable response was observed, but the tunability was much lower than near-zero permittivity. However, the negative permittivity had better cloaking performance as the scattering dip was larger than near-zero case. The performance of graphene-based cloaks was much higher than simply plasmon-based techniques due to the lower intrinsic losses in graphene. Using the total electric field calculations, upon excitation at frequency 5.17 μm, it was visualized that the dipole excitation of the object was highly suppressed and considered invisible in the far field. Upon increasing the number of layers surrounding the object, the number of dips in the scattering efficiency was the same as the number of shells/layers surrounding the object. Hence, efficient cloaking could be achieved for various frequencies. The calculations were made for up to three layers of graphene.

THz Range Applications

Similar to the previous case, but in the THz region, Chen and Alù (2011) used analytical expressions to demonstrate the ability of graphene as an effective tunable cloaking layer. Initially, the calculations were performed for a 1D scenario, in which a graphene sheet was grown over a SiO_2 film. The graphene layer of μ_C 0.13 eV caused a dip in reflectance at 3 THz; as the μ_C was increased to 0.4 eV, the reflectance dip moved toward higher frequencies with a decrease in cloaking performance. Further, a cylindrical cloak was studied, and its performance was compared with a conventional mantle cloak constructed using a conducting material with a periodic pattern. The cloaking performance of the monolayer graphene was much better than the mantle cloak. However, compared to the 1D case, the cloaking performance deteriorated, but the tunability range of the cloak was improved. The cloaking performance was visualized by studying the near field electric field distributions for an infinitely long dielectric cylinder covered by a graphene layer (Fig. 6). Also, the conductive MMs can only control in-plane heat conduction, but radiative thermal

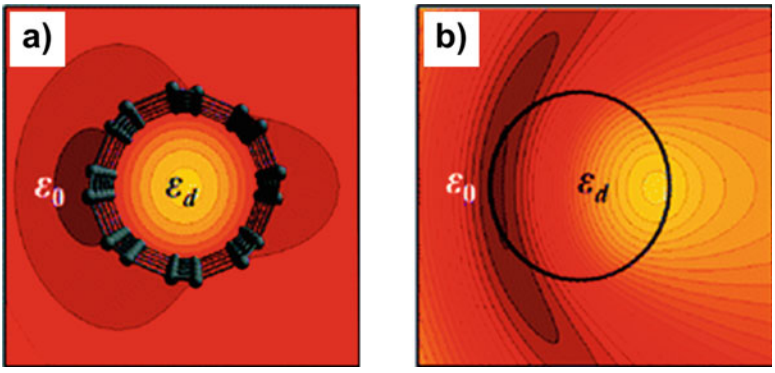

Fig. 6 Electric field distribution for an infinitely long dielectric cylinder (**a**) covered by a graphene cloak (chemical potential 0.51 eV) and (**b**) not covered by the graphene cloak for a normally incident wave of frequency 3 THz. (Reprinted from Ref. (Chen and Alù 2011), © 2011, with permission from the American Chemical Society)

MMs can control the out-of-plane emission (Song et al. 2020). Zhao et al. (2016) used patterned monolayer graphene to achieve transparency in the THz range. It was numerically studied that the transmission peak could be tuned by varying the μ_C of the graphene. The unit cell of the periodic array was composed of a rectangular dipole antenna and a wire with a pair of monopole antennas coming out toward the dipole antenna. All of the elements were constructed out of monolayer graphene. The wire acted as grounding to the antennas attached to it. The structure was placed on SiO_2 and doped silicon layers. The response of the monopoles was tuned by applying a bias voltage. The transmission response of the two structures (dipole and monopole antenna) revealed that the transparency peak at 3.4 THz of the whole MM was observed as a result of the destructive interference of the graphene dipole and monopole antennas, which acted as the bright and dark modes, respectively. As the μ_C was adjusted from 0.5 eV to 0.9 eV, the peak of the transparency displayed a blue shift and the transmission spectrum in the range 2–5.5 THz, gradually showed decreased spectral range of transparency; as two dominant peaks (for 0.9 eV) appeared instead of one (for 0.5 eV). Xiao et al. (2018) used a MM consisting of metallic structures on a monolayer graphene to achieve active modulation in the THz region. The system was analyzed theoretically using FDTD simulations. The unit cell consisted of an aluminum structure with lattice constants 80 μm and 120 μm in the x and y direction. The structure was a cut wire and oppositely oriented SRRs. The aluminum structure was placed on top of a layer of graphene. The device had a narrow transparency window at 0.65 THz; the cut wire acted as the bright mode resonator, which was canceled by the dark mode SRR structures. The electric field distribution of the structures revealed regions of high and low charges that canceled each other out in the cut wire and SRR structures that led to EIT. The transparency window was controlled by varying the μ_C of the monolayer graphene; as the μ_C was increased from 0.0 eV to 0.8 eV, the narrow transparency window was converted to a broadband absorption.

Due to MMs' excellent ability as absorbers, various graphene-based tunable MMs have been developed, which can be used RASs for stealth applications. The applications are spread in the IR, THz, and microwave frequency regions. The MMs presented consisted of simulations as well as some prototypes used as tunable absorbers. Moreover, an advanced stealth technique called cloaking was discussed; that used the EIT phenomenon to hide an object in the IR and THz frequency range. Graphene was integral to the development of these cloaking MMs.

Communication

Antennas

Antennas are used to capture EM signals which can contain important information. Efficient, small, and light antennas are crucial for defense and security applications. The radiation efficiency of plasmonic graphene antennas was increased in the THz region by introducing a MM structure (Amanatiadis et al. 2017). Radwan et al. (2015) observed that the antenna's bandwidth improved when SRR structure was

included as compared to without it. The bandwidth was increased even further when bias voltage was applied. Therefore, graphene-based MMs and FSSs can help design these new-age antennas. However, in some stealth applications, such as designing unmanned vehicles/weapons which have to be invisible to the radar is a complex task because its own antenna transmitting and receiving signals might interfere with the absorptivity of the object to the enemy's radar. FSSs can be used in antennas to prevent the detection of its signal from the radar by transmitting and suppressing only the desired frequencies. Apart from the transmitting frequency, the remaining signal is reflected away from the source. Independent tuning of the real and imaginary part of permittivity and permeability of the object is necessary to create the optimal response. Zainud-Deen et al. (2018) developed a tunable graphene-based MM antenna which achieved negative permittivity and permeability over the frequency range of 390–550 GHz. The MM unit cell comprised an SRR of width 18 μm with two gaps placed on a SiO_2 substrate and a PEC grounding layer. The reflection peak was shifted to a higher frequency as the μ_C was increased from 0.3 eV to 0.8 eV. The increase of the gap between the SRR resulted in decreasing the operating band of the MM. A 13 × 13 reflectarray was built and illuminated with a circular horn 3.9 mm away from the array to demonstrate the antenna. Various different configurations were analyzed, and a maximum gain of 22.6 dB was achieved. Seyedsharbaty and Sadeghzadeh (2017) simulated a graphene-loaded tunable patch antenna with enhanced gain in the THz band. Pyrex was chosen as the substrate for the MM, since it has lower permittivity than SiO_2, leading to higher gain. The complex structured unit cell was placed at a distance of 174 μm above the graphene layer (Fig. 7a). Upon the introduction of the MM, the return loss was significantly reduced, and the bandwidth was increased. As the μ_C of graphene was increased, the resonant frequency was shifted, and gain was increased considerably (Fig. 7b). Furthermore, the efficiency was improved by more than 100% compared to the antenna without MM. Liu et al. (2018d) simulated a microstrip antenna

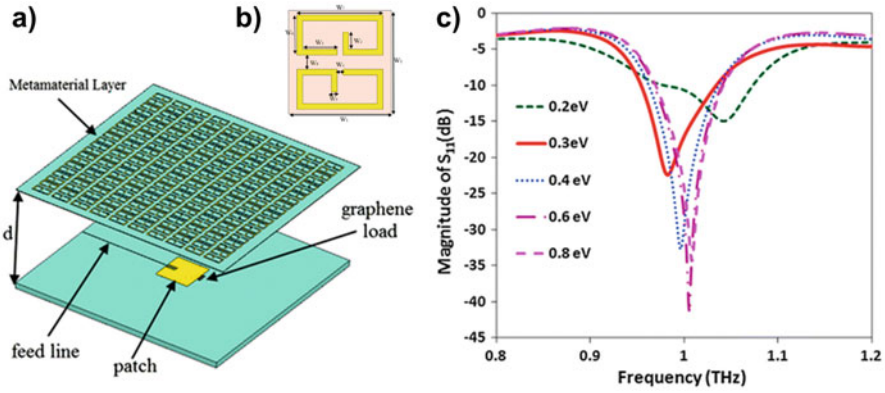

Fig. 7 (**a**) The MM patch antenna with a graphene load and (**b**) the complex MM unit cell. (**c**) MM antenna return loss for various chemical potentials of the graphene layer. (Reprinted from Ref. (Seyedsharbaty and Sadeghzadeh 2017), © 2017, with permission from Springer Nature)

designed using graphene stacks array. The structure consisted of alternating layers of graphene and Al_2O_3, which were stacked to form a cross shape with empty space in the middle square to form the unit cell of the array. The structure was built in SiO_2, which was placed on top of a silicon substrate and a copper film that grounded the system. Two resonant frequency bands were observed in the reflection coefficient plot, which covered the range of 2–5 THz. As the μ_C was increased from 0.1 eV to 0.5 eV, the main resonant band shifted from 2.8 GHz to 3.4 GHz. The sharpest response was achieved at μ_C 0.4 eV. In the microwave region, Kumar et al. (2016) developed a triple band MM-based antenna, which reached the highest gain of 3.98 dB at the center frequency of 6.98 GHz. A prototype was built whose electrical size was 19 mm^2. The unit cell of the MM comprised of an interdigital capacitor and a modified SRR structure. The size of the antenna was reduced by 49.42% as compared to the conventional patch antenna. Simulation was used to optimize the geometric parameters of the MM. Huang et al. (2016) simulated a beam scanning antenna that consisted of a straight dipole antenna placed 9 μm away from two graphene 1D arrays. The two arrays were placed on top of a SiO_2, silicon and finally a metallic grounding layer. To achieve a μ_C of 0.4 eV, a voltage of 60 V had to be applied when the thickness of SiO_2 was 100 μm. However, this requirement drastically decreases as the thickness decreased. The effect of tunability on the reflection coefficient was considered negligible due to the presence of the straight dipole antenna. However, the beam scanning performance of the antenna was significantly improved due to the tunability of the graphene arrays.

Data Storage

The vehicles, missiles, outposts, etc., used by the defense sector contain various sensors which deliver critical information for the onboard computer/personnel. Hence, fast and efficient communication technology that includes large storage can improve the performance of manned as well as unmanned defense applications. Orbital angular momentum vortex waves have gained importance in the literature due to their potential use in increasing the capacity of wireless communications. Compared to plane waves, the orbital angular momentum vortex waves have an azimuthal phase of $\exp(il\phi)$, where l is the topological charge, and its pointing vector occupies a spiral path. Hence, these waves have limitless orthogonal eigen states (Meng et al. 2019). A transmitarray antenna was developed by Meng et al. (2019) to create these vortex waves when the array was excited by a horn antenna. The transmitarray antenna was made using graphene-based MM. The unit cell of the periodic array consisted of four layers of graphene-quartz-graphene separated by air of thickness 10 μm; the thickness of quartz was 5 μm. The μ_C of the top and bottom layer graphene was varied from 0 eV to 1 eV. A frequency range of 4.2–5.6 THz and a transmission phase range of 360° was achieved. Further, the electrical tunability of graphene helped to manipulate the vortex waves; different modes could be observed over a frequency range of 4.2–5.6 THz. Polarization multiplexing enables the increase of the amount of data to be stored and transmitted. However, conventional techniques to obtain a specific polarization including, grating and waveguide-based, cause nonspecific polarization EM energy to the surrounding electronics. The

undesired EM energy can cause a disruption of the performance of the electronics. Hence, photonic crystal-based techniques are preferred as the undesired polarizations are eliminated. However, the photonic crystal-based polarization filters are hardly present in the microwave region of the EM spectrum. Hence, graphene MM-based solution was developed by Zhang et al. (2021), which used anisotropic properties to develop a device that acts as both a polarization filter and an absorber of unwanted EM waves. There were two types of unit cells used, the top layer for one was an I-shaped graphene layer on a substrate, and the other was simply the non-shaped substrate. The middle and bottom layer for both was a graphene SRR and graphene film, respectively. The transmission line theory was used to study the response of the designs. Ohmic losses were responsible for dissipating the TE waves, while the response of the TM mode could be adjusted by the arrangement of the two types of unit cells. The absorption of the TE waves was greater than 90% from 8.0 GHz to 14.2 GHz. The polarization filtering and absorption of the incident waves were demonstrated by performing simulations. Three examples were performed in which the unit cells were arranged in an 8 × 8 array. In the first example, the two types of unit cells were randomly arranged, it was observed the TE component was dissipated, and the backward reflection of the TM mode was highly suppressed. In the second example, a linear polarized wave was normally incident on the array which had the cells arranged in alternating columns of the two types of unit cells (Fig. 8). The incident wave was split into two main beams in the TM component, while the TE mode was a small single beam. When the array was arranged in the chessboard type pattern, the TM mode was split into two major beams, and the TE mode was again a single small shape. In the final example, a circularly polarized beam was incident on the array. Similar to the linearly polarized example, the TM mode was split into certain major beams, and the TE mode was a single and small major beam. Experimental variation was also performed, in which the graphene film was fabricated by reducing graphene oxide.

Slow Light

The phenomenon of EIT could be used for the development and advancement of slow light devices. Slow light is the phenomenon of reducing the group velocity of an optical carrier. It is being used to develop electronics to drastically reduce energy consumption and reduce noise in communication technologies. Liu et al. (2018a) developed an FSS whose unit cell consisted of two types of cut wire metallic structure with graphene strip of width 7 μm between them (Fig. 9a). Electrical tunability of graphene was used to tune the performance of the FSS. Details of the fabrication process which could be followed to practically construct the device were explained. Theoretical calculations were performed on CST Microwave Studio, and frequency range studied was 0.3–1.8 THz. Destructive interference between the two cut wire structures caused the EIT, which was established by analyzing the cut wire structures separately. Upon the introduction of graphene strip of μ_C 0.05 eV, the narrow transparency peak was almost at the same position. However, there was a slight decrease in peak transparency. As μ_C was varied from 0.05 eV to 0.35 eV, the peak of the transparency shifted to a slightly higher frequency and there was a drastic

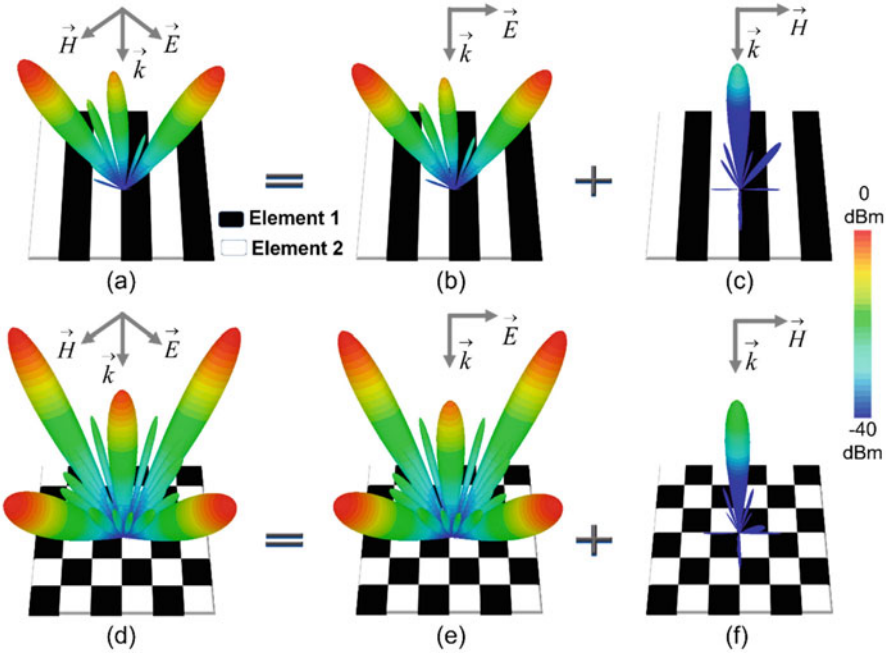

Fig. 8 (**a**, **d**) Simulated 3D scattering of the MM filter with two types of unit cells on which a 45° linear polarized wave of 10 GHz is incident. (**b**, **e**) The TM component and (**c**, **f**) TE component. (Reprinted from Ref. (Zhang et al. 2021), © 2021, with permission from Elsevier)

reduction in peak transparency (Fig. 9b). As shown in Fig. 9c, d, the surface currents in the graphene strip are strongly activated when μ_C is 0.4 eV, which causes some constructive interference, therefore, reducing the EIT. However, when μ_C is 0.05 eV, there are hardly any surface currents in the graphene strip. The slow light characteristics could be manipulated via the μ_C of graphene. The group delay was decreased from 3.6 ps to 1.4 ps as the μ_C was increased from 0.05 eV to 0.35 eV, which was detrimental to the slow light performance. Similar simulations were performed by the same group (Liu et al. 2018c) for studying the slow light response in the frequency range 1.0–2.4 THz; using an FSS which had a U-shaped metallic ring placed at the center of two metallic cut wires with a graphene strip located below the middle of the ring. The structure was placed on a SiO_2/silicon substrate. EIT was achieved through the destructive interference of the bright mode of the U-shaped ring and the dark mode of the cut wires. The electric field response of the structures was analyzed to attain this conclusion. The peak of the transparency was at 1.74 THz, and upon the introduction of the graphene strip of μ_C 0.05 eV, the peak was suppressed drastically. The graphene strip caused electric field hotspots to be created on its edges which led to a decrease in the transparency peak. As the μ_C was increased from 0.05 eV to 0.15 eV, the resonance frequency was red shifted; however, as the μ_C was increased to 0.20 eV, the EIT nature of the FSS disappeared.

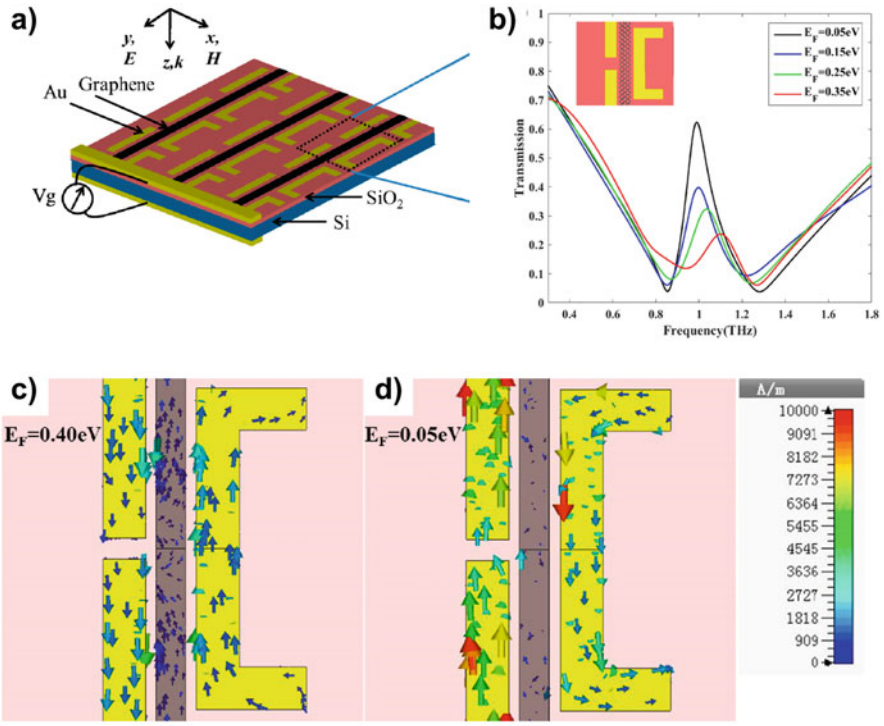

Fig. 9 (**a**) Design of the graphene-based FSS with a periodicity of 80 μm in both x and y direction. (**b**) Transmission spectrum with varying μ_C and the inset shows the unit cell of the FSS. (**c**) Surface current distributions of the unit cell at the transmission peak with $\mu_C = 0.4$ eV and (**d**) $\mu_C = 0.05$ eV. (Reprinted from Ref. (Liu et al. 2018a), © 2018, with permission from Optical Society of America)

The transparency peak was suppressed as the width of the graphene strip was increased and disappeared as the width became greater than the U-shaped ring. The transparency peak and group delay could be tuned via the μ_C of the graphene strip. As the μ_C was increased from 0.05 eV to 0.20 eV, the group delay was decreased from 3.3 ps to 0.5 ps which depreciated the slow light performance. Jia et al. (2019) developed a transparency-based thermally tunable graphene-based FSS. The unit cell consisted of a patterned graphene structure containing two hollow squares with a rectangular strip in the middle. Both the hollow squares and the rectangular strip acted as the bright modes, which coupled create the transparency. The temperature and polarization variation of the FSS was studied for the application in a slow light device. The Boltzmann transport theory was used to explain the interaction of electrons and phonons in graphene. At 60 K, the broadband transparency peak could be observed at 10.2 THz. The intensity and bandwidth of the transparency window decreased as the temperature was increased from 60 K to 500 K, however, the position of the window remained constant. The aforementioned phenomenon was due to the weak electron-phonon interaction and decreased

thermal loss at lower temperatures. The numerical results were validated by comparing with the Lorentz coupled oscillator theory. The number of dips in the transmission spectrum varied upon changing the polarization angle, hence, the FSS was deemed to be polarization sensitive.

Data Security

Apart from faster and large data transfer, "Secure" communication is of utmost importance to the defense and security systems. If sensitive information is leaked, it can lead to disastrous consequences. The use of THz for ultrawide bandwidth range by using modulators and demodulators can help prevent the theft of sensitive information using MM-based encryption systems. Although the time for signal processing increases, the chances of detection are very low. The enemy detecting the signal would have the correct code to read the signal; otherwise, the signal will be interpreted as noise (Federici and Moeller 2010). The THz beams are highly directional, undergo less scattering, and have limited propagation due to atmospheric attenuation. Hence, they are used for the encryption of beams over a wide channel spectrum. Further, the THz application range is critical because it overcomes the demerits of both microwaves (less directional) and IR waves (more directional but easily scattered in the atmosphere). Further, an autonomous EM spectrum management system is much more secure as compared to manual. Holography enables recording and reconstruction of wave information; it contains the patterns of intensity and phase variation which recreate the image (Huang et al. 2018). The need to securely transfer large data for communication, security, data storage, and 3D display can be achieved via holography (Hafiz and Qamar 2020).

Momeni et al. (2018) developed a reprogrammable graphene-based encrypted FSS for application in the THz frequency range. The unit cell consisted of a Jerusalem patterned graphene placed on an alumina/silicon/quartz substrate grounded by a gold layer. On application of bias to this unit cell, the cells in the array could be programmed as "0" or "1" states. The electrical tunability of graphene enabled it to switch between "0" and "1" based on reflectivity in the frequency range, 1.1–1.7 THz. Hence, encryption code could be quickly changed by reprogramming the FSS. The far-field response of the FSS was studied as a combined response of the individual array units using a 2D inverse fast Fourier transform. The FSS was programmed so that the secure information being sent was convolutionally encoded such that in the far-field region, the original message beams were replaced as parity beams which contained additional information masked around the original message (Fig. 10). Hence, the message cannot be read without decoding the parity beams. The message was initially directly written on the FSS and repeated in each row of the coding matrix. Next, five memory registers were used to phase encode the message, therefore, programming the FSS with the desired coding to generate the parity beams. Hence, sensitive information transfer in free space using THz waves was successfully demonstrated.

Taking a step further, graphene nanoribbon-based arrays were used to write an FSS hologram based on polarization multiplicity and re-writability by Rajabalipanah et al. (2020). The phase shift when an x or y polarized wave was

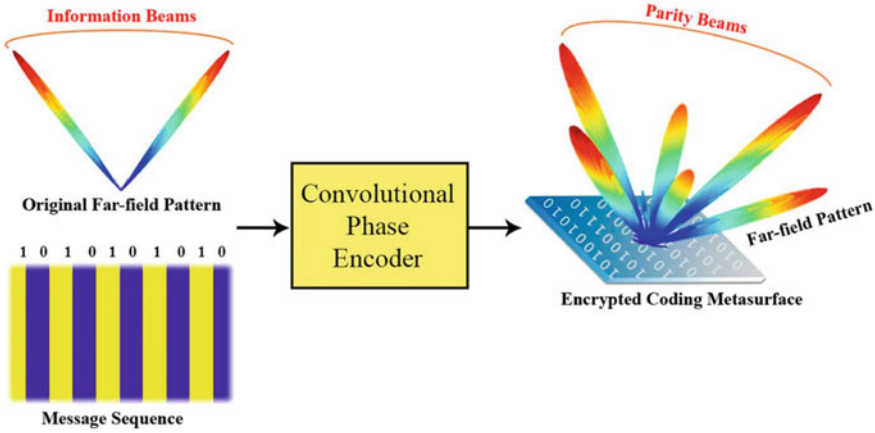

Fig. 10 Basic mechanism of the secure information transfer by the FSS. The FSS is encrypted using a convolutional encoder which modifies the original beam to be converted to a parity beam whose contents are hidden but can be recovered when decoded. (Reprinted from Ref. (Momeni et al. 2018), © 2018, with permission from Springer Nature)

incident on the graphene array was due to the propagation delay imparted by the plasmon waves activated on graphene. The resonant frequency was inversely proportional to μ_C of graphene; therefore, the working frequency of the device could be tuned. Further, a selected phase difference could be imparted by applying an appropriate bias. Therefore, anisotropy could be imparted to the FSS as the reflection phases of each polarization could be controlled by independently varying the μ_C of the x and y direction, respectively. For each polarization, a 45° phase shift was extracted to build a 3-bit graphene-based hologram; which improves the holography performance by reducing the coupling of the target unit cell with the surrounding unit cells. The 3-bit information was programmed into the FSS by applying a bias to each unit cell in the array; the position of the unit cell was defined by two digital codes. The two polarizations could produce two different images independently at the near field channels using the weighted Gerchberg-Saxton algorithm (Zalevsky et al. 1996). The device was theoretically studied by CST Microwave Studio. The imaging efficiency, signal-to-noise ratio, and root mean square error was analyzed. To demonstrate the encryption capability of the metasurface, a four-letter word was securely communicated (Fig. 11). A table with 64 elements was used to identify the position of each letter of the word. Next, the table's contents were scrambled in a complex manner such that the word could be only decoded by a person having the correct key. Applying appropriate bias to the metasurface, the coded image could be built on the metasurface using low and high intensity for "0" and "1," respectively. To further secure the message, it was encoded by a convolution algorithm. Applying this security layer, the message could be protected even if the key to the table was compromised, because generator polynomials would still be required to decrypt the word. Further, two messages

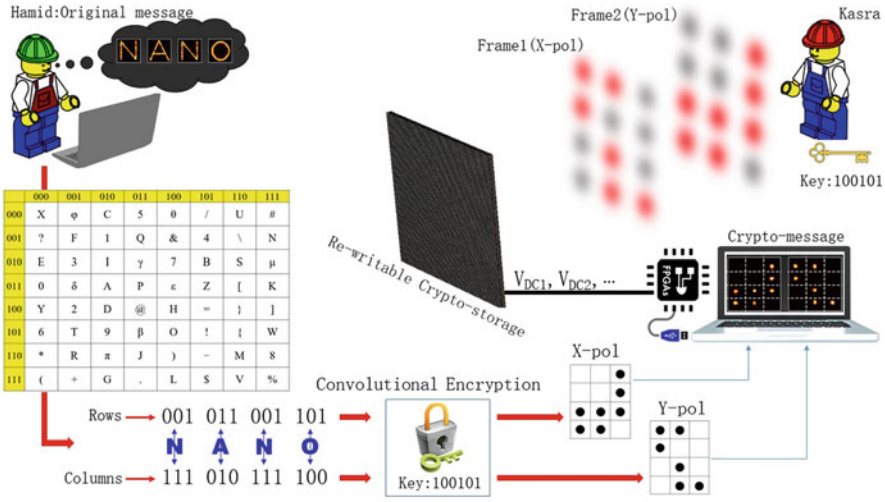

Fig. 11 The working of the FSS hologram based secure communication system. A four lettered word was defined using a 64 element table with a secret key. The information was encrypted further by using a convolutional encoder. (Reprinted from Ref. (Rajabalipanah et al. 2020), © 2020, with permission from De Gruyter)

could be sent at a time as the message in each polarization, and could be separately controlled. Conventional encrypting schemes can only transfer one letter at a time; the graphene-based FSS was demonstrated to send two four-letter words simultaneously. Therefore, the metasurface could transfer a larger amount of encrypted data in a quick and reliable fashion in THz frequencies. The device could also be used for beam manipulation and optical vortex generators.

Fast, efficient, and secure communication is of the utmost importance for defense and security systems. Graphene-based MMs have unique properties in the THz and microwave region, rarely found in conventional materials. Therefore, they are used to develop highly efficient and lightweight (because of the use of subwavelength structures) antennas, high capacity data transfer systems, high quality and low energy consuming systems based on slow light and secure communications using holography and encryption.

Imaging

THz imaging is an emerging research area that has applications in non-destructive evaluation. Security imaging is an important application finding commercial attention. THz-based imaging systems can help scan personnel and objects for unwanted objects dynamically in a safe and accurate manner. The unique properties of graphene MMs can be used to create hyperlenses, lenses, mirrors, and modulators for the next-generation THz imaging systems.

Beam Manipulation

The manipulation and detection of THz using conventional methods are complicated and usually requires strict management of environmental conditions. THz imaging was experimentally demonstrated using graphene arrays (Ullah et al. 2020); the THz waves were modified using MMs, such that the spherical waves can be converted into planar waves by changing the structure of the MMs. Zainud-Deen et al. (2017) used a graphene MM array to modulate the transmitted THz waves. The unit cell consisted of a multilayer graphene SRR with two gaps built on a SiO_2 substrate. The layers of the structure were separated by air gaps in between. The frequency range in which the MM was studied using simulations (using the finite integral technique) was 0.74–0.94 THz. The MM was used to convert EM waves into a directed beam. The transmission coefficient phase and magnitude were studied for two, three, and four layers. Three layers were deemed to be the optimal value, while 1 μm air gap displayed the best performance. The design frequency varied from 0.83 THz to 0.86 THz as the μ_C was increased from 0.4 eV to 0.8 eV. A 13 × 13 array of the tunable MM was simulated to understand the practical working of the lens. At 0.5 eV μ_C, a gain of 19 dB could be achieved. Hosseininejad et al. (2019) developed a flat reflective MM-based mirror to focus THz waves in the near field region. The tunability via μ_C of graphene helps to change the position and depth of focus of the mirror. The unit cell consisted of two layers of graphene separated by a high-density polyethylene (HDPE) layer; the structure was placed on a silicon substrate that was grounded. The two layers offered greater flexibility to the response of the mirror as μ_C of each layer could be controlled separately. The system was simulated in CST Microwave Studio, and the results were validated by comparing with a theoretical approach. The mirror was deemed capable of performing 3D confocal imaging by developing appropriate algorithms. Each unit cell was controlled using two bits, and the length of the square surface was considered nine times the incident wavelength. Hence, the mirror was able to focus at the desired location and depth with an error of less than 6%. Luo et al. (2019) designed a graphene-based tunable negative refractive index MM antenna for dynamic beam tilting in the THz range. The unit cell consisted of a metallic structure with embedded graphene placed on SiO_2 and p-type silicon layers grounded by a PEC. The source of the THz waves was a rectangular microstrip antenna with a central frequency of 4.26 THz. It was observed that when there was no MM, the gain was 5.9 dB and the main beam was at transmitted 0°. When one MM layer with μ_C of 0 eV was introduced in front of the rectangular antenna, the angle of the main beam was −8°, and the gain increased by 8.6%. As the μ_C was increased to 0.5 eV, the angle was increased to −11°. Furthermore, when two MM layers with μ_C of 0.5 eV were introduced, the gain increased by 10.4%, and the angle further increased to −27°. Therefore, the steering angle was modulated.

Yao et al. (2018) simulated a MM lens tunable from 4 THz to 6 THz, and adjustable focus (up to 10 μm) using graphene ribbons. The lens was designed to incorporate rows of graphene ribbons (subwavelength dimension) in an ion gel on top of a dielectric and a metallic mirror. The focal length for 5 THz wave was 96 μm

and varied by 10 μm when the μ_C was changed from 0.2 eV to 0.6 eV. Liu et al. (2018e) developed a tunable THz MM lens whose structure consisted of rectangular apertures of gold strips arranged vertically and having various lengths and rotations. The gold was covered by monolayer graphene deposited via the CVD method. The THz focal plane imaging system was used to characterize the lens. The frequency of the incident right hand circularly polarized (RCP) wave was 0.75 THz. The focal length of the lens was 10.46 mm under 0 V (no applied bias) and 12.24 mm when bias was 2 V (shift of 4.45λ). Furthermore, the modulation speed of the lens was estimated to be 30 MHz. Similar to the previous lens, Hamzavi-Zarghani et al. (2018) designed a tunable graphene MM lens for a circularly polarized wave that could be tuned to both converge and diverge the wave. The unit cell consisted of two graphene SRRs placed opposite to each other, between a dielectric substrate. The unit cells, when arranged appropriately by changing the rotation angle of the SRRs normal to the surface, the lens arrangement could converge the incident wave. The rotation angle could be changed to obtain an opposite response. This was demonstrated by simulating 18 unit cells in CST Microwave Studio. The working range of the lens was shifted from 21 THz to 19.63 THz by varying the μ_C from 1.6 eV to 1.4 eV.

Hyperlens

The natural diffraction limit hinders the resolution achieved in THz imaging, and scanning near field THz microscopy is used to overcome this problem. However, the process is very slow and expensive; MMs offer a simpler alternative due to their unique properties. Hyperlens can be used in microscopy to enhance an image beyond the subwavelength scale by converting evanescent waves to propagating ones. Another application of hyperlenses is power concentration. Graphene-based MM was used to demonstrate a THz hyperlens by Andryieuski et al. (2012). The unit cell consisted of a graphene wire, which was placed inside a dielectric (TOPAS), consisting of a number of graphene layers. The reflection and transmission of the hyperlens for various incidence angles were calculated to study the lens's performance. The simulations were performed to demonstrate the working of the hyperlens in vacuum. Two sources separated by one-fifth of the incident wavelength were successfully separated and displayed two peaks when using the hyperlens (Fig. 12). However, when graphene wires were removed from TOPAS, only a single peak was observed. The working frequency of the hyperlens was about 6 THz. The simulations were performed in CST Microwave Studio and COMSOL; both showed similar results, and the slight variation was due to the meshing techniques of the software.

Modulators

Modulators are used for imaging, beam shaping, adaptive optics, electro-optic switching, and hyperspectral imaging. Conventional liquid crystal-based modulators suffer from low modulation speed in the IR region and beyond. Various graphene-based modulators are being developed by research groups for high-speed thermal imaging (Fan et al. 2017; Hajivar and Hosseini-farzad 2019), which is a crucial

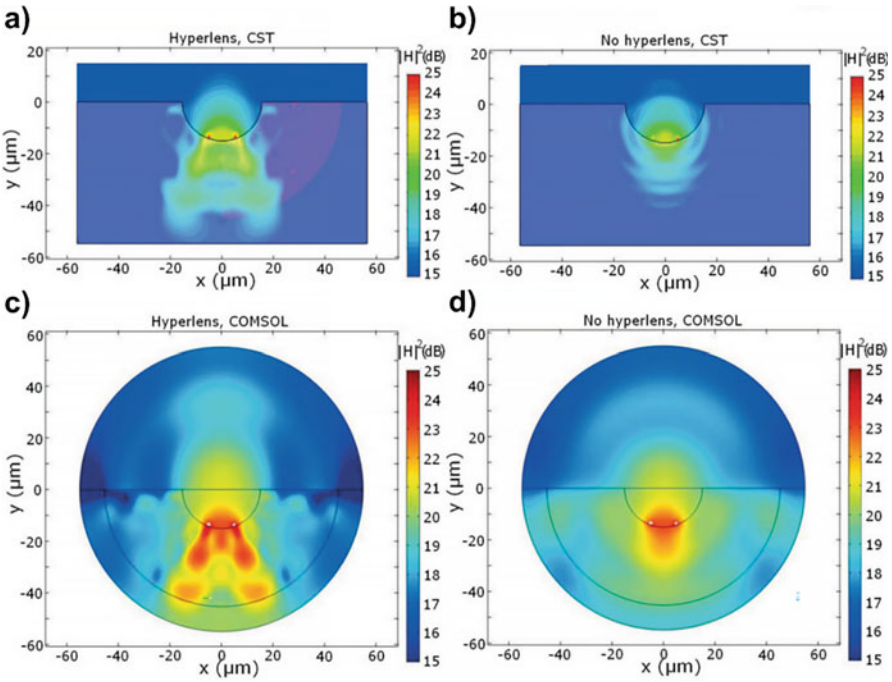

Fig. 12 Two magnetic line sources separated by 10 μm and emitting TM polarized waves were simulated to be incident on (**a**, **c**) hyperlens and (**b**, **d**) dielectric (TOPAS). The simulations were performed in (**a**, **b**) CST Microwave Studio and (**c**, **d**) COMSOL. (Reprinted from Ref. (Andryieuski et al. 2012), © 2012, with permission from the American Physical Society)

defense application. Moreover, the IR-based systems find applications in navigation and medical detection.

IR Range Applications

Yao et al. (2014) built a tunable mid-IR modulator using optical antennas built on graphene. The optical cavity creates a perfect absorber due to the critical coupling condition. The modulation depth of 100% was recorded when the applied bias voltage was changed from 0 V to 80 V. The working range of the modulator was 5–7 μm. The periodic structures and graphene were placed on a dielectric (AlO_x) and an aluminum grounding layer. The graphene monolayer was fabricated using CVD, and the dielectric thickness was 300 nm, which was optimized using the FDTD simulation results. The structures were deposited using electron beam evaporation and consisted of 10 nm palladium and 30 nm gold, the gaps between the structures were 40 nm wide. The absorber was considered ultrathin, and it was stated by the authors that it was the thinnest absorber with more than 95% absorption at the time of publication. The response time of the absorber for tuning was less than 10 ns, which was also confirmed experimentally. The previous modulator had a very high working voltage. Zeng et al. (2018) experimentally demonstrated an IR modulator that

operated at GHz speeds but used low voltage. The modulator achieved modulation depth up to 90% while working at room temperature. The unit cell consisted of I-shaped patterned monolayer graphene on top of an ultrathin dielectric layer (Al_2O_3), amorphous silicon, followed by a grounding layer of gold. The gate voltage required to tune the μ_C of graphene depended upon the thickness of the dielectric layer; therefore, the gate voltage was low due to the ultrathin layer of Al_2O_3. The light incident on the structured array was trapped inside the cavities, and a combination of absorption by the structure and destructive interference leads to a very low reflection over a certain wavelength window. The dip in the reflectance shifted from 8.3 μm to 7.3 μm, and the reflection minima increased from 5% to 16%, as bias voltage was changed from −3 V to 7 V. The plasmonic response was visualized using near field imaging of the unit cell while applying bias to the FSS. The trapping of the light could be seen near the resonance wavelength. The switching speed of the modulator was 7.2 GHz, which was the fastest in the mid-IR range using an electrical approach at the time of publication. To demonstrate the application of the modulator, a 6 × 6 array was used to selectively reflect the EM waves in the mid-IR region; the frame rate achieved was 23 kHz, which was much higher than the commercially available cameras.

THz Range Applications

Sensale-Rodriguez et al. (2013) developed experimentally and theoretically a modulator for video-rate THz imaging. The unit cell consisted of monolayer graphene grown via CVD and transferred to SiO_2/p-type silicon substrate. To demonstrate the capabilities of the device, a 4 × 4 square array was prepared. Graphene was chosen because of the very high switching speeds. Each pixel was controlled separately via titanium/gold contacts and grounding layer. Reflection modulation was selected due to its higher transmission depth as compared to transmission mode. The modulator was used to construct an object by turning off and on the desired unit cells. Therefore, the incident waves were reflected off only from the particular unit cells which help form the object. The very high switching speed of graphene enables the modulator to achieve a speed of 24 frames per second. More than 50% reflectance was achieved with a modulation depth close to 100%. Electrolytic grating improves the conventional dielectric modulators because of the presence of higher charge densities. Malevich et al. (2020) designed a modulator consisting of a 16 × 16 array integrated on a PCB (Fig. 13a). Each pixel contained an electrolyte sandwiched between graphene ribbons grown using CVD (Fig. 13b). Bias was applied between the corresponding rows and columns of the array, which was controlled externally. The transmission response of the modulator was studied using a THz source and camera. Upon application of bias across the arrays, the transmission of the EM waves could be controlled. The transmission response and tunability of a single unit cell were studied. As the bias was increased from 0 V to 4 V, the transmission decreased drastically over a broadband range (Fig. 13d). However, bias voltages greater than 3 V caused hysteresis and slow response. The relaxation time increased drastically (20 ms to 0.7 s) as the bias was increased from 0.7 V to 2.0 V. The modulator could operate with switching speeds up to 1 kHz. The modulator was able

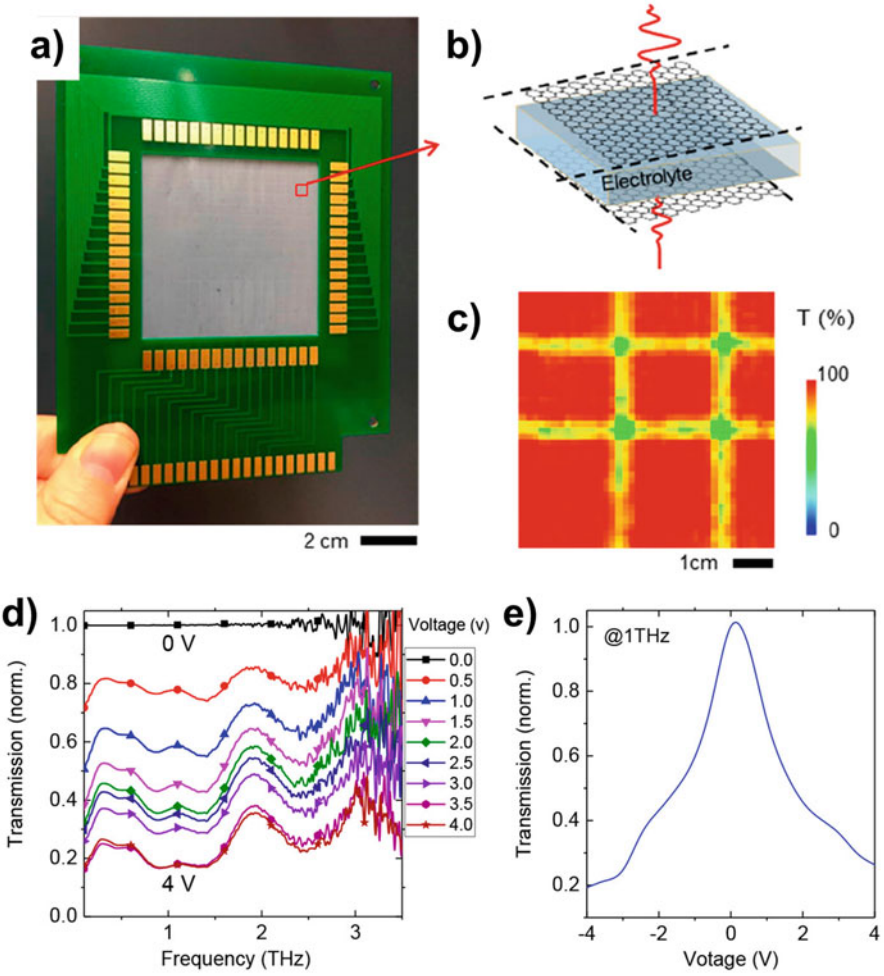

Fig. 13 (**a**) Prototype of the THz modulator, (**b**) the unit cell consisted of an electrolyte sandwiched by graphene nanoribbons. (**c**) Image obtained at 0.1 THz with two rows and columns biased at 1 V and −1 V, respectively. (**d**) Transmission spectrum of the modulator with varying bias voltage. (**e**) Transmission response to the applied bias voltage at 1 THz. (Reprinted from Ref. (Malevich et al. 2020), © 2020, with permission from ACS Publications)

to generate patterns with a high frame rate. This was experimentally confirmed by passing THz waves through two patterns (cross and two lines) in front of the modulator. The modulator was able to correctly recreate the two patterns.

Microwave Range Applications

Further, in the microwave frequency range, Li et al. (2017) developed a modulator with tunable reflection using a MM whose unit cell consisted of a gold graphene structure placed on SiO_2/silicon/air substrate and grounded by aluminum. The MM

showed good reflectance performance in the range of 25.3–38.2 GHz. The electrical tunability of graphene was used to vary the reflection amplitude from 78% to 26%.

Chiral Imaging

Some compounds relevant to biological activity, such as glucose, possess chirality in the optical region, which can help quantify these compounds. However, conventional spectroscopy requires highly sophisticated and expensive instruments. A planar MM-based lens was demonstrated to successfully map the exoskeleton of a beetle which displays chiral properties (Khorasaninejad et al. 2016). Polarization manipulation in the THz region by chiral MMs is being looked into for small and lightweight devices. Chiral FSSs alter the polarization state due to their unique properties. On exposure to optical pumping, Graphene can alter its properties; therefore, a tunable chiral MM can be developed. Single-layer graphene does not interact strongly with optical pumping; however, multiplayer graphene can be used to manipulate its THz response. Masyukov et al. (2020) developed a chiral MM whose polarization states could be dynamically controlled. The unit cell consisted of a combination of multilayer graphene and gold, which was arranged in the shape of a chiral petal. The gold and graphene layers mirrored each other and were sandwiched in a TPX substrate. The working frequency of the MM was 0.5–0.9 THz. The system was simulated by employing the temporal coupled-mode theory. The simulation output was validated by comparing it with analytical results. The resonant frequency of the MM to the left-handed circularly polarized (LCP), and RCP waves depended on the optical pumping power. Up to 0.65 THz, there was almost no effect of the optical pumping power on the transmission response. However, the transmission was close to zero for the RCP wave at 0.84 THz for high optical pumping power. Therefore, at this frequency, only LCP waves can travel through the MM. The polarization sensitivity was confirmed by calculating the ellipticity angle spectrum, +45° angle corresponds to RCP wave and −45° pertains to LCP wave, and 0° represents a linearly polarized wave. Li et al. (2020) also developed a chiral MM, which had opposite absorption responses to LCP and RCP waves. The μ_C of graphene was varied to modulate the response of the MM. The design showing chiral dichroism was optimized using the advanced Jones matrix theory. Two structures were proposed, which were mirror images to each other. The structures consisted of a split ring pattern made out of aluminum on a polyimide substrate. The ring also contained an aluminum strip at a particular twist angle that was different for both structures. A graphene window was added to the structure to tune the response of the MM. It was reported using simulations that under normal incidence, one of the structures absorbs most of the RCP waves and the other absorbs most of the LCP waves. The EM energy absorption in the structures was due to ohmic losses. The varying surface currents in the structures were responsible for the opposing response to RCP and LCP waves. There was no significant spectral shift upon changing the μ_C from 0.1 eV to 1.0 eV. At 1.02 THz and graphene width 4 μm, the modulation range was between 39% and 77%. Further, the MM was simulated to validate its use in dynamic imaging. A 16 × 16 array was created, there was an opposite response of the two types of structures to the LCP and RCP waves. The response at μC 0.1 eV

and 1.0 eV was compared, it was reported that 1.0 eV had higher contrast to the image obtained due to the higher modulation depth.

THz imaging is an essential application for the future of defense and security systems. Therefore, THz lenses and mirrors are required, which, as previously established, can be effectively accomplished using graphene MMs. Further, hyperlenses can improve the resolution of THz imaging systems. Modulators based on graphene in the IR, THz, and microwave frequency range can achieve high switching speeds for faster frame rate imaging and video.

Conclusions

The defense and security systems are at the forefront of developing new and exciting technologies, which give a nation not only the scientific edge but also an upper hand in foreign policy and diplomacy. Due to their structure and the selected materials, MMs have unique applications in which conventional materials cannot be used. FSSs are flat MMs with thickness less than tens of the working wavelength. The shape and the dimensions of the structures of MMs greatly influence the response of the system. Moreover, the subwavelength structures can reduce the size of the component required for certain applications. The evolution of fabrication techniques such as lithography and nanoimprinting has paved the way for cheaper and accurate periodic structures. Therefore, making MMs a viable alternative for new-age applications.

Graphene, due to its excellent properties, like high electrical and thermal conductivity, high strength, and zero bandgap, can improve the response of MMs even further. Graphene-based MMs have defense and security-based applications in the IR, THz, and microwave regions. Furthermore, graphene fabrication technology has tremendously improved over the past few years, which has enabled researchers to build working models of the exciting theoretical phenomenon which were extensively reported over the last decade. Graphene is a promising material that can be integrated into electronic circuits due to its good compatibility with complementary metal-oxide semiconductor (CMOS) technology, low voltage requirement, and switching speed in the GHz range. The tunable response of graphene to the EM waves is one of the most important properties employed in graphene-based MMs. Most monolayer graphene MMs have narrow bandwidths and a low range of tuning. However, patterning of the graphene layer is used to obtain a broadband response and high tuning range. The unit cell structure of a MM defines its frequency response over a wide range, and graphene can be used to further tune the response enabling dynamic use over a frequency range. Different defense applications require different MM responses, narrow or broadband peaks, which can be achieved by graphene MMs. The response of graphene can be tuned chemically by doping or electrically by applying bias, both of which change the μ_C or E_F of graphene, which leads to the variation of its conductivity. A field-effect transistor (FET) type structure has to be fabricated to apply bias to graphene. The bias is applied between two conducting contacts, which act as source and drain. They are placed on the graphene layer, and

the dielectric is placed between the top layer of graphene and the grounding layer. The dielectric usually has semiconducting properties to effectively apply the bias. The thickness of the dielectric determines the amount of gate voltage that has to be applied for biasing. A thinner dielectric layer leads to a lower requirement of the gate voltage to achieve a certain μ_C. The metal layer at the bottom acts as a reflecting surface which can enhance the interaction of the light with graphene.

In the current chapter, the applications of graphene-based MMs in the defense and security systems have been divided into three main categories, stealth, communication, and imaging.

Stealth is a major area of research in the defense sector because it is of the utmost importance on the battlefield on land, water, or sea. IR, THz, and microwave-based systems are used to detect objects which can lead to a tactical advantage. Graphene-based MMs having broadband and tunable absorption response are used as RASs. Specifically, graphene FSSs were used to prevent detection from thermal systems (Safaei et al. 2017; Lin et al. 2019; Yao et al. 2021). THz frequency region is gaining importance in stealth applications, and graphene has an excellent absorption response in the THz region. Therefore, several research groups have studied graphene MM-based absorbers (He et al. 2016, 2018; Liu et al. 2018b; Mishra et al. 2018; Wu et al. 2019; Jiang et al. 2020; Yan et al. 2020). Apart from stealth applications, the absorbing surfaces can be used to shield sensitive electronic components that may be damaged on exposure to unwanted EM energy. Further, the absorber should have a good absorbance over a wide range of incident angles and different polarization conditions. Microwaves are widely used to detect objects over long-range, and graphene-based MMs can improve the efficiency and the weight of microwave absorbing surfaces (Huang et al. 2014; Balci et al. 2015). The development in the understanding of MMs and advancement of simulation techniques have led to the development of cloaking techniques in the IR (Alù and Engheta 2005; Farhat et al. 2013) and THz region (Chen and Alù 2011; Zhao et al. 2016; Xiao et al. 2018; Song et al. 2020) using graphene MMs. Cloaking can be the future of stealth technology.

Intel is the best weapon one can have over the battlefield. Hence, having fast and secure communication technology can influence the outcome of a battle. Graphene-based MMs can improve the efficiency and reduce the size of antennas, and in some cases even hide the antennas from the enemy. Various THz (Seyedsharbaty and Sadeghzadeh 2017; Liu et al. 2018d) and microwave antennas (Huang et al. 2016; Kumar et al. 2016) have been developed. Further, the principle of orbital angular momentum waves (Meng et al. 2019) and polarization multiplexing (Zhang et al. 2021) can be used to increase the amount of data that can be transferred. Slow light devices can significantly reduce the energy consumption of the data transfer (Liu et al. 2018a, c; Jia et al. 2019). Secure transfer of data is vital to the success of a security system. Graphene MM-based holography systems can the used to encrypt data and transfer it securely using THz waves (Momeni et al. 2018; Rajabalipanah et al. 2020).

THz imaging is at the forefront of new security imaging technology. The manipulation and capturing of the THz waves are important for the advancement of the aforementioned technology. As previously discussed, graphene MMs are highly

prevalent in THz technology because conventional materials require extreme environmental conditions to manipulate THz waves. Ultrathin lenses (Liu et al. 2018e; Yao et al. 2018) and mirrors (Hosseininejad et al. 2019) with adjustable focus were designed using graphene MMs. Further, hyperlenses (Andryieuski et al. 2012) were used to go beyond the diffraction limit of the THz technologies. High-speed modulators were developed in the IR (Yao et al. 2014; Zeng et al. 2018) and THz (Sensale-Rodriguez et al. 2013; Malevich et al. 2020) frequency range to improve the speed and efficiency of IR and THz imaging systems.

Despite the significant progress in MM technology, it still lacks experimental data. There is a lot of simulation and analytical-based literature to understand the physics of the graphene-based MMs. However, the assumptions and idealistic parameters included in the simulations do not present the true response of the structures and various losses. The improvement of fabrication technologies has led to an increase in the number of test devices being built which still requires optimization and improvement for real-life applications. Once the technologies are applied in the defense sector, they can be upscaled and used in day-to-day life to improve the life quality of human beings.

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Active Metamaterial Frequency Selective Surface (FSS) Based Tunable Radar Absorbing Structure (RAS)

4

Saptarshi Ghosh

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Abstract

This book chapter presents a comprehensive review of various active metamaterial frequency selective surface (FSS)-based switchable and tunable radar absorbing structures (RASs). Passive absorbing structures offer limited post-fabrication flexibility and cannot be used for any multifunctional characteristic. In contrast, an active RAS can be exploited for manifold applications by regulating the external control. Further, fabrication and installation error can be compensated by using a tunable structure, unlike a passive design. Therefore, active FSS-based RASs are being widely investigated since the last decade. The reported active RASs can be broadly classified into three categories based on their realization techniques, such as mechanically actuation, magnetically regulation, and electronically control. Among them, electronically reconfigurable absorber designs are often preferred due to their high switching speed, wideband tuning, low cost, and compact size. Electronically regulated active components, viz.

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varactors, p-i-n diodes are mounted across the gaps etched in the FSS designs and their variable characteristics are precisely regulated by applying required biasing voltages. Based on use of these active components, different types of electronically actuated structures have been realized, such as switchable absorber/reflector, switchable absorber, and tunable absorber, spread across different frequency regime. This book chapter summarizes detailed information about the working principle, design topology, biasing mechanism, and experimental procedure of those developed FSS-based switchable and tunable RAS designs, focusing across microwave frequency range.

Introduction

In the previous chapters, several types of narrow-band and broadband FSS-based absorbing structures are presented. The designs are mostly passive absorbers having fixed reflectivity characteristics. These RASs offer limited post-fabrication flexibility and cannot be used for any multifunctional characteristic. In contrast, an active absorber can be exploited for manifold applications by regulating the external control. Further, fabrication and installation error can be compensated by using a tunable structure, unlike a passive design. Therefore, active metamaterial FSS-based RASs are being widely investigated since the last decade. The reported active RASs can be broadly classified into three categories based on their realization techniques, such as mechanically actuation, magnetically regulation, and electronically control. Among them, electronically reconfigurable absorber designs are often preferred due to their high switching speed, wideband tuning, low cost, and compact size. Electronically regulated active components, viz. varactors, p-i-n diodes are mounted across the gaps etched in the FSS designs and their variable characteristics are precisely regulated by applying required biasing voltages. Based on the use of these active components, different types of electronically actuated structures have been realized, such as switchable absorber/reflector, switchable absorber, and tunable absorber, focusing across microwave frequency range.

This chapter presents a comprehensive review of different techniques used for realizing metamaterial FSS-based switchable and tunable RASs. Several examples are illustrated and explained with respect to their principles of operation, design topology, biasing mechanism, and experimental procedure. A brief comparison among the designs across each category has also been carried out to describe their advantage(s) and disadvantage(s) with respect to the state-of-the-art.

Electronically Regulated RAS

Electronically regulated RASs are realized by mounting active components across the gaps present in the FSS designs and their responses are being controlled by the external biasing voltage. P-i-n diodes, varactors, RF MEMS are generally used to

realize these active absorbing structures. These components have low response time (of the order of ns) and can be precisely regulated from remote locations. The tuning range is also wide in nature, thereby covering multiple applications. Low cost and compact size are other benefits of these components. However, adequate circuitries are required in the design to bias the components with required voltages. Proper care has to be taken while separating the biasing voltage from the electromagnetic (EM) signal propagation. Therefore, the active RASs need dedicated investigation along with the illustration of the biasing circuits.

Based on the principle of operation, these active FSS-based reconfigurable RASs can be broadly divided into three categories, such as switchable absorber/reflector, switchable absorber, and tunable absorber. A detailed review of each of these types of RAS designs are presented below.

Switchable Absorbers/Reflectors

Metamaterial FSS-based switchable absorbers/reflector structures give rise to absorption during one biasing state of the active components, while exhibit reflection under another biasing state. They can be particularly used in stealth technology, where the incident EM wave can be absorbed to hide the object in the absorptive state, and it can work as a perfect reflector in the reflective state (Bahret 1993). They have another application in the area of compressive imaging, where an array of switchable reflectors can be used to form frequency-selective spatial light modulators (Shrekenhamer et al. 2013a).

Narrow-Band Switchable Absorbers/Reflectors

Over the past decade, several electronically controlled FSS-based switchable absorbers/reflectors are reported in literature. In (Yoo and Lim 2012, 2014a), single-band switchable absorber/reflector structures have been designed, where two different configurations for two different biasing states are presented, as shown in Fig. 1a. In one configuration, the unit cell geometry consists of an electric field-driven LC (ELC) resonator printed in a periodic pattern on a grounded dielectric substrate. FR4 (dielectric constant of 4.4, and loss tangent of 0.02) has been used as the dielectric, and copper (conductivity of 5.8×10^7 S/m) has been used as the metal element. The ELC resonator has a connected central wire and exhibits a reflection dip at 10.50 GHz, as observed from Fig. 1c. This state has been described as the ON state. During the OFF state, the central wire of the resonator is disconnected. Therefore, the electric resonance of the structure is destroyed and the absorption mode is switched to the reflection mode. Two different prototypes are fabricated, as shown in Fig. 1b, and their corresponding responses are compared with those of the simulated results. However, no active device has been used in these geometries; rather two different configurations of a single structure have been used to realize the switching operation.

Later the same group of authors has presented another switchable absorber/reflector design, where the above concept has been realized by using semiconductor

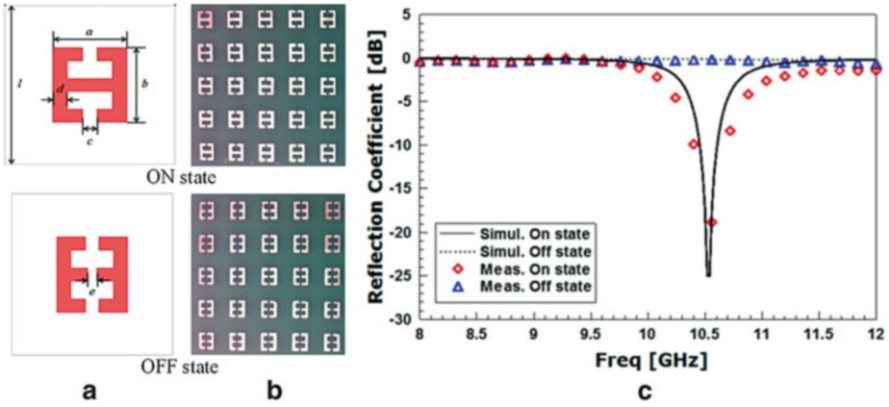


Fig. 1 Switchable absorber/reflector structure, using two different configurations. (**a**) Unit cell geometries, and (**b**) simulated reflection coefficients under ON and OFF states. (Reproduced from Yoo and Lim (2014a) with permission. Copyright 2014, Taylor & Francis)

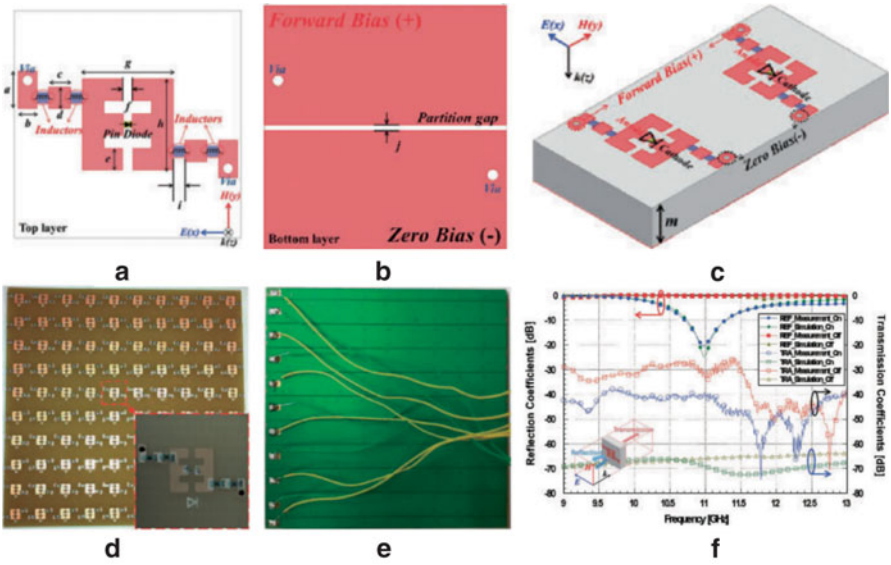


Fig. 2 Switchable absorber/reflector structure, based on the use of p-i-n diode. (**a**) Top view, (**b**) bottom view, and (**c**) perspective view of the proposed unit cell geometry. (**d**) Top layer and (**e**) bottom layer of the fabricated sample. (**f**) Comparison between simulated and measured scattering coefficients. (Reproduced from Yoo and Lim (2014b) with permission. Copyright 2014, IOP Publishing)

switches (p-i-n diodes) (Yoo and Lim 2014b). Similar shaped ELC resonator has been used in the basic design, where p-i-n diodes are soldered across the gap present in the central wire. Figure 2a, b, c illustrate the unit cell of the proposed design. The

diode has a small resistance during forward bias, thus acting as a short circuit condition. This mimics the unit cell topology with central conductor connected, as illustrated in Fig. 1a Due to this connection, the structure, under forward bias, exhibits a narrow-band reflection dip at 10.79 GHz with reflectivity of −25 dB. Since the structure is laminated by a complete metal plane, there is no transmission and the structure results in a near unity absorption at the same frequency (10.79 GHz). Thus, the design operates in the absorptive mode under forward bias condition. In contrast, a large capacitance is resulted during zero (or reverse) bias voltage, which is equivalent to the unit cell topology with no central conductor connected. Then the electric resonance is destroyed and the structure is switched to the reflective mode. Hence, a switchable absorber/reflector structure has been realized from a single topology by the use of p-i-n diodes.

These p-i-n diodes have three layers, viz., P-type layer, I-intrinsic layer, and N-type layer. P-type layer contains positive charge carriers (i.e., holes), whereas N-type layer consists of negative charge carriers (i.e., electrons). The intrinsic layer is very thin (of the order of 10 to 200 microns) and comprises of undoped semiconductors. The operating principle of the p-i-n diode is much similar to that of a normal p-n junction diode. In addition, owing to the presence of the intrinsic layer, the depletion region in the p-i-n diode is larger in size, thereby having a high reverse breakdown voltage, and a low level of junction capacitance. Another useful aspect of the p-i-n diode is observed during high frequency signals, where the diode appears as a resistor rather than a non-linear device, and produces no significant rectification or distortion. Therefore, p-i-n diode is often preferred over a normal p-n junction diode in a number of applications, such as RF switch, high voltage rectifier, and photodetector (Bill n.d.).

Although p-i-n diode is a passive device, an external voltage source is required to regulate the diode characteristics. The voltage should be provided from a direct current (DC) supply, with minimal fluctuation in current or power ratings. To operate the diodes under forward bias, a voltage above the threshold value (typically 0.7 Volt for Silicon-based diodes) should be provided across each diode. Either zero or a fixed negative voltage is generally applied to operate the diodes under reverse bias. Different diodes exhibit different values of resistance (during ON state) and capacitance (during OFF state) under different current ratings, which can be obtained from their datasheets. Therefore, careful selection of bias voltages and associated currents is essential while operating the p-i-n diodes under different biasing conditions. In addition, the external voltage source should be protected from the time-varying radio frequency (RF) signal during operation.

Since the diodes are connected across a periodic pattern in the FSS geometry, the voltage source cannot be connected across each diode separately. Instead, a dedicated biasing circuitry needs to be designed across the RAS, such that the diodes get connected with one another either through series or parallel connection and the required voltage can be provided from a remote location. Thus an additional biasing network along with the design topology is required to be designed for an active FSS-based RAS.

The researchers in Yoo and Lim (2014b) have used an innovative method to regulate the diodes in the proposed design. An M/A-COM MA4PBL027 beam-lead p-i-n diode is used as the switching component mounted across the gap in the central conductor. The upper and lower sections of each ELC resonator are then connected to two different sections of the ground plane through metallic vias. A photo-imageable solder resist ink (PSR) is used to isolate the two sections of the ground plane. These two sections are then separately connected to the positive and negative terminals, respectively, of the external voltage source, as depicted in Fig. 2d, e. Two chip inductors (from Murata LQW04AN5N1C00), each of them having a value of 5.1 nH, are connected in series at each end of the ELC resonator. The inductors have a self-resonance frequency (SRF) greater than 10 GHz and behave as RF choke (high impedance) for protecting the external voltage source from RF signals. At the same time, DC voltage is able to transmit through these inductors, thus ensuing proper biasing across each active component.

After implementing the required biasing circuitries, the scattering coefficients of the fabricated sample have been measured under different biasing voltages. During ON state, 1.3 V has been applied across each p-i-n diode, and the measured reflection coefficient and transmission coefficients are obtained as −22.41 and −40.92 dB, respectively, at 10.97 GHz. Since the ground plane is discontinuous due to the presence of PSR ink, the transmission coefficient is also measured, but found to be insignificantly small, as observed from Fig. 2f. Thus, near unity absorption is achieved at 10.97 GHz with an absorption coefficient of 99.40%. During OFF state, the voltage across the p-i-n diodes is reduced to zero, and the resonance effect gets destroyed. This gives rise to the reflective behavior of the structure, showing a high value of reflection coefficient (0.28 dB at 10.97 GHz). Thus, the switchable absorber/reflector concept has been experimentally demonstrated in the proposed metamaterial FSS-based active RAS.

Zhu et al. have designed another narrow-band switchable absorber/reflector for two opposite polarizations, where the biasing network is provided through metallic vias (Zhu et al. 2010a). Two sets of ELC resonators are patterned with orthogonal orientations in the unit cell, where each set consists of two identical circular arcs coupled together through p-i-n diodes, as shown in Fig. 3a. When a zero or negative bias is applied across the diode, it exhibits a large capacitance that enables opposite charges to be accumulated on the two arches resulting in a strong resonance. On the contrary, during forward bias, the diodes exhibit a small resistance that cancels out the opposite charges on the neighboring arches and the resonance gets destroyed. Thus, by changing the bias voltage, the absorptive mode can be switched to the reflective mode or vice versa. Since the topology has used two sets of resonators in orthogonal configurations, the proposed RAS can respond to any incident EM wave irrespective of its polarization characteristic.

During full-wave simulation, the proposed design has been modeled on a FR4 dielectric substrate, whereas the metal patterns are considered to be made of copper. The transmission coefficient is happened to be zero due to the presence of metal plane at the back side, and the absorptivity is obtained from the reflection coefficient only. At zero bias, the diodes are operating under OFF state and a perfect absorption

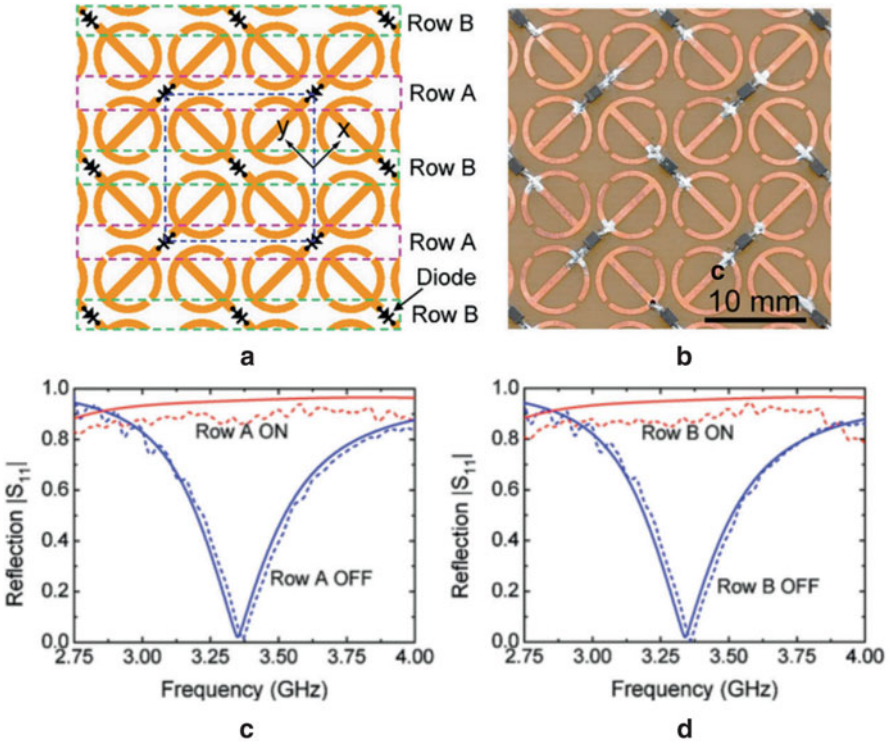


Fig. 3 Switchable absorber/reflector structure for two opposite polarizations. (**a**) Schematic view of the structure, and (**b**) fabricated sample prototype. Comparison between simulated (*solid lines*) and measured (*dashed lines*) reflection coefficient under different diode states: (**c**) along *x*-axis and (**d**) along *y*-axis. (Reproduced from Zhu et al. (2010a) with permission. Copyright 2010, AIP Publishing)

is achieved at 3.34 GHz. In contrast, the absorptivity is dropped while increasing the bias voltage and the structure is subsequently switched to the reflective mode.

To verify the proposed concept, a sample prototype has been fabricated as depicted in Fig. 3b. The biasing network has been arranged at the back side of the structure, and vias are mounted through the substrates for providing the required connection to the diodes. Two different biasing layers have been considered for regulating two different sets of diodes (mounted across rows A and B), which can to be turned ON/OFF separately without disturbing the other set of switching components. The fabricated sample is then measured under different biasing voltages and their reflection coefficients are compared with those of the simulated responses. Figure 3c, d depict the responses obtained under different polarized EM waves, incident along *x*-axis and *y*-axis, respectively, and the design is found to exhibit similar switching characteristic for both the cases. Thus, a polarization-insensitive switchable metamaterial absorber/reflector RAS has been designed and experimentally validated.

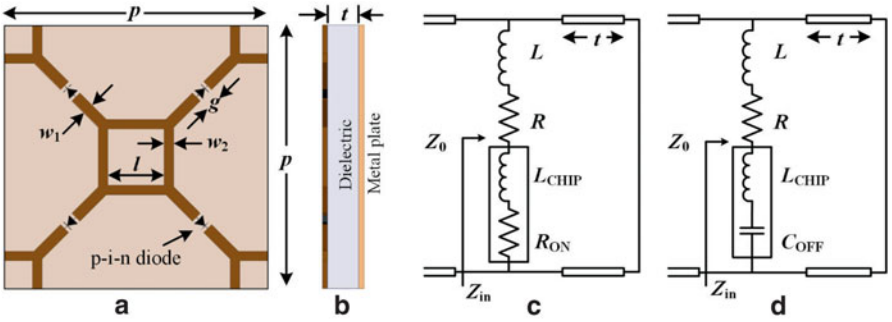


Fig. 4 Polarization-insensitive narrow-band switchable absorber/reflector structure. (**a**) Unit cell geometry, (**b**) side view. Equivalent circuit model under (**c**) ON state and (**d**) OFF state. (Reproduced from Ghosh and Srivastava (2016a) with permission. Copyright 2016, IEEE)

Ghosh and Srivastava have later designed another polarization-independent switchable absorber/reflector RAS based on a simplified topology and biasing network (Ghosh and Srivastava 2016a). The proposed geometry is comprised of periodic square loops connected through diagonal elements, where the p-i-n diodes are symmetrically mounted across the diagonals. The geometry is patterned on FR4 dielectric substrate terminated by a metal ground, whose top view and side views are depicted in Fig. 4a, b, respectively. During forward bias, the diode exhibits a small resistance (in series connection with a parasitic inductance) and behaves as a short circuit line. Then the surface current flows continuously across the top layer, without encountering any gap in the metallic pattern. Under reverse bias, the diode provides a large capacitance. The equivalent circuit model for both ON and OFF states are illustrated in Fig. 4c, d, respectively.

Owing to absence of any equivalent capacitance at the top surface, the structure exhibits a perfect reflective behavior under ON state. Since the grounded substrate is electrically thin, only inductive effect is obtained from the dielectric, and thus no overall capacitance is resulted. Under OFF state, the diode capacitance and the distributed inductance from the metallic pattern contribute to an absorption frequency at 3.71 GHz. Thus, the proposed design behaves as a reflector/absorber under different biasing conditions. The geometry has also been studied for different polarization angles and found to exhibit identical responses owing to its four-fold symmetric topology.

To validate the proposed concept, the design has been fabricated and characterized under normal incidence. The sample prototype consists of 4 × 4 unit cells, covering a total of 41 square loops as depicted in Fig. 5a. BAP 70–03 p-i-n diode from NXP are mounted symmetrically across the gaps in the structure. To regulate the active components, two vertical metal lines are printed at the two sides of the sample and narrow-width horizontal metal lines are drawn across the square loops away from those vertical lines. This type of biasing configuration enables the cathodes and the anodes of all the p-i-n diodes to get simultaneously connected with the left- and right-sides bias lines, respectively. When the right and left vertical

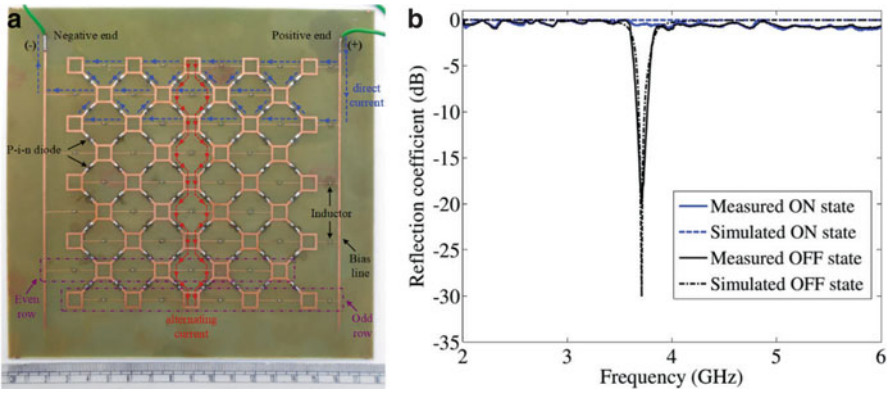


Fig. 5 Polarization-insensitive narrow-band switchable absorber/reflector structure. (**a**) Photograph of the fabricated prototype. (**b**) Comparison between simulated and measured reflection coefficients under normal incidence. (Reproduced from Ghosh and Srivastava (2016a) with permission. Copyright 2016, IEEE)

metal lines are joined with the positive and negative terminals of a supply voltage, a continuous direct current flows through all the diodes in parallel connection. Blue colored arrows indicate the flow of DC through the p-i-n diodes and the biasing circuitries, whereas the red dotted lines designate the alternating current flow across the geometry. Although the time-varying AC propagates through the diode elements in the geometry, the current value is too small (of the order of μA) to damage the active devices. However, the RF signals need to be restricted from entering into the DC voltage source. Hence, the lumped inductors, having a value of 12 nH and an SRF of 3.62 GHz, are mounted between the successive metal lines to attain the required isolation. The horizontal lines are also made of narrow-width such that they behave as distributed inductances and further restrict the propagation of RF signal. Thus the biasing network has no considerable effect on the design and the desired response is obtained.

The reflectivities of the prototype has been measured under normal incidence for both ON and OFF states and the structure exhibits a complete reflection and perfect absorption at 3.70 GHz, respectively, as observed in Fig. 5b. The sample has also been characterized under different polarization angles and the consistency in the reflection coefficient confirms the polarization-insensitive behavior of the proposed geometry. Further, the biasing circuitries have been established under parallel configuration, and only a nominal voltage is required to regulate all the diodes simultaneously. Thus, the proposed design concept can be suitably implemented for various active radar absorption applications.

Broadband Switchable Absorbers/Reflectors

The switchable absorber/reflector designs discussed in the previous section exhibit single-band absorption responses. However narrow-band absorbers have limited applications; therefore, a few wideband switchable geometries have also been

reported recently. These structures offer reflective behavior under one of the biasing conditions, whereas broadband absorption is observed during the other switching state of the p-I-n diodes.

P. Kong et al. have presented an active metamaterial RAS, where the geometry exhibits broadband absorption during OFF state and complete refection under ON state (Kong et al. 2015). The proposed geometry is comprised of three concentric split rings patterned on a FR4 dielectric, and a honeycomb spacer is used as an isolation layer between the ground plane and the dielectric, as depicted in Fig. 6a, b. Lumped resistors and switchable p-i-n diodes are mounted in a parallel connection across the two splits in the outer ring. A set of inner split rings are introduced for further enhancing the absorption bandwidth.

During OFF state, the p-i-n diode exhibits a high capacitive effect and the RF signal is unable to propagate through the active component with ease. As a result, the EM wave passes through the lumped resistor mounted at other end of the outer ring and contributes to large ohmic loss, thereby resulting in wideband absorption from 4.5 to 7.7 GHz. On the contrary, the diode offers a small resistance during ON state, thus producing a short circuit path. Then the EM wave mostly transmits through the diode, without encountering the resistor component, and gives rise to the reflective effect over the frequency band. Unlike earlier case, the reflection band is not perfect, with an average reflectivity of −2.3 dB in 4.5–7.7 GHz, due to presence of parasitic components associated with the lumped resistor models. The proposed geometry is asymmetric in nature and results in a switchable absorption/reflection behavior for a particular polarization angle only.

While validating the proposed concept, the biasing circuitries have been designed in the top layer only, with the unit cells being connected in series combination. To isolate the biasing lines from time-varying RF signal, inductors are mounted between the successive unit cells. Since lumped inductors exhibit narrow-band impedance response, three different inductors, having different SRF values, are connected in series such that the biasing lines have a high impedance characteristic over a wide frequency band. In the fabricated prototype, lumped resistors of 100 Ω

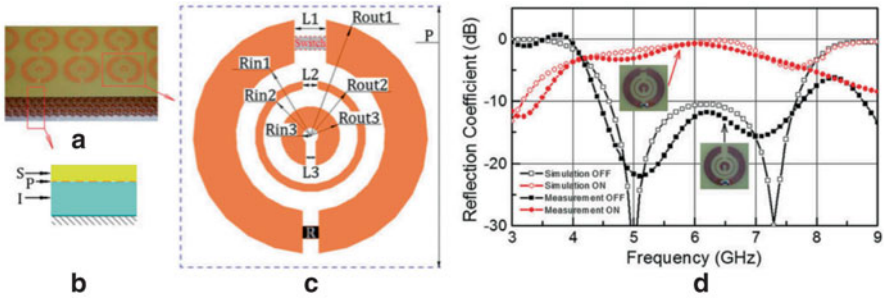


Fig. 6 Wideband switchable absorber/reflector structure. (**a**) Photograph of the fabricated prototype, (**b**) side view of FSS, and (**c**) unit cell geometry. (**d**) Comparison between simulated and measured reflection coefficients under normal incidence. (Reproduced from Kong et al. (2015) with permission. Copyright 2015, Taylor & Francis)

are soldered across one gap and BAP 70-03 p-i-n diodes are mounted across the other gap in the split ring.

While measuring the sample prototype under normal incidence, reflection coefficient is observed below −9 dB over the frequency range of 4.1 to 7.3 GHz during OFF state (zero bias voltage). Under ON state, the bias voltage is increased to 10 V (such that all the diodes connected in series over one horizontal line are simultaneously turned on) and wideband reflective behavior is observed with an average reflectivity of −2.8 dB. The comparisons between the measured and simulated responses are depicted in Fig. 6b and they are found in good agreement with each other. Thus, the concept of wideband switchable FSS-based absorber/reflector is demonstrated.

The above topology, as discussed, is polarization-sensitive and can be operated for single polarization only. A further advancement has been carried out by Ghosh and Srivastava, by designing a polarization-insensitive broadband switchable absorber/reflector RAS (Ghosh and Srivastava 2016a). The proposed design is an extended version of their previous work on narrow-band switchable absorber/reflector RAS, where lumped resistors have been mounted across the gaps in the square loop topology. An air spacer has also been introduced between the ground plane and the FR4 dielectric, to maintain the quarter-wavelength condition. The proposed geometry has been illustrated in Fig. 7a.

The resistors absorb the incident EM wave owing to high ohmic loss, thereby exhibiting large absorption bandwidth under OFF state. On the contrary, p-i-n diodes behave as short circuit during ON state, and the structure exhibits good reflection over the entire frequency range. During full-wave simulation, reflection coefficient is observed below −10 dB over the frequency range of 3.48 to 8.04 GHz, under OFF state. However, when the diodes are operated under forward bias, good reflectance performance (maximum reflection coefficient of −1.56 dB) is observed over the C-band (4–8 GHz) frequency range. The lumped resistors absorb the incident EM wave even during the ON state, thereby producing a finite ohmic loss and hence near perfect reflective behavior cannot be achieved unlike the single-band switchable structures.

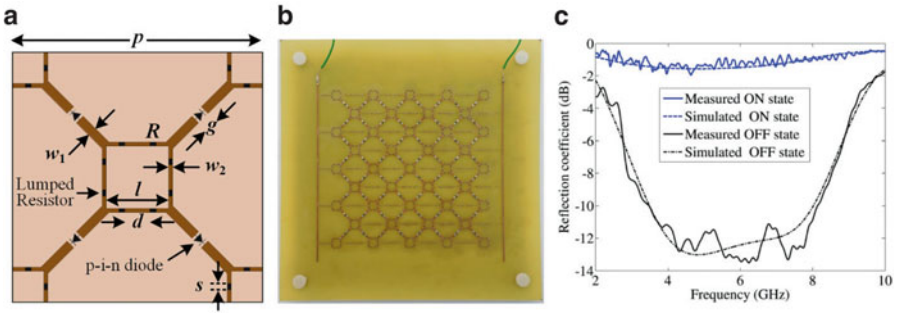


Fig. 7 Polarization-insensitive broadband switchable absorber/reflector structure. (**a**) Unit cell geometry. (**b**) Photograph of the fabricated prototype. (**c**) Comparison between simulated and measured reflection coefficients under normal incidence. (Reproduced from Ghosh and Srivastava (2016a) with permission. Copyright 2016, IEEE)

The fabricated sample is made of 128 mm × 128 mm, containing 41 square loops, as depicted in Fig. 7b. Lumped resistors and p-i-n diodes are mounted in their desired locations. The biasing circuitry has been made similar to its narrow-band switchable counterpart explained in the previous section. In addition, four different valued inductors are connected between the successive unit cell geometries such that the overall impedance response of the biasing line has a wideband characteristic across the required frequency band.

After realizing the biasing network, the reflection coefficient of the prototype has been measured under normal incidence and a broadband absorption bandwidth ranging from 3.56 to 8.16 GHz (with a fractional bandwidth of 79.52%) has been achieved under OFF state. The structure exhibits good reflection performance in the same frequency band during ON state. The measured and simulated responses are compared and found to have a reasonable agreement, as depicted in Fig. 7c. Thus, the polarization-insensitive metamaterial FSS-based switchable absorber/reflector RAS structure has been successfully demonstrated.

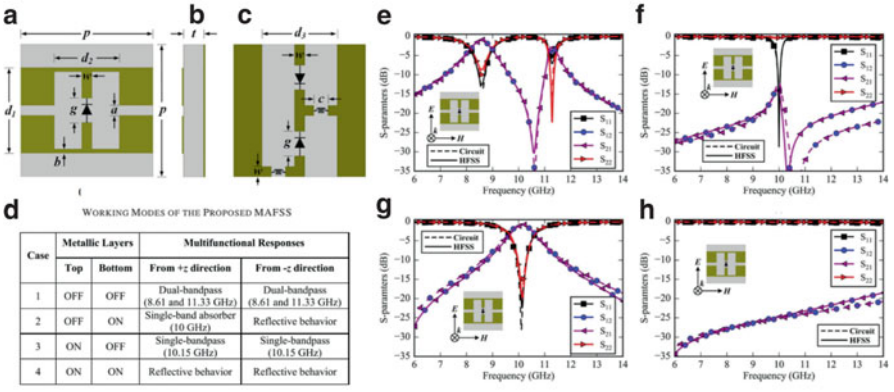
Subsequently, several other narrow-band and wideband switchable absorber/reflector designs have been presented exhibiting various characteristics, such as compactness (Ghosh and Srivastava 2016b), multi-layer topology (Li et al. 2020), multifunctional behavior (Yang et al. 2019), and so forth. H. Li et al. have designed a wideband polarization-insensitive switchable absorber/reflector FSS design with multifunctional characteristics (Yang et al. 2019). The design employs four unitary elements with similar configurations, where p-i-n diodes and lumped resistors are symmetrically mounted across each element. One pair of via holes is drilled between the top and bottom side metallic patterns such that the p-i-n diodes oriented under different polarizations can be regulated independently. When dc voltage has been separately applied across the horizontally and vertically loaded p-i-n diodes, four different working states are arisen, viz., '00', '11', '01', and '10', where '1' and '0' correspond to the diode operating conditions under forward and reverse bias voltages, respectively. For the states '00' and '11', the geometry is exhibiting polarization-insensitive absorption and reflection responses, respectively, whereas for other states '01' and '10', the responses (absorption/reflection) are getting reversed under orthogonal polarizations. During absorption mode (appearing under different working states), a −10 dB absorption bandwidth is observed from 4.88 to 12.57 GHz, whereas the reflective behavior is obtained over a large frequency band under the reflective mode.

The proposed AFSS structure has been fabricated on FR4 substrates, and lumped components have been mounted accordingly. An air spacer and a metal ground plane have been appended below the periodic geometry. Due to continuous connection across the unitary elements, there is no need of dedicated bias lines. Diodes in each row and column are biased through two sets of vertical and horizontal bias lines, respectively, printed at each side of the fabricated prototype. This type of biasing configuration enables independent regulation of the diodes, without any additional circuitry. Afterwards, the fabricated sample has been measured and the responses are in reasonable agreement with those of the simulated results. Thus, the multifunctional absorber/reflector active FSS structure has been demonstrated.

Recently, Phon et al. have designed another multifunctional active FSS geometry based on the switching responses of active components (Phon et al. 2019). By properly controlling the biasing states of the diodes mounted across different layers, several EM characteristics, such as absorption, reflection, and/or transmission responses can be obtained. The proposed geometry is comprised of connected metallic grids periodically arranged on the top and bottom sides of a dielectric substrate, as illustrated in Fig. 8a, b, c. P-i-n diodes are mounted across the gaps in the metallic strips such that the biasing of each set of the diode can be independently accomplished.

When the top and bottom layer geometries are separately characterized, it is observed that the top layer exhibits a combination of bandpass and bandstop responses, under ON as well as OFF states. But, the bottom metallic pattern offers a transmissive and reflective behavior, under OFF and ON conditions, respectively. When the overall geometry is analyzed, four different EM responses are attained, based on the combinations of different working states, as depicted in Fig. 8d. Among them, one set of working states (ON-ON) offers a reflective behavior, whereas another working state (OFF-ON) exhibits a narrow-band absorber at 10 GHz. Other two working states offer single−/dual-band transmissive responses at microwave frequencies. The simulated scattering parameters for different working states are also depicted in Fig. 8e, f, g, h.

The proposed geometry has been fabricated and the biasing circuitries have been established. The topology has been designed such that no dedicated biasing line is required to regulate the diode components; the voltage can be provided at the end of the sample only. Since the diodes either in top or bottom layer are connected through parallel circuit, a nominal voltage is sufficient to simultaneously activate all the diodes under forward bias condition. After realizing the biasing network, the sample



Case	Metallic Layers		Multifunctional Responses	
	Top	Bottom	From +z direction	From -z direction
1	OFF	OFF	Dual-bandpass (8.61 and 11.33 GHz)	Dual-bandpass (8.61 and 11.33 GHz)
2	OFF	ON	Single-band absorber (10 GHz)	Reflective behavior
3	ON	OFF	Single-bandpass (10.15 GHz)	Single-bandpass (10.15 GHz)
4	ON	ON	Reflective behavior	Reflective behavior

Fig. 8 Multifunctional active FSS structure. Unit cell geometry: (**a**) top view, (**b**) side view, and (**c**) bottom view. (**d**) Working states of the proposed design. Simulated scattering parameters under normal incidence: (**e**) OFF-OFF state, (**f**) OFF-ON state, (**g**) ON-OFF state, (**h**) ON-ON state. (Reproduced from Phon et al. (2019) with permission. Copyright 2019, IEEE)

prototype has been measured and the responses are recorded for different working states. The results have also been compared with those of the simulated responses and are found to have good agreement with one another. Therefore, a multifunctional switchable active FSS structure has been designed, developed, and experimentally demonstrated/validated.

Switchable Absorbers

Switchable absorbers can be defined as the structures whose absorption response can be switched between two different frequency bands by an external stimulus. Electrically controllable switchable absorbers generally use p-i-n diodes in the design, and the absorption frequency or bandwidth can be regulated depending on the supply voltage. Recent development in military and commercial applications necessitates switchable RAS designs to be employed in electromagnetic compatibility test facilities, radar camouflage, tagging and asset tracking systems, and so forth. These absorber structures can also be used in beam steering antenna for beam forming applications, where the radiation pattern can be programmed by regulating the frequency response of switchable geometry (Han et al. 2019). It has another exciting application in the area of hyperspectral imaging, where frequency selectivity can be exploited to design a broadband detector. At terahertz frequency, single-pixel detector can also be designed using an array of switchable FSSs to encode the scene of interest in the area of compressive imaging (Chana et al. 2008).

Since the past few years, several electronically controlled metamaterial FSS-based switchable absorbers have been reported in literature. Most of the geometries employ p-i-n diodes as the active components, whose response can be tuned by external voltage. Under forward bias, the diodes provide small resistance, whereas large capacitance are generated during zero or reverse bias voltage. Consequently, different types of switchable RAS designs have been developed exhibiting narrow-band and wideband absorption characteristics.

Narrow-Band Switchable Absorbers

In 2010, a narrow-band switchable absorber structure has been developed, where p-i-n diodes are implanted between two ELC resonators as observed in Fig. 9a (Zhu et al. 2010b). The resonators have their own absorption frequencies, governed by the distributed inductance and capacitance originated from the geometries. When the diode is mounted across the slot position, its switchable impedance comes in parallel with the gap capacitance and changes the overall coupling effect. During OFF state, the diode exhibits large capacitance that lowers down the resonance frequency (2.6 GHz) of the RAS. In contrast, the diode offers small resistance under forward bias voltage, and the absorption response shifts to higher frequency (2.9 GHz), as depicted in Fig. 9c. Consequently, the absorption response is switchable between two frequency bands by controlling the p-i-n diode biasing conditions in the proposed RAS geometry.

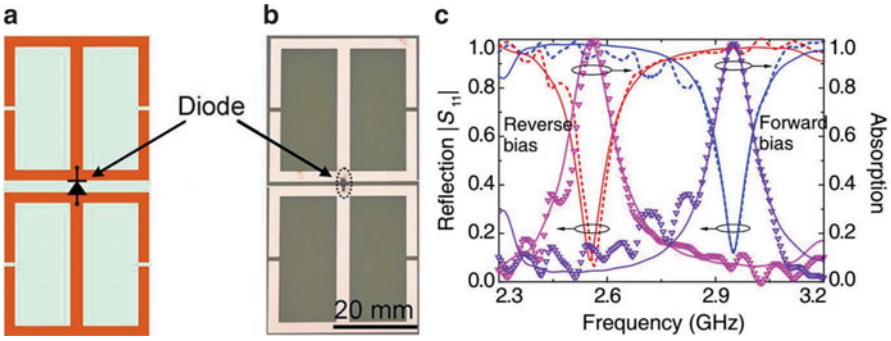


Fig. 9 Single-band switchable absorber. (**a**) Schematic view of the unit cell geometry, (**b**) fabricated sample, and (**c**) comparison of simulated and measured responses. (Reproduced from Zhu et al. (2010b). Copyright 2010, The Electromagnetics Academy)

To validate the performance, the proposed design has been fabricated on a metal-backed FR4 sheet having an overall size of 270 mm × 290 mm, consisting of 7 × 4 unit cells. A portion of the sample prototype is depicted in Fig. 9b. The diodes are mounted across the gaps in the fabricated geometry, and thereafter biased through series connection. Under OFF condition, the prototype exhibits a perfect absorption response at 2.56 GHz (with a peak absorptivity of 99.5%), whereas the absorption frequency shifts to 2.94 GHz maintaining the peak absorptivity at 98%. The measured responses, while compared with those of the simulated results, are in good agreement, thereby verifying the switchable absorption property of the proposed active RAS design.

A similar switchable absorber design has been demonstrated by Xu and Sonkusale in 2013, where p-i-n diodes are mounted in series with the unit cells in each column of the geometry (Xu and Sonkusale 2013). Figure 10a depicts the absorber geometry and the diode arrangement. When the diodes are biased under forward bias, the ELC resonators are connected with one another through short circuit. The geometry will exhibit an absorption response, based on the ELC resonators. However, additional capacitive effect will appear when the diodes are biased under reverse bias, thereby resulting in a change in the absorption frequency. Near perfect absorption responses are obtained in both the states, 4.10 GHz under OFF state and 4.78 GHz under ON state, as shown in Fig. 10b. The capacitance value gets significantly reduced while changing the bias voltage from positive to zero value, and a lowering in the absorption band is observed.

The sample prototype has been fabricated using printed circuit board technique and characterized using two different waveguides covering the entire frequency of operation. Since the diodes are connected in series across each column in an array, adequate dc voltage (9 V) has been provided across each column to simultaneously turn on all the diodes, whereas 0 V is applied to operate the diodes under OFF state. The measured responses exhibit the absorption frequency switching between 3.5 GHz (OFF state) and 5.6 GHz (ON state) while varying the bias voltage.

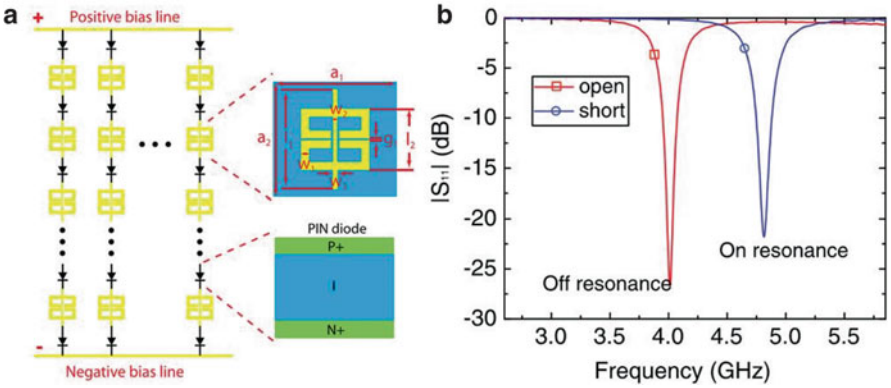


Fig. 10 Single-band switchable absorber. (**a**) Schematic view of the proposed structure, and (**b**) simulated reflection coefficients under different biasing states. (Reproduced from Xu and Sonkusale (2013) with permission. Copyright 2013, AIP Publishing)

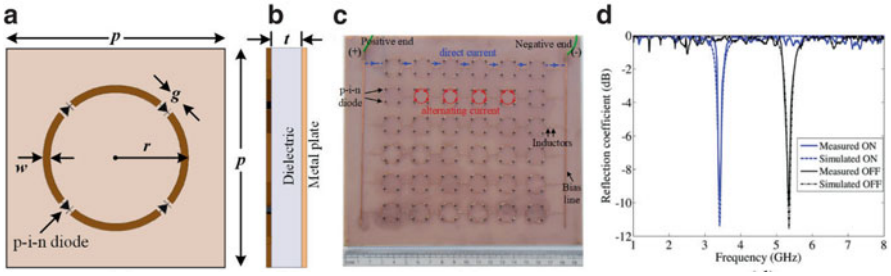


Fig. 11 Polarization-insensitive single-band switchable absorber. Unit cell: (**a**) top view, (**b**) side view. (**c**) Photograph of the prototype. (**d**) Simulated and measured reflection coefficients under different biasing states. (Reproduced from Ghosh and Srivastava (2016c) with permission. Copyright 2016, IEEE)

The above active FSS-based RAS designs demonstrate dynamically switchable absorber concepts, however, the topologies are asymmetric by design and can be used for single polarization only. To increase the design usability, S. Ghosh and Srivastava have presented a four-fold symmetric absorber structure that can be employed for switchable applications under any angle of polarization (Ghosh and Srivastava 2016c). The unit cell geometry comprises of an annular ring, where p-i-n diodes are symmetrically mounted across the gaps in the diagonal positions. The overall design is imprinted on a FR4 dielectric substrate, whose bottom side is completely metal laminated. Figure 11a, b depict the top and side views, respectively, of the proposed switchable RAS structure.

During forward bias voltage, the diodes exhibit short circuit condition owing to small resistance value. As a result, the geometry behaves as a complete annular ring and resonates at 3.41 GHz having reflection coefficient of −11.38 dB. However, the diodes exhibit large capacitance during OFF state, and the corresponding absorption

response shifts to the higher frequency (5.35 GHz). The proposed design has also been studied for different polarization angles and the absorption response is observed to be constant, thereby validating its polarization-insensitive characteristic.

During experimental verification, one sample prototype has been fabricated using PCB technique and p-i-n diodes are mounted symmetrically across the gaps in the structure. Narrow-width metal lines and inductors of requisite values are mounted between the consecutive unit cells to provide the biasing connections across the diodes. Two different inductors, having values of 3.9 and 12 nH are used such that the biasing circuitries can be protected from RF signal under both the switching conditions. These metal lines are further connected to two vertical bias lines present at the sides of the fabricated sample, where the dc voltages are being supplied. The prototype and its biasing configuration are shown in Fig. 11c. When the diodes are forward biased, the structure exhibits a reflection coefficient at 3.41 GHz. During zero bias voltage, this resonance frequency shifts to 5.35 GHz. The measured results have been compared with those of the simulated responses in Fig. 11d and the plots have good compatibility with each other, under normal incidence.

The earlier reported narrow-band switchable absorbers have used series biasing connections, where the diodes present in a row or column are connected in series circuit and a large biasing voltage is required to simultaneously turn on all the diodes. Further, there is an increasing chance of fabrication error, as a damaged diode will prevent the current flow across that row or column, thereby resulting in a significant measurement deviation. Thus, parallel biasing is preferred over series connection, while designing switchable RAS structures.

The above limitations have been resolved in Ghosh and Srivastava (2018), where a switchable FSS-based absorber structure has been designed based on parallel biasing configuration. The proposed unit cell consists of rotated cross-dipole geometry at the top layer of a dielectric substrate, which is connected with the ground plane through a metallic via drilled at the centre point. P-i-n diodes are symmetrically mounted across the gaps in the cross-dipole arms. Figure 12a, b illustrate the top and side views of the unit cell geometry, respectively.

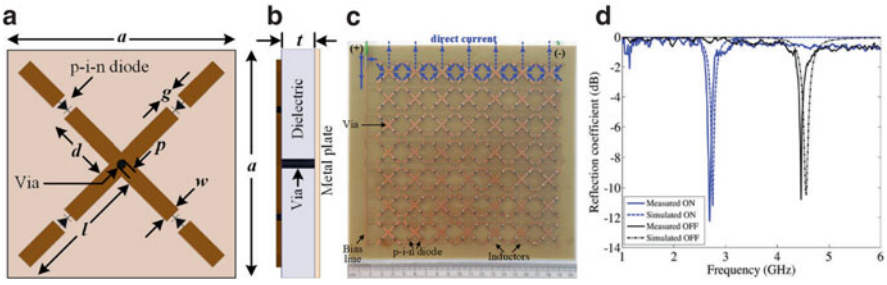


Fig. 12 Polarization-insensitive single-band switchable absorber. Unit cell geometry: (**a**) top view, (**b**) side view. (**c**) Photograph of the fabricated prototype. (**d**) Simulated and measured reflection coefficients under different biasing states. (Reproduced from Ghosh and Srivastava (2018) with permission. Copyright 2018, IEEE)

During forward bias voltage, the diodes exhibit small resistance and current flows through the entire length of the cross-dipole. However, during OFF state, the diodes have a capacitive effect that restricts the overall length of the cross-dipole for current propagation. In that case, the equivalent inductance is resulted from the reduced length that gives rise to an absorption response at higher frequency. Thus, there is a switching activity in the absorption frequency between 2.75 GHz (ON state) and 4.56 GHz (OFF state). The design symmetry further confirms the polarization-independent behavior of the proposed switchable RAS.

To validate the switching concept, a sample prototype having an overall size of 168 mm × 168 mm has been fabricated using PCB technique and diodes are mounted across the gaps, as depicted in Fig. 12c. Horizontal bias lines are drawn across the structures that connect one of the diode terminals to the supply voltage. The other ends of the diodes are connected to the ground plane through metallic vias. Since the diodes are connected in parallel circuit, only a small voltage is required to attain the forward biasing condition. Inductors of requisite values are mounted across the gaps in the metal lines to protect the biasing circuitries from the RF signal. During measurement, a nominal voltage is provided under forward bias and zero voltage is provided to attain the reverse bias. While measured the structure, a reflection dip is observed at 2.69 GHz and 4.46 GHz during ON and OFF states, respectively. Good compatibility is also observed between the measured and simulated responses, as shown in Fig. 12d, and thus a metamaterial FSS-based switchable RAS has been demonstrated exploiting parallel biasing connection.

Later, S. Ghosh and Srivastava have combined the previous two design topologies in single geometry and achieved a dual-band switchable absorber (Ghosh and Srivastava 2017a). The cross-dipole geometry is inscribed inside the annular ring, and the p-i-n diodes are mounted symmetrically in both the resonators. The metallic via is also present at the centre, connecting the cross-dipole with the ground plane. Figure 13a, b illustrate the top and side views of the proposed dual-band unit cell geometry, respectively.

Each of the resonators independently exhibits single-band switchable absorption response, as discussed above. When the diodes from both the geometries are regulated, their absorption frequencies combinedly exhibit dual-band responses, switching between different frequency bands. The simulated reflection coefficient is depicted in Fig. 13c. Four different working states are observed; one of them (OFF-ON biasing condition) offers a bandwidth-enhanced absorption behavior, whereas other three cases exhibit discrete absorption frequencies. Their responses are listed in Fig. 13d. Since the overall design topology is four-fold symmetric, polarization-independent responses are obtained under all biasing conditions. Angular stability is also observed under different modes.

Later, the proposed geometry has been fabricated, and similar biasing configuration has been implemented for independently regulating the p-i-n diodes mounted in different positions. The photograph of the sample prototype is show in Fig. 13e. The diodes present in the annular ring are biased by the voltage source applied between two vertical metal lines drawn at the sides of the prototype, whereas the diodes present in the cross-dipole geometry are regulated by the voltage source applied

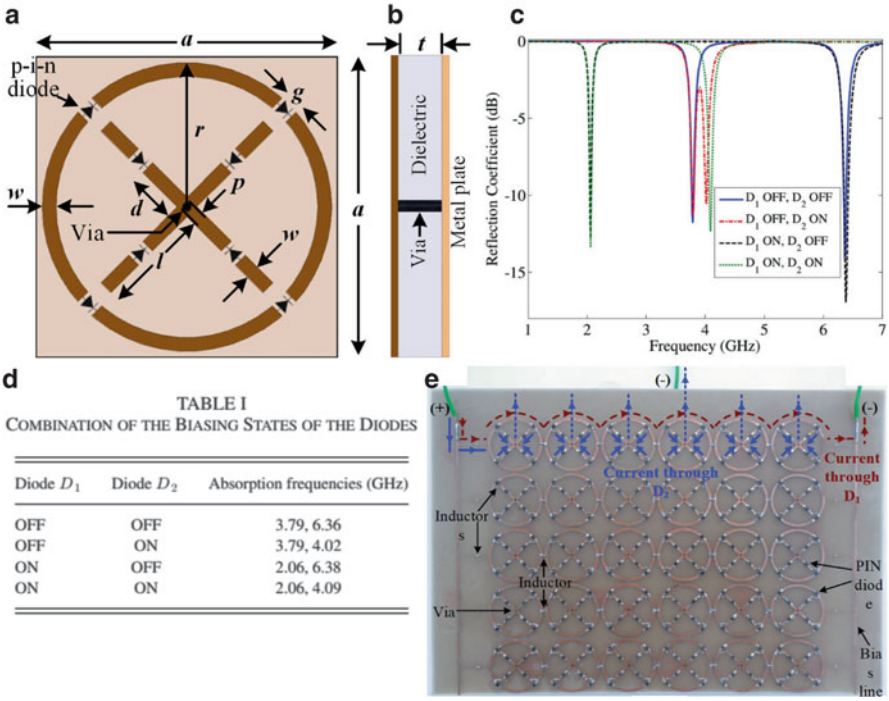


TABLE I
COMBINATION OF THE BIASING STATES OF THE DIODES

Diode D_1	Diode D_2	Absorption frequencies (GHz)
OFF	OFF	3.79, 6.36
OFF	ON	3.79, 4.02
ON	OFF	2.06, 6.38
ON	ON	2.06, 4.09

Fig. 13 Polarization-insensitive dual-band switchable absorber. Unit cell geometry: (**a**) top view, (**b**) side view. (**c**) Simulated reflection coefficients under different biasing conditions. (**d**) Responses under different working states. (**e**) Photograph of the fabricated prototype. (Reproduced from Ghosh and Srivastava (2017a) with permission. Copyright 2017, IEEE)

between one metal line and the ground plane. Different valued inductors are placed at requisite positions to restrict the RF signal from affecting the biasing circuitries. When the bias voltages across the diodes are separately varied, the absorption frequencies are switched accordingly, thereby validating the proposed concept. Thus, a dual-band switchable FSS-based RAS design has been successfully accomplished.

Broadband Switchable Absorbers

The above presented switchable absorber structures offer either single-band or dual-band responses, switching between different frequency bands. However, their bandwidths are limited, having restricted applications in radar technology. Hence, broadband switchable absorber designs are often preferred, which can exhibit absorption responses over wide frequency bands.

In 2002, Zhang et al. have first developed a switchable microwave absorber, covering different frequency bands under different biasing conditions (Zhang et al. 2012). The proposed geometry is comprised of three layers; the top layer has a *RC* network comprising of one lumped resistor and one lumped capacitor, whereas the p-i-n diodes

are mounted in the middle layer in each unit cell. The bottom layer is completely metal laminated. Air spacers are inserted between the successive layers. When zero bias voltage has been applied across the active components, the diodes exhibit capacitive impedance and resonate at a particular frequency. Except the resonance point, the diode layer behaves as a transmitter, allowing the incident EM wave to transmit and get reflected from the bottom ground plane, thereby resulting in a wideband absorption in lower frequency range. On the contrary, when the p-i-n diodes are forward biased, continuous current flows through the diode layer, thereby behaving as a reflective surface (as explained in Ghosh and Srivastava 2016b). This virtual ground plane restricts the RF signal from further transmission and reduces the effective substrate height of the geometry under ON state. Since the broadband absorption phenomenon depends on the quarter-wavelength condition, the operating bandwidth gets shifted while changing the biasing condition. Thus, −10 dB reflection coefficient band is observed in the ranges of 0.85–1.88 GHz and 2.66–5.23 GHz under OFF state and ON state, respectively.

To verify the proposed design concept, the structure has been fabricated and biasing circuitries have been provided from the bottom side. Two separate metallic layers have been considered, one for connecting the positive terminal and other for negative terminal of the external voltage. The cathodes and anodes of the diodes are connected to these terminals through metallic vias, as described in Zhang et al. (2012). During measurement, the sample prototype exhibits a good agreement between the simulated and measured response. Angular stability under transverse electric (TE) mode has also been observed till 30° angle of incidence.

The above switchable geometry has two limitations, viz., the topology is polarization-sensitive and the biasing configuration is complicated to realize. Later, Ghosh and Srivastava have presented an improved design by resolving the above shortcomings (Ghosh and Srivastava 2017b). The proposed switchable broadband RAS consists of a lumped resistor-loaded-square loop topology in the top layer, whereas the middle layer consists of diode-embedded cross-dipole geometry, as shown in Fig. 14a, b, c. The principle of operation remains same as before: the middle diode layer acts as an adaptive ground plane based on the supply voltage. During OFF state, the diode layer behaves as a transmitter and during ON state, reflective behavior is exhibited. This results in a switchable response covering the absorption bandwidth across S band (2–4 GHz) and C band (4–8 GHz) in one state, whereas X band (8–12 GHz) under another biasing state. The simulated response is depicted in Fig. 14d.

Since both the top and middle layer geometries are four-fold symmetrical, the overall design exhibits identical response under different polarization angles. Further, the biasing circuitries have been realized in the same diode layer with narrow-width metal lines and inductors with requisite values. This effectively reduces the biasing complexity as well as enhances the design usability. The photograph of the fabricated prototype is shown in Fig. 14e. The prototype, while measured under normal incidence, exhibits two different absorption spectra under different biasing conditions: 3.18–12.78 GHz bandwidth (below −10 dB reflectivity level) under ON state and 1.91–4.89 GHz under OFF state. The measured response has a good consistency with the simulated result. Thus, broadband switchable microwave absorber has been demonstrated.

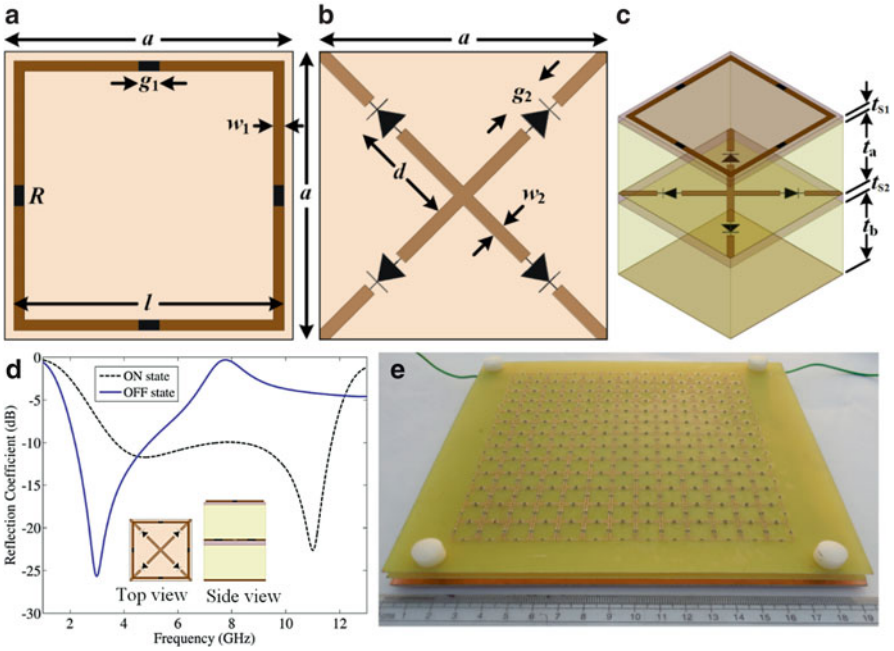


Fig. 14 Polarization-insensitive broadband switchable absorber. Proposed design: (**a**) top layer, (**b**) middle layer, and (**c**) perspective view of the unit cell geometry. (**d**) Simulated reflection coefficient under different biasing states. (**e**) Photograph of the fabricated prototype. (Reproduced from Ghosh and Srivastava (2017b) with permission. Copyright 2017, IEEE)

Several other wideband switchable metamaterial FSS-based RAS designs have also been presented, covering different absorption frequency bands under different biasing states (Lee et al. 2017). Their underlying concepts are aligned with the above presented papers, and have used p-i-n diodes for exhibiting switchable responses in microwave frequency bands.

Tunable Absorbers

The third kind of reconfigurable absorber is the tunable absorber, where the absorption frequency or/and the absorption bandwidth can be tuned over a large number of states based on the supply voltage. Thus, the tunable absorber can be considered as an extension of the switchable absorber, in terms of number of intermediate states. However, the principle of operation of a tunable absorber is much different from that of a switchable absorber.

Electronically regulated switchable absorbers mostly exploit the two different states of the p-i-n diodes; ON state is attained by providing the forward bias voltage, whereas the OFF state is obtained by applying the zero/reverse bias voltage. On the contrary, tunable absorbers generally employ varactor diodes for regulating the

absorption frequency in a continuous way. These varactors exhibit a nonlinear variation in the junction capacitance under the reverse bias voltage that controls the precise tuning of the absorption band. P-i-n diodes are also used in the tunable absorbers to regulate the absorption bandwidths (instead of absorption frequencies), but those diodes are employed under the reverse bias voltage only (below the threshold voltage level). With precise tuning of the applied voltage, the resistance value across the p-i-n diodes can be tuned from very large to very small, thereby resulting in a tunable bandwidth.

In view of the current scenario, tunable absorbers have manifold applications. They can be used in several sectors where the absorption frequency and/or bandwidth need to be regulated. They can be used in hybrid radome for reducing the RCS (Bahret 1993). When the FSS radome is implanted in an aircraft, the resonant frequency can electronically be tuned through the external bias voltage. Thus, the aircraft can absorb the EM wave incoming from the enemy radar, while it can communicate with the friendly radar simultaneously. Tunable narrow-band absorbers, operating across a wide frequency range, can replace a broadband absorber in a few applications. They are also useful in radar camouflage and deception roles, low frequency EMC measurement facilities, telematics, and tagging applications (Chambers and Ford 2000). Furthermore, tunability can compensate FSS fabrication and installation errors not possible with passive RASs.

P-i-n Diode Regulated Tunable Absorbers

Various designs of tunable absorbers have been presented so far. Chambers and Ford haves first investigated the concept of tunable radar absorbers by exploiting variable resistances in the top layer of a Salisbury screen (Chambers and Ford 2000). Salisbury screen is a single-band absorber structure where a resistive sheet is placed at a specific height above a ground plane (Salisbury 1952). The incident EM wave gets absorbed by the top resistive sheet at a particular frequency (where the spacer thickness matches the quarter-wavelength condition). Then, by varying the top layer surface resistance, a dynamically tunable RAS geometry has been realized.

Later, Tenannt and Chambers have improved the design topology, where the conventional resistive layer of the Salisbury screen has been replaced by an active FSS geometry, controlled by p-i-n diodes (Tennant and Chambers 2004). The overall structure consists of a periodic grid of bow-tie dipole elements patterned on a dielectric substrate, where the diodes are embedded across the gaps in the centre. The structure is terminated by a ground plane. By regulating the bias current, the reflectivity response can be controlled over the frequency band from 9 to 13 GHz. In the following years, the tunability operation has further been explored by Mias in several FSS-based applications (Mias 2005), where varactor diodes are embedded in the biasing grids to result in tunable high impedance surfaces, and filtering responses. Subsequently, other tunable RAS geometries have been started developing across different frequency ranges, exploring various applications.

In 2016, J. Li et al. have developed a broadband tunable FSS-based RAS, where the p-i-n diodes are loaded between two double-semicircle dipole elements (Li et al. 2016a). Teflon and honeycomb core are used as the constituent dielectric substrate

with a metal sheet acting as a ground plane. When the supply voltage across the diodes are varied in the range of 0 to 0.95 V across each diode, the resistance offered by the diodes are varying from large value to small value. This corresponds to a continuous variation in the reflectivity level between 2 and 11.3 GHz with a varied bias voltage.

To implement the concept, the proposed design has been fabricated and p-i-n diodes are mounted accordingly. Bias lines have been drawn across the structure to provide the necessary voltages across the diodes. During varying the bias voltage in the range of 0.0–0.95 V, the diodes exhibit a dynamically tunable resistance value and gives rise to a continuously varying absorption response. In the absence of bias voltage, the structure exhibits a reflective behavior, whereas the reflectivity curve shows multiple nulls/single null at different frequency bands with increasing the bias voltage value. The measured responses are also in agreement with the simulated results, thereby establishing the proposed tunable FSS-based RAS concept.

In a similar way, P. Kong et al. have developed another tunable broadband FSS-based RAS (Kong et al. 2014), where both p-i-n diodes and varactors have been implemented for resulting in a tunable absorption response. The components are alternatively mounted between the successive unit cell topologies, as depicted in Fig. 15a, and the voltages across them are separately varied for dynamically tunable bandwidth. The variable resistance is attained by using the p-i-n diode, whereas the variable capacitance is realized by mounting the varactor across the gaps. This results in maintaining good impedance matching across a wide frequency range and a broadband tunable absorber covering from 1 to 5 GHz has been obtained, as depicted in Fig. 15b.

A sample prototype has been fabricated with a size of 180 mm × 180 mm to validate the proposed design concept. The p-i-n diodes and varactors are mounted alternatively across the structure, as depicted in Fig. 15c. Independent bias lines have been used for separately regulating the diode components. The forward bias voltage has been applied across the former device, whereas the latter device works under reverse bias. By precisely regulating the bias voltages, variable resistance and capacitance are obtained from the p-i-n diodes and varactors, respectively. The

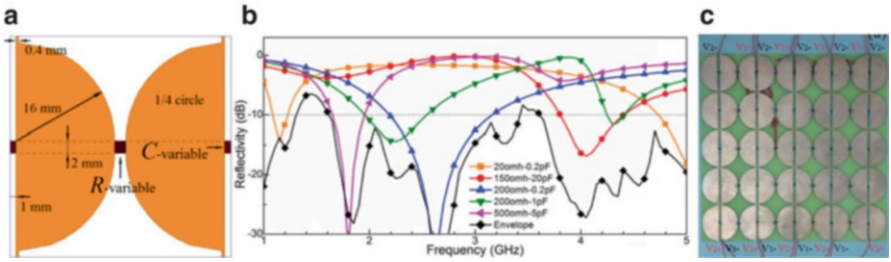


Fig. 15 Tunable absorber using p-i-n diode and varactors. (**a**) Top view of the unit cell pattern, (**b**) simulated reflection coefficient of the structure under different resistance and capacitance. (**c**) Photograph of the fabricated prototype. (Reproduced from Kong et al. 2014. Copyright 2014, Optical Society of America)

measurement has been carried out by the arch test method and the reflectivity is found to have below −10 dB over the frequency range from 1.3 to 4.7 GHz. The measured responses are in similar trend with those of the simulated responses, confirming the proposed mechanism. A tunable RAS structure, exploiting both the p-i-n and varactor diodes, thus has been successfully demonstrated.

The above designs presented on tunable FSS-based RSA structures are asymmetrical in nature and operate for single polarization angle only. For practical applications, a polarization-independent tunable absorber is preferred due to its ability to absorb any arbitrary polarized EM wave. H. Wang et al. have developed such a polarization-insensitive tunable absorber, based on four-axial symmetrical topology (Wang et al. 2016). The geometry is patterned on a multi-layer geometry, as shown in Fig. 16a. FR4 substrate and honeycomb core are used as the intermediate dielectrics, with metal sheet as the ground plane. The top layer periodic pattern is made from four quarter circles, whereas p-i-n diodes are symmetrically mounted across them. While varying the diode voltage, the reflection coefficient envelope is found to have below −8 dB in the frequency range of 1.5 to 8.0 GHz, depicted in Fig. 16b.

The fabricated prototype has an overall size of 180 mm × 180 mm, where the diodes are mounted across the gaps, as shown in Fig. 16c. All the p-i-n diodes are connected in series ensuring the diodes in each unit cell receive the same bias voltage, conforming to the simulation. Inductors of requisite values are embedded diagonally between the successive metallic plates to attain the RF-isolation, without impacting the RAS's operation. The measured reflectivity has an envelope value below −10 dB covering a frequency range of 1.6 to 4.3 GHz and 5.4 to 8.0 GHz. The polarization-insensitivity as well as the broadband tunability in the design have been successfully established.

Several other wideband polarization-independent metamaterial FSS-based tunable RAS designs have also been presented, exploiting the p-i-n diodes (Zhao et al. 2020). The resistance values of the diodes are regulated by applying different bias voltages and subsequently, the absorption responses are obtained over a wide frequency band in microwave regime.

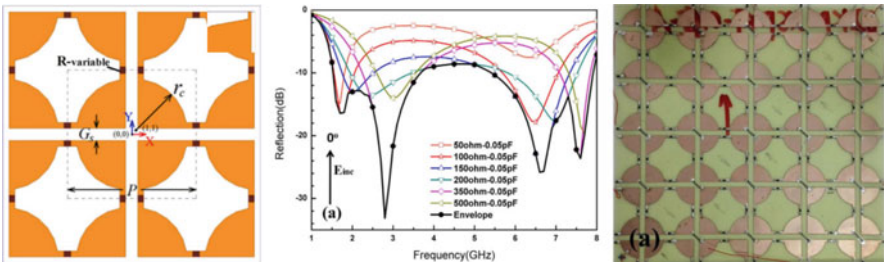


Fig. 16 Polarization-insensitive tunable absorber using p-i-n diode. (**a**) Top view of the unit cell pattern, (**b**) simulated reflection coefficient of the structure under different resistor values. (**c**) Photograph of the fabricated prototype. (Reproduced from Kong et al. 2014. Copyright 2016, Springer Nature Publishing)

Varactor Regulated Tunable Absorbers

Varactor diodes are also used in metamaterial FSS-based tunable absorber structures, in particular, for varying the absorption frequencies rather than the absorption bandwidths. The depletion region width across the p-n junction of a diode is associated with its junction capacitance through a non-linear relationship. By applying an external voltage, this junction capacitance can be precisely regulated. If the varactor diode is embedded in FSS geometry, by varying the bias voltage across the diodes, the absorption frequency can be widely tuned, thereby giving rise to a tunable absorber.

Since the last few years, several varactor-controlled tunable absorbers have been reported in literature. H. Yuan et al. have developed a single-band metamaterial tunable absorber structure based on a microstrip line resonator loaded with a varactor diode, as shown in Fig. 17a (Yuan et al. 2015). Two biasing lines are also printed at the top and bottom sides of the unit cell. Since the unit cell pattern is asymmetrical in nature, the topology will exhibit tunable operation for single polarization only. When the line resonator is oriented along the electric field direction, the surface current propagates through the geometry as well as the varactor. The junction capacitance of the diode then contributes in the structure resonance that can be dynamically varied with the supply voltage. With the variation in the reverse bias voltage, the absorption frequency can be continuously tuned from 8.48 to 8.92 GHz, as depicted in Fig. 17b.

The proposed design has been fabricated and SMV1231–011 varactors are soldered across the gaps. Figure 17c depicts the photograph of the sample prototype. The junction capacitance value corresponding to the varactor has a wide range between 2.35 pF at 0 V to 0.46 pF at 15 V. Thus, when the supply voltage has been varied, the junction capacitance value has been changed, and correspondingly the absorption frequency has been tuned from 8.48 GHz (during 0 V) to 8.92 GHz (during 15 V). The measured responses are in reasonable agreement with the simulated results under normal incidence, thus validating the proposed tunable RAS design concept.

A similar tunable absorber has been presented in Ma et al. (2016), where two ELC resonators are embedded in a single unit cell and one varactor diode is mounted

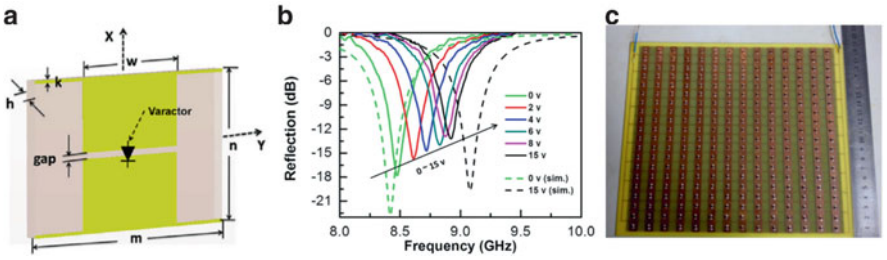


Fig. 17 Single-band tunable absorber using varactor diode. (**a**) Schematic view of the unit cell geometry, and (**b**) simulated reflection coefficients of the structure under different bias voltages. (**c**) Photograph of the fabricated prototype. (Reproduced from Yuan et al. (2015) with permission. Copyright 2015, AIP Publishing)

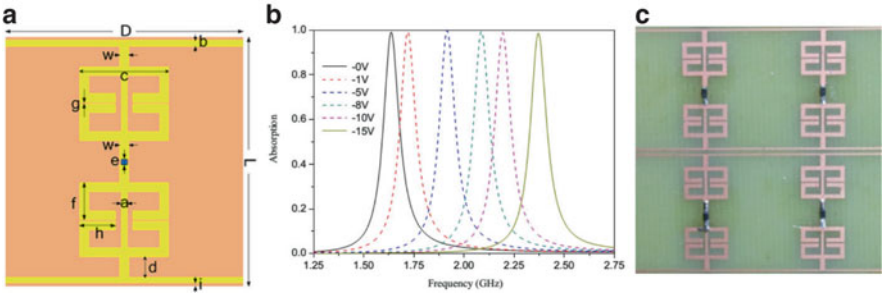


Fig. 18 Single-band tunable absorber using varactor diode. (**a**) Top view of the unit cell geometry, (**b**) simulated absorptivities of the structure under different bias voltages. (**c**) Photograph of the fabricated prototype. (Reproduced from Ma et al. (2016) with permission. Copyright 2016, Elsevier)

between the two resonators, as shown in Fig. 18a. The diode voltage has been regulated through the biasing lines, drawn at the upper and lower sides of the unit cell. While increasing the reverse voltage from 0 V to −15 V, the depletion width becomes larger and results in a decreased capacitance. This leads to a blue shift in the resonance frequency. The absorption band is thus found varying from 1.63 to 2.37 GHz, producing a tunable range of 45.39%, as depicted in Fig. 18b. The proposed design has also been fabricated and similar biasing configuration has been used to regulate the varactor diodes. Figure 18c shows the photograph of the fabricated prototype. The sample has been measured under different reverse bias voltages and the absorption responses are found in good agreement.

The above presented tunable RAS designs are polarization-dependent, and will exhibit the desired performance for a particular orientation of EM wave only. J. Zhao et al. have resolved this limitation by developing one polarization-insensitive tunable absorber (Zhao et al. 2013). Initially, the design has been constructed for single polarization and later, four-fold symmetry has been incorporated to enhance its usability. Figure 19a illustrates the unit cell topology of the polarization-sensitive tunable RAS, where the varactors diodes are mounted at the top layer between two resonators to exhibit wide frequency tuning. In order to make the absorber insensitive to wave polarization, the basic units are placed orthogonally, as demonstrated in Fig. 19b. In both the geometries, the biasing network has been provided from the back side of the structure, and thus requires an additional feeding network and an insulating layer between the dielectric substrate and the ground plane.

The voltage across the diodes are varied through this feeding network and the design exhibits a tunable absorption response covering from 4.35 to 5.85 GHz, while varying the reverse voltage from 0 to −19 V. The measured responses under different bias voltages are shown in Fig. 19c. As the experimental results are found to be consistent with those of the simulated responses, the frequency tunable behavior has been successfully validated in the proposed design.

The previously reported articles have employed different types of biasing configurations to provide the supply voltage across the varactors, either through using

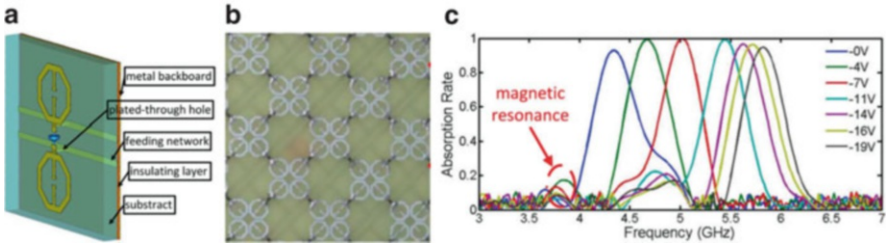


Fig. 19 Single-band tunable absorber using varactor diode. (**a**) Perspective view of unit cell geometry of polarization-sensitive tunable RAS. (**b**) Photograph of the fabricated polarization-insensitive sample prototype, and (**c**) its measured absorptivities under different bias voltages. (Reproduced from (Zhao et al. 2013). Copyright 2013, IOP Publishing)

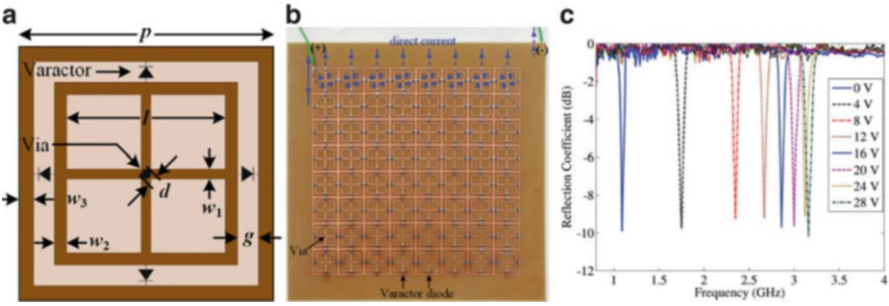


Fig. 20 Polarization-insensitive single-band tunable absorber. (**a**) Top view of the unit cell geometry. (**b**) Photograph of the fabricated prototype, and (**c**) its measured responses under bias voltages. (Redrawn from Ghosh and Srivastava 2017c)

dedicated bias lines or from the bottom surface. But, it would have been convenient if the effect of the biasing lines can be reduced; one plausible solution is to implement the embedded biasing circuit instead of the external biasing grid.

The design proposed by S. Ghosh and Srivastava (2017c) has been based on the above concept, where the biasing of the varactor diodes has been achieved without using any external circuitries. Figure 20a depicts the top layer of the unit cell topology that consists of two concentric square loops, and the diodes are symmetrically embedded between them. The inner loop is connected with the back metal plane through a metallic via drilled at the centre, whereas the outer loop is joined with the neighboring unit cells. When the bias voltage is varied from 0 to 28 V, the junction capacitance corresponding to the varactors is decreased, and the resonance frequency is increased from 1.08 to 3.16 GHz, as observed from Fig. 20b. Thus the tunable range is nearly 98.11% with respect to the centre frequency. Due to four-fold symmetry, the proposed design topology is polarization-insensitive as well as angularly stable for different polarizations.

While going through the surface current distribution, it is found that the current flows through the square loops and the varactor diodes continuously, without

producing any parasitic capacitance. Even if any gap capacitance is generated, the same appears in parallel with the large valued varactor capacitance, and subsequently gets nullified. Therefore, the resonance frequency of the design is mostly governed by the junction capacitance of the diode and the equivalent inductance of the metallic patterns. This gives rise to a wideband tuning range, by varying the diode capacitance through the external bias voltage. It is also observed that the metallic vias present at the centre of the unit cell doesn't contribute in generating the absorption frequency. Instead, it is primarily used to supply the required bias voltage from the ground plane.

To validate the proposed design, a prototype has been fabricated using the PCB technique and diodes from BB837 are mounted across the gaps. Since the unit cells are interconnected with one another through metallic patters, the cathodes of all the varactors are connected through parallel circuit. Thus, single bias voltage applied at far end will be sufficient to regulate the diodes simultaneously. The anode terminals of the diodes are connected to the ground plane through the metallic vias. Therefore, when the positive and negative ends of the supply voltage are connected to the top layer and ground plane, respectively, the structure exhibits a tunable response with the variation of the external voltage. Figure 20c displays the photograph of the fabricated prototype and the biasing configuration. The sample has been measured using free space technique and a wide tuning range covering from 1.09 to 3.16 GHz has been observed for a voltage variation of 0 to 28 V. The wide tuning range as well as embedded biasing circuitries in the FSS-based tunable RAS have been attained in this work. The proposed design is also polarization-insensitive as well as angularly stable under both TE and transverse magnetic (TM) polarizations.

A similar frequency tunable absorber has been developed using fishnet-like resonator, where the embedded biasing configuration has been used to reduce the design complexity (Kim et al. 2016). The proposed absorber utilizes a single varactor in each unit cell, and the metallic patterns surrounding the diode are used for generating the absorption frequency as well as biasing the varactor diode model. While regulating the reverse bias voltage from 7 to 19 V, the absorption frequency is tuned between 3.96 and 5.29 GHz. The structure has also been fabricated using PCB technique and its responses are measured under different biasing conditions. However, the proposed design is asymmetrical in nature and is able to exhibit the tunable response for single polarization only.

Another polarization-sensitive tunable absorber has been designed in Zhu et al. (2015), where two varactors are loaded in a split square loop printed in each unit cell geometry. By regulating the diode voltage between 0 and −12 V, the absorption frequency can be varied in the range of 5.12 to 5.68 GHz. The biasing lines have been provided from the bottom surface, similar as described in Zhao et al. (2013), to precisely regulate the absorption responses. The structure has been fabricated as well as measured under normal incidence, thereby validating the proposed concept. Later, multiple concentric metallic loops have been designed in the unit cell, and diodes have been mounted across the gaps in each of the loops. Thus, three narrow-band absorption responses are generated, which can be independently tuned through bias voltages applied across different sets of varactor diodes. However, the concept has

been limited to numerical simulation only, and no biasing mechanism has been illustrated in support of the design.

J. Yuan et al. have first experimentally demonstrated a multi-band tunable absorber, where one band exhibits the tunable response and the other absorption band is constant (Yuan et al. 2014). The proposed unit cell geometry consists of a central resonator, where a varactor diode is mounted in the middle and horizontal bias lines are connected at two ends. This part of the resonator generates the tunable response. The other absorption frequency is originated from the split ring resonator (SRR) section, which remains almost constant throughout the tuning voltage. Figure 21a depicts the unit cell topology of the proposed dual-band tunable RAS design. While tuning the bias voltage across the varactor, the diode junction capacitance gets varied that results in the shift of one of the absorption frequencies. A tunable frequency range is obtained, as shown in Fig. 21b, covering from 2.69 GHz to 3.37 GHz, while changing the reverse voltage from 0 to −15 V. The other absorption band, realized due to the SRR geometry, remains constant around 4 GHz.

The photograph of the fabricated prototype is displayed in Fig. 21c, where the biasing configuration is also illustrated. The unit cell geometries are connected in each row through the horizontal biasing lines, which are joined with one another at the side through two vertical metal lines. When the supply voltage has been provided across these two lines, the varactors are getting biased, and the tunability operation has been realized in the fabricated prototype. The measured response is also in good agreement with the simulated response under normal incidence.

Although the previous metamaterial RAS is designated as a dual-band tunable absorber, the geometry exhibits only one tunable absorption band and the other absorption frequency remains constant throughout the tuning states. Further, the topology is polarization-sensitive due to its design asymmetry. In 2017, S. Ghosh and Srivastava have developed an improved dual-band polarization-independent FSS-based tunable absorber, resolving the above limitations (Ghosh and Srivastava 2017d). Initially, two single-band tunable absorber structures have been independently designed, and later combined together to exhibit a dual-band tunable response.

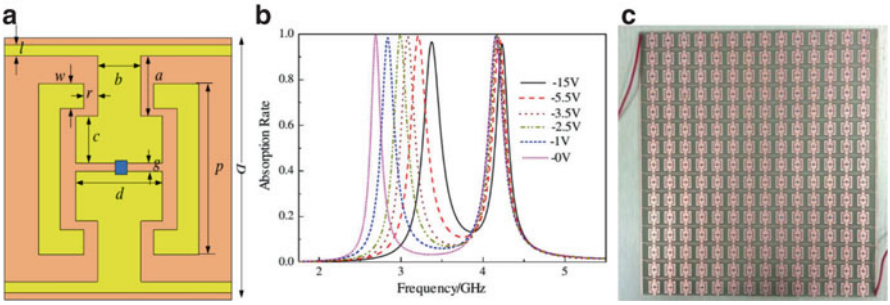


Fig. 21 Dual-band tunable absorber. (**a**) Top view of the unit cell geometry. (**b**) Simulated absorptivities of the structure under different reverse bias voltages. (**c**) Photograph of the fabricated prototype. (Reproduced from Yuan et al. (2014). Copyright 2014, The Electromagnetics Academy)

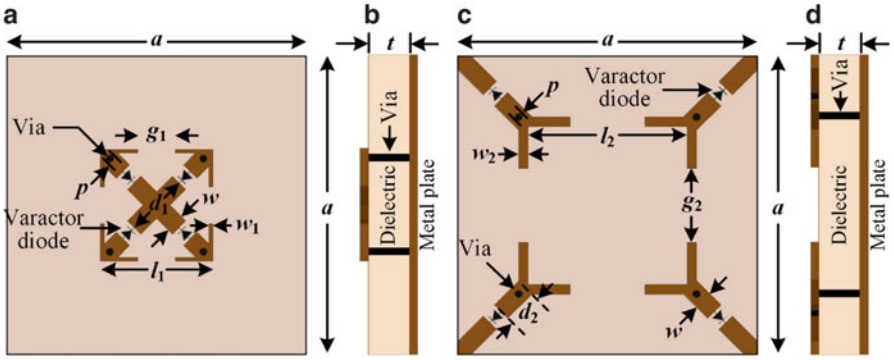


Fig. 22 Polarization-insensitive single-band tunable absorber I: (**a**) top view, and (**b**) side view. Polarization-insensitive single-band tunable absorber II: (**c**) top view, and (**d**) side view. (Reproduced from Ghosh and Srivastava (2017d) with permission. Copyright 2017, IEEE)

Figure 22a, b depict the top view and side view, respectively, of the unit cell topology of the first single-band tunable absorber (marked as 'absorber I'). The proposed geometry consists of a rotated cross-dipole inscribed inside a split square loop. The top metallic pattern is connected with the ground plane through four metallic vias present at the corners, whereas varactors are mounted across the gaps in the cross-dipole geometry. When the EM wave is incident on the structure, the capacitance is resulted from the varactor capacitance only, as there is no distributed capacitance originated from the topology. Thus, the resonance frequency can be tuned for a wide frequency range covering from 0.98 to 3.76 GHz by varying the reverse bias voltage from 0 to −28 V. The design is polarization-insensitive as well as angularly stable.

In a similar way, the second single-band tunable absorber (designated as 'absorber II') has been developed such that the former topology can be engraved inside the later one, while maintaining the tunable response. The unit cell of the second design, as depicted in Fig. 22c, d, is comprised of a split square loop connected with adjacent unit cells along the diagonals. Vias are engraved at the corners of the square loop, whereas varactors are mounted across the gaps on the diagonals. When the junction capacitance of the diodes are varied with the reverse bias voltage, the absorption frequency is increased from 0.88 GHz at 0 V to 3.33 GHz at 28 V, thus producing a tunable range of 278.41% with respect to the lower frequency. This topology is also polarization-independent as before.

When the above two geometries (absorber I and II) are embedded in single unit cell, a dual-band tunable absorber has been resulted, as depicted in Fig. 23a. The geometries are constructed such that both the single-band structures can independently exhibit tunable responses, while varying the corresponding bias voltage. At the same time, their polarization-insensitive characteristics will prevail in the combined design topology.

To realize the proposed design, one sample prototype has been fabricated and proper biasing circuits have been installed around the structure. Two vertical metal lines are

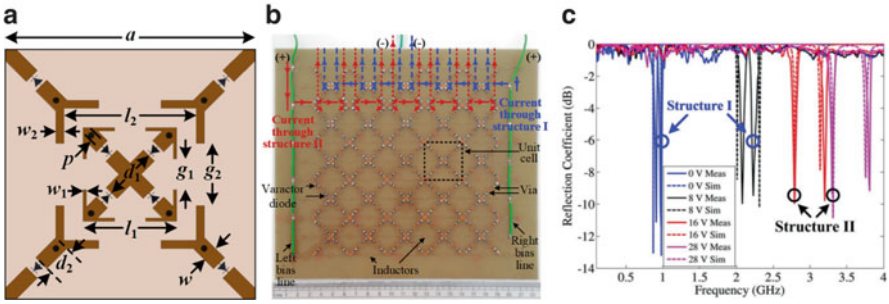


Fig. 23 Polarization-insensitive dual-band tunable absorber. Unit cell geometry: (**a**) top view, and (**b**) side view. Simulated reflection coefficients for (**c**) different bias voltages across structure I, and (**d**) different bias voltages across structure II. (Reproduced from Ghosh and Srivastava (2017d) with permission. Copyright 2017, IEEE)

printed at two sides (one left and another right) of the geometry. Narrow-width metal lines are horizontally drawn across the unit cells such that the lines passing through the inner square are connected to the right side bias line, whereas the metal lines passing through the outer square are connected to the left side bias line. Another set of connection is provided to the ground plane. When the right (left) sided bias line and the ground plane are connected between the two ends of a supply voltage, connection is established across the varactor diodes mounted in the inner (outer) square loops. Thus, by varying the bias voltage, the absorption frequency corresponding to the particular geometry can be regulated. Figure 23b illustrates the photograph of the fabricated sample and its biasing configuration. With the proper biasing network, the fabricated sample has been measured and it is found that the structure exhibits different sets of tuning responses, depending on the applied voltages. The measured responses are also found to be in reasonable agreement with the simulated responses, under different situations, as depicted in Fig. 23c. Thus, a dual-band polarization-insensitive metamaterial FSS-based tunable RAS design has been realized.

Magnetically Regulated RAS

Although electronically controlled active RAS designs are widely reported, other types of reconfigurable structures have been explored across different frequency ranges. One of the other techniques is magnetic actuation, where the absorption response can be regulated by using ferrite-based substrates. By varying the magnetic field intensity, the ferrite material characteristics (such as real and imaginary parts of relative permeability) can be varied, thereby resulting in a tunable or switchable absorber structure.

Huang et al. have first experimentally demonstrated a magnetically tunable metamaterial absorber, where a ferrite substrate has been integrated as a substrate or superstrate into a conventional planar absorber (Huang et al. 2014). The unit cell is made of an ELC resonator printed on a grounded dielectric substrate, to realize a

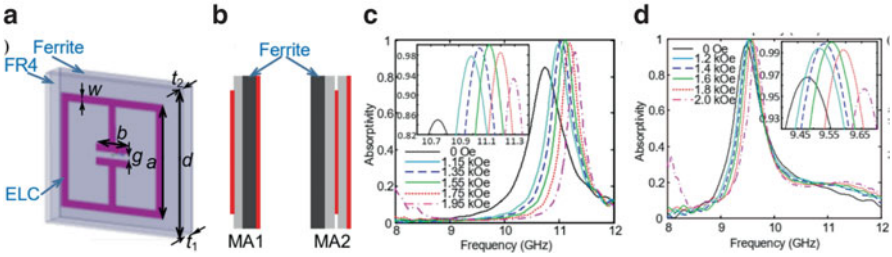


Fig. 24 Ferrite-based tunable metamaterial absorber. (**a**) Unit cell geometry, (**b**) side views of two MA models. Simulated absorptivities under different magnetic fields for (**c**) MA1 model and (**d**) MA2 model. (Reproduced from Huang et al. (2014). Copyright 2014, Optical Society of America)

passive absorber design. FR4 has been used as the constituent dielectric, whereas copper metal is used as the complete ground plane. Two different models have been considered; in the first case (MA1), one ferrite layer has been inserted between the FR4 and the ground plane, whereas the ferrite layer has been employed as a superstrate in the second case (MA2). Figure 24a, b illustrate the geometric configurations of the proposed tunable absorber structures.

When the applied magnetic field intensity is varied, the permeability of the ferrite layer is shifted that results in tuning the absorption frequency of the structure due to the change in background media parameters. For the first case (MA1), the resonant frequency gradually moves from 10.99 to 11.29 GHz with the variation in the magnetic field from 1.15 KOe to 1.95 KOe. Similar frequency shift between 9.52 and 9.66 GHz has been found for MA2 with the similar variation in the magnetic field intensity. Figure 24c, d depict the simulated absorption properties under different magnetic fields for MA1 and MA2, respectively. During measurement, the samples are fabricated by PCB technique and the ferrite layer is then combined with the sample in two different configurations. The prototypes are measured in a rectangular waveguide and the results are found in good agreement with those of the simulated responses under normal incidence. Thus the proposed magnetically tunable metamaterial RAS design has been successfully validated.

Later, Lei et al. have demonstrated another tunable metamaterial absorber based on ferromagnetic resonance, where the ferrite layer has been used as the constituent dielectric instead of using it as a substrate or superstrate (Lei et al. 2016). The proposed geometry, as depicted in Fig. 25a, consists of an array of ferrite rods adhered on a thin metallic ground plane. Yttrium iron garnet (YIG) has been chosen as the ferrite material, whereas copper has been considered as the ground plane. When the EM wave is incident on the structure, the ferrite rods and the ground plane strongly couple with the magnetic and electric field components, respectively, and result in a ferromagnetic resonance. By varying the applied magnetic field, both the effective permittivity and permeability can be tuned over a wide frequency range, thereby generating a tunable absorber design. Figure 25b shows the simulated reflection coefficient of the proposed design, where the resonance dip shifts from 8.97 to 10.02 GHz as the magnetic field is increased from 2000 Oe to 2400 Oe.

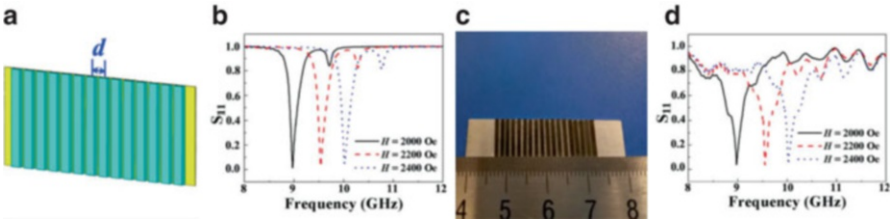


Fig. 25 Ferrite-based tunable metamaterial absorber. (**a**) Unit cell geometry, and (**b**) its simulated reflection coefficient under different magnetic fields. (**c**) Photograph of the fabricated sample, and (**d**) its measured reflection coefficient under different magnetic fields. (Reproduced from Lei et al. (2016) with permission. Copyright 2016, AIP Publishing)

The geometry has been fabricated and its scattering parameters are recorded through waveguide measurement technique. Figure 25c, d show the photograph of the prototype and its measured responses, respectively. The close resemblance between the simulated and measured response confirms the magnetically tunable behavior of the proposed metamaterial RAS design.

In 2016, another group of researchers have demonstrated a similar type of ferrite-based tunable metamaterial absorber, where periodic metallic strips have been combined with a metal-backed ferrite substrate (Li et al. 2016b). While varying the external magnetic field, the resonance frequency has been tuned in an ultra-wide frequency range from 0.2 to 7.6 GHz, thereby exhibiting a tunable RAS.

Later, a dual-band tunable metamaterial absorber has been presented in Wang et al. (2018), where both the absorption bands can be tuned by the applied magnetic field and modulated by the permittivity of its components. The proposed geometry consists of eight cuboid ferrite places on a metal ground plane. Yttrium iron garnet (YIG) has been used as the ferrite material. When the incident EM wave and applied magnetic field act on the substrates simultaneously, the ferrites produce strong ferromagnetic resonances and shows two distinct absorption peaks. While varying the applied magnetic field from 2500 Oe to 2800 Oe, two resonance peaks are drifted to different frequencies, exhibiting a dual-band tunable behavior. The sample has also been fabricated and the measured circumstances are similar as the simulation ones.

The work has further been extended by the same group of researchers, by presenting a broadband tunable metamaterial absorber, using the similar concept (Wang et al. 2019). The geometry is made from eight U-shaped ferrite particles deposited on a ground plane. By optimizing the geometric dimensions, the proposed design exhibits a wide absorptivity above 70% in the frequency range from 8.55 to 10 GHz under a magnetic field of 2700 Oe. It is observed that the absorption spectrum is blueshifted with the applied field intensity increasing from 2500 Oe to 3000 Oe. During variation, the absorption bandwidth remains constant, while maintaining the absorptivity above 60% throughout the bandwidth. The structure has been fabricated and measured in a waveguide environment, and the measured responses have good consistency with those of the simulated ones.

A few other magnetically reconfigurable RAS designs have been presented, exhibiting manifold characteristics (Yang et al. 2012). Most of the geometries are numerically simulated and/or experimentally demonstrated, providing a viable solution to develop active absorbers using magnetic actuation.

Mechanically Regulated RAS

Mechanically actuation is another technique to realize active absorber designs, where different types of mechanical forces, such as stress-strain, pressure etc. can be applied to regulate absorption responses. A few articles have also been reported in this domain across microwave frequencies.

Zhang et al. have demonstrated a mechanically stretchable and tunable metamaterial absorber structure comprising of ceramic bricks adhered on a thin conductive rubber layer (Zhang et al. 2015). Figure 26a shows the schematic view of the proposed design, whereas the fabricated prototype and its measurement set up are illustrated in Fig. 26b, c, respectively. When the EM wave is incident on the geometry, oscillating magnetic dipoles are generated inside the dielectric brick, indicating the occurrence of Mie resonance behind the absorption behavior. The conductive rubber plate restricts the transmission response as well as couples with the dielectric resonator. Thus, a narrow absorption peak is observed at 11.56 GHz under no stress condition.

When a uniaxial stress has been applied along the magnetic field direction (y-axis), the conductive rubber and dielectric resonators' interspacing have been

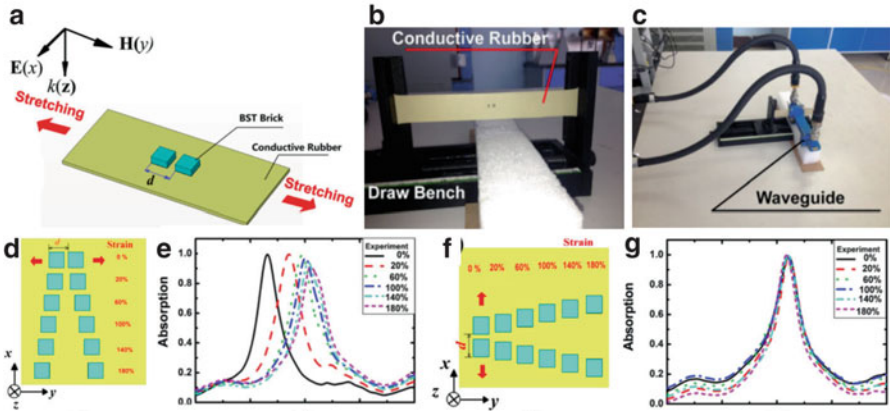


Fig. 26 Mechanically tunable metamaterial absorber. (**a**) Schematic of the proposed geometry, (**b**) photograph of the fabricated prototype, and (**c**) experimental set up for measurement. (**d**) Schematic of stretchable absorber under uniaxial stress along H-field (y-axis), and (**e**) corresponding absorption spectra as a function of strain. (**f**) Schematic of stretchable absorber under uniaxial stress along E-field (x-axis), and (**g**) corresponding absorption spectra as a function of strain. (Reproduced from Zhang et al. (2015) with permission. Copyright 2015, AIP Publishing)

stretched. This leads to a variation in the mechanical strain from 0% to 180%, and subsequently the absorption peak increases from 11.15 to 11.56 GHz. With increasing strain, interspacing effect becomes weaker, thereby leading to the increasing absorption frequency. Figure 26d, e display the arrangement of stress on the geometry and its corresponding change in the absorption response. On the contrary, the effect of the uniaxial stress along E-field direction (*x*-axis) has also been studied, as shown in Fig. 26f, g, and it is observed that the absorption profile remains constant with the strain variation. Unable to oscillate the magnetic dipoles (which are transversely coupled) with this particular stress, the geometry exhibits no dynamic change in this orientation. Both the simulated responses are well supported by the measured responses under normal incidence. Thus, the proposed active RAS design exhibits a mechanically control selective dynamic behavior under different polarizations.

Later, H. Jeong and Lim have presented a similar article on stretchable absorber (Jeong and Lim 2017), where different stretchable dielectric substrates and top metallic patterns are used in the geometry. The proposed design, under different stress conditions applied along different directions, exhibits a frequency tunable absorption behavior. While stretched along y-axis, the absorption frequency shifts by a wide range, between 10.4 and 11 GHz. On the contrary, there is no considerable shift under *x*-axis stress condition. The concept has also been successfully validated by fabrication and measurement.

Another technique of mechanically actuated frequency reconfigurable metamaterial absorber has recently been proposed in Kim et al. (2019). The proposed design consists of a top metallic pattern and three dielectric layers vertically stacked one another, viz., one FR4 layer (top dielectric), one variable thickness air spacer (middle dielectric), and another FR4 layer (bottom dielectric). The overall geometry is terminated by a ground plane for zero transmission. Figure 27a shows the schematic of the unit cell geometry. The geometry, based on the metallic patterns and the dielectric substrates, exhibits a narrow-band absorption response. The air spacer thickness can be regulated from back by a mechanical actuator. This effectively changes the overall substrate thickness as well as the absorption frequency. When the spacer thickness is varied from 17 to 26 mm, the resonant frequency

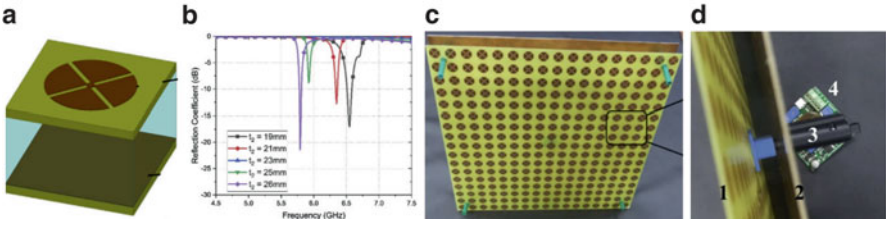


Fig. 27 Mechanically actuated tunable metamaterial absorber. (**a**) Schematic of the proposed unit cell geometry, and (**b**) simulated reflection coefficient responses under different thicknesses. (**c**) Photograph of the fabricated prototype, and (**c**) magnified pictures of the actuator and linear actuator control board. (Reproduced from Kim et al. (2019) with permission. Copyright 2019, Elsevier)

gradually decreases from 6.96 to 5.96 GHz, as plotted in Fig. 27b. The design remains polarization-insensitive during the thickness variation.

The proposed design has been constructed using the PCB technique, and a mechanical actuator has been attached at the back side of the bottom FR4 layer. The actuator, under the external voltage, can push or pull the bottom FR4 layer that effectively changes the air spacer thickness. With different actuator positions, the overall thickness gets varied and results in a tunable absorption frequency. The fabricated prototype and the mechanically regulated tunings mechanism are illustrated in Fig. 27c, d, respectively. The measured responses of the fabricated sample are also found in agreement with the simulated results under normal incidence.

A similar method has been employed to design another mechanically controllable active RAS, where the discrete variation in the air spacer thickness has been used for tuning activity (Jeong et al. 2018). The proposed structure employs a 3-D printed frame at the backside of the FR4 layer, and the spacer height has been varied using the frame. By changing the spacer thickness from 1 to 3.5 mm, the absorption frequency is obtained between 8.9 and 8.0 GHz, respectively. The performance of the proposed design has been demonstrated through full-wave simulation and measurement.

Several research articles have also been published on pressure-driven reconfigurable RAS designs. Those geometries have exploited the microfluidic technology, where microchannels have been engraved in an elastomeric substrate. Afterwards, different types of fluids (either conductive or dielectric) have been inserted/extracted through the microchannels, thus exhibiting dynamically modulated characteristics. The injection/extraction of fluids has been carried out either manually or through some mechanical actuators, thus labeling the designs as mechanically active RASs.

A frequency-switchable metamaterial absorber based on the microfluidic technology has been demonstrated in Ling et al. (2015). The proposed design consists of a metallic resonator at the top layer, and microfluidic channels are loaded across the geometry. The design has been printed on a grounded FR4 dielectric. Figure 28a

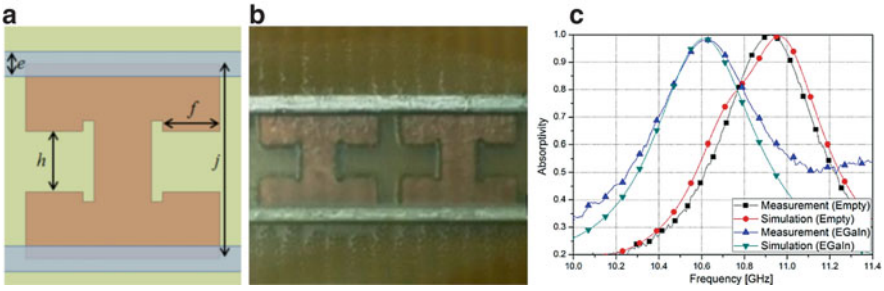


Fig. 28 Frequency-switchable metamaterial absorber using microfluidic technology. (**a**) Unit cell geometry. (**b**) Zoomed view of the liquid metal-filled fabricated prototype. (**c**) Comparison between simulated and measured absorptivities under different conditions. (Reproduced from Ling et al. (2015). Copyright 2015, MDPI)

depicts the unit cell geometry. When there is no liquid across the channels, the structure resonates at 10.96 GHz with 99% absorptivity. In contrast, when a liquid metal EGaIn has been injected into the microchannels, the effective inductance in the top layer is enhanced, thereby shifting the absorption response to 10.82 GHz. The design has been fabricated and displayed in Fig. 28b, whose absorption characteristics are measured in a waveguide environment. Both the simulated and measured responses exhibit the switchable absorption behavior of the proposed geometry, as depicted in Fig. 28c, thereby establishing the proposed concept of a pressure-driven mechanically regulated RAS design.

This microfluidic technology has also been used in realizing switchable broadband absorption behavior (Ghosh and Lim 2018). The concept has previously been explained in Ghosh and Srivastava (2017b), where a switchable layer is made using electronically controllable p-i-n diodes. In this proposed design, the diode layer has been replaced by two orthogonally oriented liquid-metal-encased dielectric substrates, sandwiched between the top resistive pattern and bottom ground plane. With control in the liquid metal flow, the structure can exhibit a wide absorption bandwidth switching between two frequency regimes. Without the presence of any metal flow, the design exhibits a 90% absorption spectrum over the frequency range of 4.02 to 8.20 GHz, whereas the bandwidth shifts to higher side (8.04–12.04 GHz) with the inclusion of a liquid metal in the microchannels. Two orthogonally oriented microchannels have been used in the design to widen the absorption bandwidth as well as make the topology polarization-insensitive.

The proposed metasurface has been fabricated through various steps. Initially, the top resistive layer has been constructed using PCB technique and later mounting the lumped resistors. The microchannels have been engraved in PDMS substrates through laser etching, and joined together through bonding films. Liquid metal EGaIn has been injected/extracted through the channels manually; however, the same can be carried out by using a mechanical actuator. The final prototype has been measured using free space technique and the absorption spectra under different working states have been compared with those of the simulated responses. A good agreement between the results successfully demonstrates the switching characteristics of the proposed RAS design.

There are several other articles reported on pressure-driven switchable/tunable metamaterial RS designs, where the liquid flow has been regulated to exhibit various characteristics (Lim and Lim 2018). Instead of liquid metal, water has been used in a few switchable absorber designs (Xiong and Yang 2020), where by exploiting the dispersive characteristics of water molecules, broadband absorption has been obtained. At the same time, switching activity has been incorporated through injection/extraction method, thereby resulting in pressure-driven wideband switchable metamaterial RAS designs.

In addition to the above reported reconfigurable techniques (electronically control, mechanically control, and magnetically control), various other methods have been developed by researchers to realize active RAS designs. The charge density in the graphene material can be precisely controlled and a wide variety of graphene based reconfigurable absorbers have been realized across microwave and

terahertz frequencies (Balci et al. 2015). Micro-electro-mechanical system (MEMS) based reconfigurable metamaterial absorbers have also been reported, where the resonating structures are actuated by external voltages and the tunability in resonant frequency are achieved by changing the physical pattern of the unit cells (Pitchappa et al. 2014). The voltage dependent birefringence property exhibited by liquid crystals has been explored in metamaterial unit cells, leading the development of tunable absorbers (Shrekenhamer et al. 2013b). Other types of reconfigurable techniques, such as photodoping (Chen et al. 2008), electronic charge injection (Chen et al. 2006), and temperature control (Driscoll et al. 2008) have also been demonstrated in recent years to attain frequency agile and reconfigurable metamaterial RAS designs.

Conclusions

This chapter presents a widespread review of metamaterial FSS-based active RAS designs. An active absorber structure is often preferred over a passive design due to its various advantages, such as multifunctional characteristics, wider usability, etc. Since the last decade, a wide variety of reconfigurable techniques have been developed across different frequency ranges to realize dynamically switchable and tunable absorber designs. Among them, electronically control, mechanically actuation, and magnetically regulation are the most reported techniques for attaining such reconfigurable properties. A wide number of articles have been discussed, along with their geometric configurations, simulated absorption responses, fabricated prototype, and measured results in detail. The biasing circuitries, in particular, for diode-regulated active RAS designs, have also been described in the chapter. Owing to wider availability in electronically controlled absorber articles, the discussion has been divided into different sections based on the EM responses, viz. switchable absorber/reflector, switchable absorber, and tunable absorber. Each of the sections reviews narrow-band and broadband absorber articles and their corresponding analysis has been presented. In case of mechanically actuation and magnetically regulation, several types of EM absorbers have been explored citing their design configurations and working principles. Later, some other types of reconfigurable RAS designs have been reported, indicating the vast domain of metamaterial FSS-based active RAS designs. A comprehensive summary of those FSS-based active RAS designs have been presented in Table 1 for easy understanding of the readers.

This review work will be immensely helpful to understand the basic mechanism of active absorber geometries and their working principles. The existing limitations in the reported designs and the plausible solutions can be devised while going through this chapter. This chapter will also be beneficial for identifying the unexplored regions in the reconfigurable absorber topic. With the advent of full-wave simulation facilities and advanced fabrication technologies, active RAS designs will be further explored in the coming days and this book chapter will be a helpful guide for carrying out the research activities in the domain of active absorbers.

Table 1 A comprehensive review of electronically modulated metamaterial FSS-based active RAS designs

Reconfigurable FSSs	Type of geometries	Type of absorption bands	Switching/ tuning range	Polarization-insensitivity	Biasing configuration
Yoo and Lim (2014b)	Switchable absorber/ reflector	Single-band absorption (ON state)	10.79 GHz (absorption mode)	No	Use of via, parallel connection
Zhu et al. (2010a)	Switchable absorber/ reflector	Single-band absorption (ON state)	3.34 GHz (absorption mode)	Yes	Use of via, parallel connection
Ghosh and Srivastava (2016a)	Switchable absorber/ reflector	Single-band absorption (OFF state)	3.71 GHz (absorption mode)	Yes	Metal lines with inductor, parallel connection
Kong et al. (2015)	Switchable absorber/ reflector	Broadband absorption (OFF state)	4.5–7.7 GHz (absorption mode)	No	Metal lines with inductor, series connection
Ghosh and Srivastava (2016a)	Switchable absorber/ reflector	Broadband absorption (OFF state)	3.48–8.04 GHz (absorption mode)	Yes	Metal lines with inductor, parallel connection
Zhu et al. (2010b)	Switchable absorber	Single-band absorption (ON/OFF states)	2.9 GHz (ON state)/ 2.6 GHz (OFF state)	No	Diodes between unit cells, series connection
Xu and Sonkusale (2013)	Switchable absorber	Single-band absorption (ON/OFF states)	4.78 GHz (ON state)/ 4.10 GHz (OFF state)	No	Diodes between unit cells, series connection
Ghosh and Srivastava (2016c)	Switchable absorber	Single-band absorption (ON/OFF states)	3.41 GHz (ON state)/ 5.35 GHz (OFF state)	Yes	Metal lines with inductor, series connection
Ghosh and Srivastava (2018)	Switchable absorber	Single-band absorption (ON/OFF states)	2.75 GHz (ON state)/ 4.56 GHz (OFF state)	Yes	Use of via, parallel connection
Ghosh and Srivastava (2017b)	Switchable absorber	Broadband absorption	3.68–11.80 GHz (ON state)/	Yes	Metal lines with inductor,

(continued)

Table 1 (continued)

Reconfigurable FSSs	Type of geometries	Type of absorption bands	Switching/ tuning range	Polarization-insensitivity	Biasing configuration
		(ON/OFF states)	2.08–4.94 GHz (OFF state)		parallel connection
Yuan et al. (2015)	Tunable absorber	Single-band absorption (All states)	8.48–8.92 GHz	No	Diodes between resonators, parallel connection
Ma et al. (2016)	Tunable absorber	Single-band absorption (All states)	1.63–2.67 GHz	No	Diodes between resonators, parallel connection
Ghosh and Srivastava (2017c)	Tunable absorber	Single-band absorption (All states)	1.08–3.16 GHz	Yes	Use of via, embedded biasing
Yuan et al. (2014)	Tunable absorber	Dual-band absorption (All states)	2.69–3.37 GHz (tunable), 4 GHz (constant)	No	Diodes between resonators, parallel connection
Ghosh and Srivastava (2017d)	Tunable absorber	Dual-band absorption (All states)	0.98–3.76 GHz (absorber I), 0.88–3.33 GHz (absorber II)	Yes	Metal lines with inductor and vias, parallel connection

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Metamaterial-FSS Based Radome Using SIW Technology

5

V. Krushna Kanth and S. Raghavan

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Abstract

In this chapter, metamaterial (MTM) inspired frequency selective surfaces or frequency selective screens (FSSs) are demonstrated using a substrate integrated waveguide cavity (SIWC) technology. SIW cavity is formed by the series of thin metallic wires around the FSS unit cell connecting the top and bottom metal

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layers of the dielectric substrate. The material characterization of MTM-FSS based on SIW is carried out and proved the metamaterial behavior of the structure. A combined work on MTM inspired FSS in conjunction with SIW is presented for the radome application. The SIW cavity is applied to MTM-FSS to improve the unit cell performance characteristics. The SIW cavity appends the superiority of the rectangular waveguide (RGW) to the FSS structure such as sharp cutoff frequency, high – Q factor, and high-power handling capabilities. The technique to reduce and improve the bandwidth of FSS using SIW is discussed. Further, a planar radome wall has been developed using the proposed MTM-FSS based on the SIW cavity. The radome wall characterization and testing procedures to qualify for the airborne application is presented in brief. The chapter may be found interesting for the Rf and Microwave community specific to the EMI protection and radome engineers.

Keywords

Active electronically scanned array · Frequency selective surfaces · Matamaterials · Substrate integrated waveguide · Radome

Introduction

As a convention, frequency selective surfaces or frequency selective screens (FSS) are planar periodic arrays that act as a selective filter for electromagnetic (EM) waves in the free space hence, popularly known as spatial filters. Based on the requirement, these structures are composed of a dielectric substrate with thin metallic patches or slots in a specific pattern to exhibit low-pass, high-pass, bandstop, or bandpass responses. The specific design with lossy dielectrics or lossy metallic patterns made them act as absorbers. From the past few decades, FSS has been extensively studied for its potential applications in the EMI shielding of RF and microwave and millimeter-wave applications as spatial filters, radomes [Wu (1995), Vardaxoglou (1997), Munk (2000) and Wang (2015)]. The application of these structures includes ground plane to the antenna [Narayan et al. (2016)], radomes [Nair R. U., and Jha R. M. (2014)], etc. for selected frequency bands in the wide frequency spectrum. The combination of different types and layers of FSS element structures can be used for high-performance multi-band and wideband structures [Kesavan et al. (2016)]. Indeed, these planar periodic arrays play a crucial role in absorbing EM radiation impinging on its surface and reducing the radar cross-section of the structure significantly [Li et al. (2016), Sharma et al. (2020) and Kazantsev et al. (2020)].

Conventional FSS resonant structures are arranged in two dimensions (2-D) on a plane, and have been widely investigated and used for various applications for decades. These applications consist of electromagnetic (EM) spatial filters in free space, EM absorbers, high-impedance surfaces (HIS), artificial electromagnetic bandgap structures (EBG), etc. These structures are usually arranged in infinite planar periodic arrays in two dimensions. Thereof, the frequency behavior of the

FSS is entirely determined by the structural geometry in one period called a unit cell. The studies have mostly focused on the band-stop and band-pass characteristics produced by these surfaces. The periodic structures with bandpass characteristics have specular application in stealth. In the past decade, bandpass FSS attained the interest for modeling the stealth applications and developing FSS unit cells with dimensions much smaller than a wavelength ($<< \lambda$) is of high priority. The electrically small dimensional FSS is useful for exhibiting effective filtering response, flexible, and conformal applications. In traditional designs, the frequency-selective properties result from mutual interactions of the adjacent unit cells. These unit cells are modeled and behaved as an equivalent lumped resistor (R), inductive (L), and capacitive (C) elements and are properly arranged in the equivalent circuit (EC) so they couple to the incoming magnetic and electric fields of an impinging EM wave.

However, the applications of these conventional 2-D planar periodic arrays are limited and suffered from the limited bandwidth (BW), an unstable behavior at higher oblique incidence, frequencies, and polarization angles. Also, these 2-D periodic arrays are not thermally stable because of the printed thin metallic layers. These thin metallic arrays may burn/deform and causes breakdown at high microwave powers and lose their basic performance characteristics [Munk et al. (1971), Liu and Behdad (2013)]. The dielectric or metallic breakdown in the periodic structure is occurred due to high temperatures (K), induced high voltage (V), and currents (I) due to high microwave power (HMP) along the conducting path in the structure. These are the un-avoiding effects in a conventional periodic FSS. Increasing the gap between elements and increasing the thickness of metallic elements reduces the chances of breakdown conditions. However, the real challenge occurs when periodic FSS is used for airborne applications. Also, increasing the dielectric substrate and metallic thickness is not a recommendable solution for the airborne application, as it increases the overall payload of the system. Recently, the substrate integrated waveguide (SIW) technique is developed to design microwave and millimeter wave components, circuits, and subsystems. The metallic SIW cavity isolates the sensitive electronic components in the printed circuit board (PCB). The SIW structure is an equivalent model of the rectangular waveguide, and it appends the superiority of conventional metallic waveguides, such as high Q-factor, high selectivity, cutoff frequency characteristic, the high power capacity of the FSS with better EM performance characteristics [Zhang et al. (2005), Deslandes and Wu (2003), Luo et al. (2005), Qi et al. (2008) and Luo (2007a)].

SIW is a planar transmission line that has the better transmission behavior and was popularized in the past two decades by researchers. In high-frequency applications, microstrip devices are lossy and not efficient due to their poor performance and high transmission losses. In addition, wavelengths at high frequencies are small. Hence, microstrip device manufacturing requires very tight tolerances and fabrication accuracies. Basically, at high frequencies, waveguide devices are commonly preferred for higher efficiency. However, their manufacturing process is costlier and more difficult. Therefore, a new concept SIW is emerged as a trade-off between conventional microstrip and rectangular waveguide (RWG) and dielectric-filled waveguide (DFWG) technologies. Dielectric filled waveguide is converted to SIW

structures with the help of vias by replacing the sidewalls of the waveguide [Bozzi et al. (2011), Krushna Kanth and Raghavan (2019a), Feng Xu and Ke Wu (2005)]. SIW technology is applied to the development of high-efficient periodic FSS structures by improving their EM response (Luo et al. 2005, 2006, 2007a).

When SIW is used for the FSS, it results in multiple coupling paths, which induce a transmission zero near the passband. Consequently, enhancing the upper or lower stopband of the frequency responses improve the bandwidth and roll-off performance. With the further enhancement of the frequency selectivity, two periodic FSS shapes individually with one transmission zero have been combined to obtain multiple transmission zeros (Luo et al. 2008). Moreover, this combined periodic FSS has a design complexity that is an arduous task for the design engineer and real-time usage. A cavity-based technique to improve and reduce the bandwidth of the FSS unit cell is proposed without altering the design dimensions of the unit cell [Krushna Kanth and Raghavan (2019a)].

The latest development in the defense systems opted for the active electronically scanned array (AESA) for improved EM performance and better scanning and tracking of the targets. It requires a superior-performance radome. This includes a very large operational bandwidth (~1 GHz), flat-top response, minimum transmission loss (<1 dB), and abrupt roll-off performance (at both edges of passband) characteristics. The conventional monolithic, sandwich/hybrid, and FSS radomes may not offer such superior performance in terms of selectivity and out-of-band rejection. On contrary, these radomes degrade their functional performance at millimeter wave frequencies. The aforementioned high performance can be achieved with the usage of MTM, MTM-based FSS, Multi-layered MTM/FSS radome walls as they show stable EM performance over a wide incident angle in the millimeter wave frequency region. Given the abovementioned radome design limitations, in this chapter, periodic FSS based on SIW cavity technology has been investigated for airborne radome application. The spectacular advantages of FSS based on SIW such as sharp roll-off, flat passband, highly stable, high Q-factor, high-power handling, and ease to fabricate using conventional PCB process have been considered to design a novel Multilayer radome to improve the electrical performance along with the mechanical and structural requirements of modern facets of advanced radar system requirements [Krushna Kanth and Raghavan (2019a)]. The wide variety of applications such as polarizers [Mollaei M. S. M. (2017), Akbari et al. (2017), Krushna Kanth and Raghavan (2018a)], dual-band FSS [Krushna Kanth and Raghavan (2020)] can be realized using SIW technology.

In this chapter, two applications of FSS based on SIW have been discussed: (1) a technique to control the bandwidth of FSS using SIW cavity technology, and (2) design methodology of Metamaterial-FSS (MTM-FSS) based radome using SIW technology. In the first case, the SIW cavity is applied to an FSS to reduce the BW and improve the BW. In the second case, an MTM-FSS unit cell is designed and its performance was analyzed. Further, an SIW cavity is modeled to incorporate the proposed MTM-FSS unit cell. Further, the unit cell is extended for the radome wall application. The detailed design and analysis of the proposed concepts are explained in the subsequent sections of this chapter. The details of the fabrication method and experimental verification are described in brief.

FSS Bandwidth Controlling Using SIW Technology

A technique to control the bandwidth of FSS using SIW cavities is discussed in the subsequent sections. Two cavity techniques were explained to reduce and improve the bandwidth and selectivity of FSS. The detailed design is explained in the following sections.

Cross Slot FSS Element

In this work, a simple dipole cross slot element is considered to design and prove the influence on FSS bandwidth by appending the SIW cavity model to the unit cell. Initially, the dipole cross-slot FSS is designed to operate at the 10.0 GHz in the center of the X-band frequency region by considering its potential applications in radar, missile guidance, tracking, and scanning of enemy targets. The maximum incident EM wave is coupled when the dimension of the structure is equal to half of the operating wavelength. The resonance frequency of the geometry depends on the total length of the slot. It is estimated as the half of operating wavelength (Luo et al. 2005), the approximated expression for $f_{r,s}$ is:

$$f_{r,s} = \frac{0.5c_o}{d\sqrt{\varepsilon_{eff}}} \tag{1}$$

In Eq. (1), d is the length of the slot element and ε_{eff} $(=(\varepsilon_r + 1)/2)$ is the effective dielectric permittivity of the substrate material. The dimensions of the cross slot are the length of cross slot (d) and the width of cross slot (w) are 12.5 mm and 2.0 mm, respectively. The periodicity (P) of the unit element is half ($\lambda/2$) of the operating wavelength of the unit cell, which is calculated as 15.0 mm ($\lambda/2$). The slot is etched from a copper layer having a thickness of 35 μm and the copper layer is backed by a dielectric substrate, a commercially available Rogers RT Duroid 5880 substrate with dielectric permittivity, $\varepsilon_r = 2.2$, loss tangent, $tan\delta = 0.0009$). The thickness of the substrate is $t = 1.57$ mm (commercially available) used for physical support. In addition, the substrate material has a very low loss tangent which is more effective for aerospace applications with very low transmission loss. The design of the slot element and its frequency response in the X-band region is presented in Fig. 1a–c. Further, to understand the influence of design parameters, the parametric study has been conducted w.r.to the length and its transmission/reflection coefficients are evaluated and shown in Fig. 2a, b. The obtained response through parametric study helps in better insight observation on the dependence of FSS slot dimensions on frequency response. The plotted response reveals a $-$ 1 dB (>90%) relative transmission bandwidth of 11.18% from 9.36 GHz to 10.468 GHz. The structure shows a maximum insertion loss (IL) of 0.04 dB and a maximum return loss (RL) of 24.1 dB at the center frequency of the X-band.

Furthermore, in the upcoming studies, an SIW cavity has been incorporated for the FSS design. The effect of SIW cavity on FSS performance is studied and

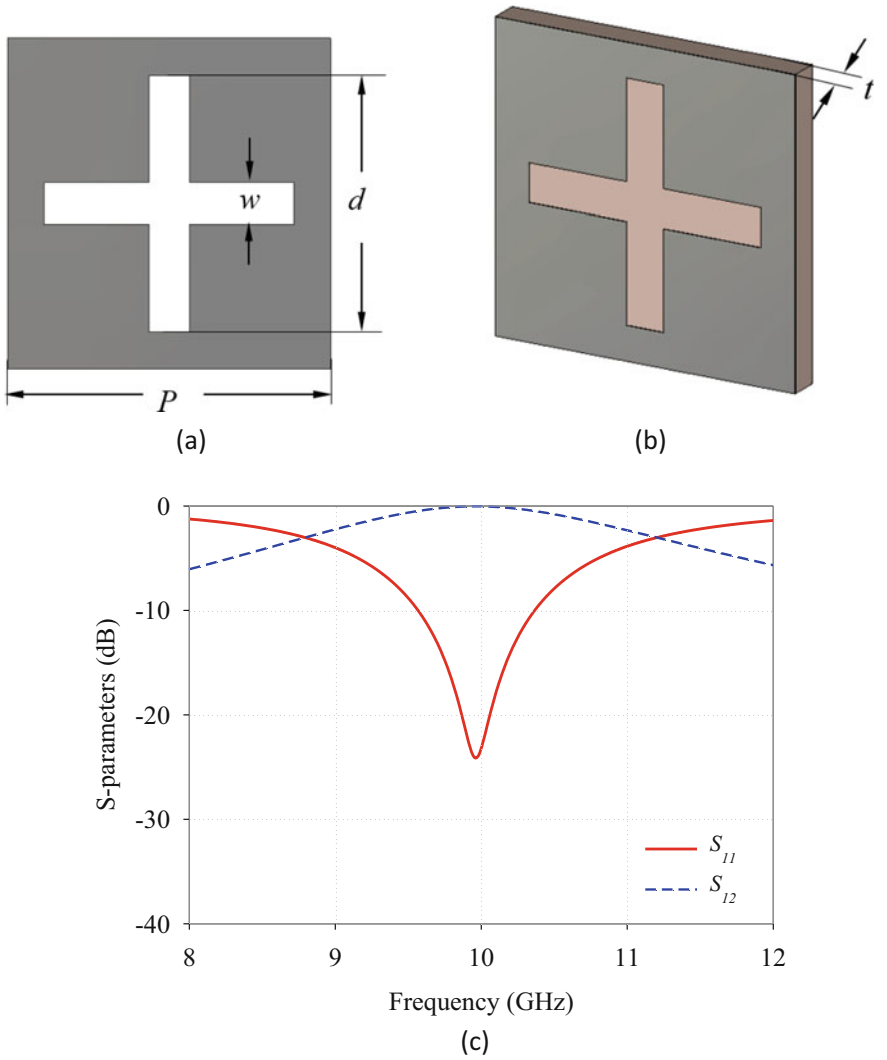

Fig. 1 Dipole cross slot FSS element: (**a**) unit cell model, (**b**) perspetive view, (**c**) frequency response in the X-band

conclusions are made from the observations. Two cavity models are reported to reduce and improve the bandwidth characteristics of the FSS. Commercially available full-wave EM simulator CST Microwave studio is used for all the unit cell simulations. The periodic boundary conditions on the unit cell have been used on four sides of the FSS with Floquet port plane wave excitation on top and bottom surfaces of the unit cell to impinge the EM wave on the FSS geometry.

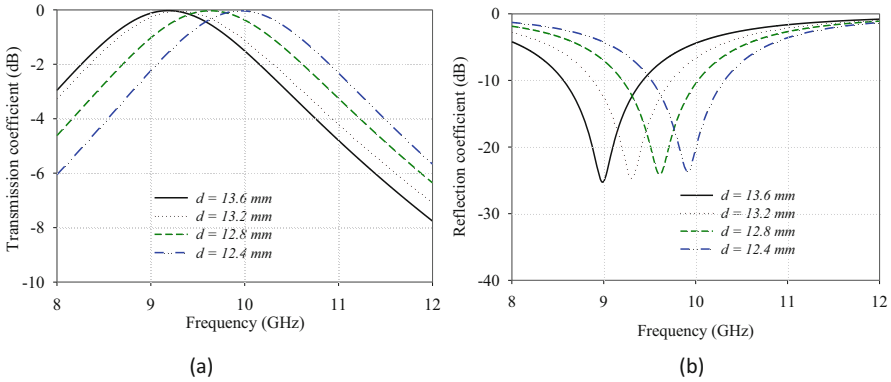

Fig. 2 Influence of cross slot length on the frequency response of single layer FSS: (**a**) Transmission coefficient, (**b**) Reflection coefficient

Reducing the Bandwidth of FSS

In convention, it is known that reducing the BW intern results in high selectivity, which found its specific applications in high precession scanning applications. A closed SIW cavity model is presented in this section to reduce the bandwidth of FSS. For designing the closed SIW cavity, FSS slots (shown in Fig. 1) with equal dimensions are etched on either side of the dielectric substrate material. Further, a series of metallic vias are aligned around the cross slot FSS element to form a closed SIW cavity as shown in Fig. 3a, b. The closed SIW cavity is specified with the name Cavity-I to avoid ambivalence. The modeled closed cavity will act as a virtually thick screen and reduce the thickness. The necessary and sufficient conditions were used to design and align the metallic vias along the slot element to avoid the EM leakage [Bozzi M., Georgiadis A., and Wu K. (2011), Wang H. B., and Cheng Y. J. (2016), Luo G. Q., Hong W., Hao Z.-C., et al. (2005)].

$$d_r < \frac{\lambda_0}{5} \tag{2}$$

$$d_p < 0.5\lambda_0\sqrt{\varepsilon_r} \text{ and } d_p < 4d_r \tag{3}$$

The values of d_v, d_p, and d_q are 0.6 mm, 1.5 mm, and 1.2 mm respectively. d_s is the distance between the two vias at the edges of the arm. The numerical value of the d_s is 3 times that of d_q which is 4.5 mm. In the above equation, λ_0 represents the free space wavelength at 10 GHz.

The presented SIW cavity introduces the parallel inductance in the equivalent circuit model (ECM) representation of the conventional single-layer cross slot FSS. This increases the equivalent inductance of the overall structure. The representation of analytical expressions is detailed as follows:

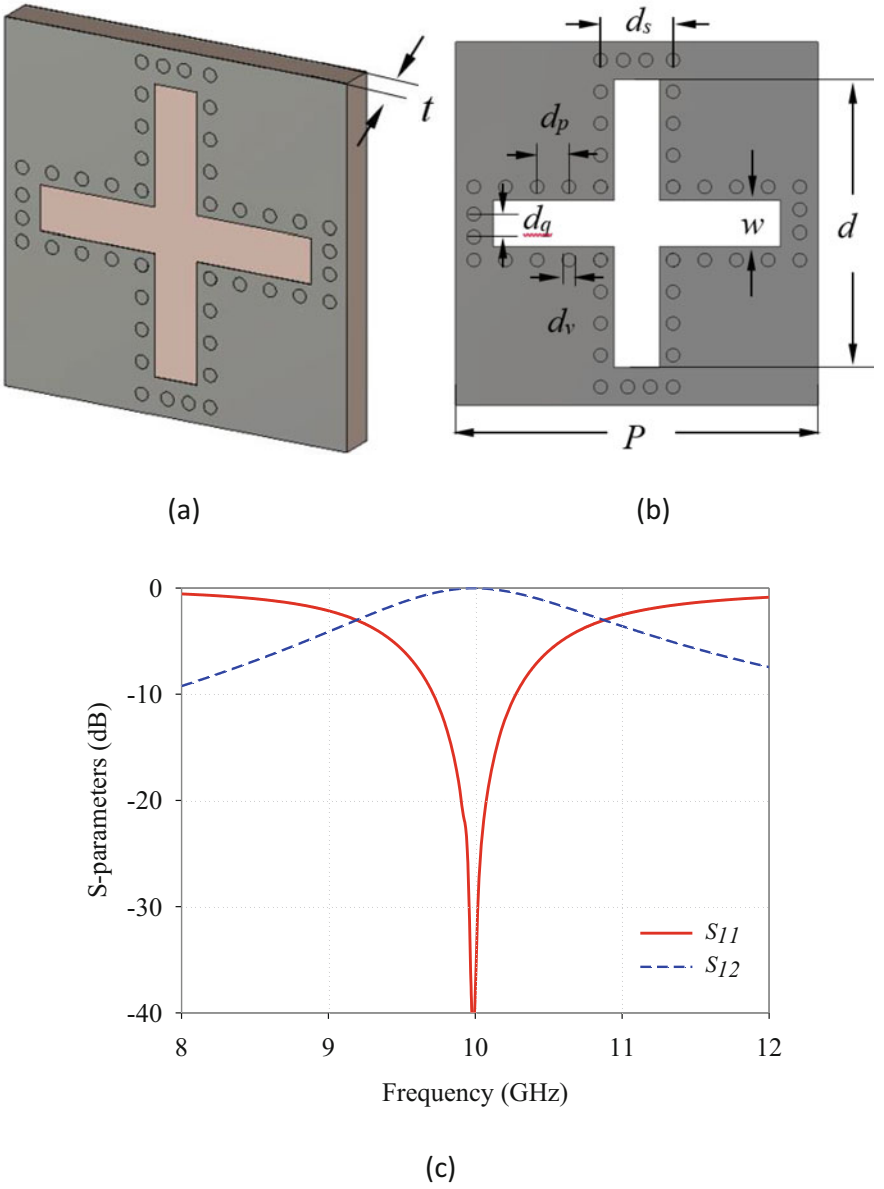

Fig. 3 Cross slot FSS element design with the help of SIW cavity-I: (**a**) unit cell structure, (**b**) perspetive view of the unit cell, (**c**) corresponding frequency response

$$Q = 2\pi f_r L/R \tag{4}$$

$$f_r = 1/2\pi\sqrt{LC} \tag{5}$$

where f_r, L, R, and C are the resonance frequency (f_r), inductance (L), resistance (R), and capacitance (C) of the model, respectively. Equations (4) and (5) shows the direct dependence of inductance (L), resistance (R), and capacitance (C) on quality factor (Q) and resonance (f_r). As the L increases, Q increases, and f_r decreases. This results in the reduction of bandwidth by improving the cutoff frequency with sharp selectivity of the unit cell model [8].

For clear insight observation, the ECM of the proposed SIW cavity-based FSS configuration has been presented in Fig. 4 [Suresh Kumar T. R., Venkatesh C., Salil P., and Subbarao B. (2015)]. As the proposed FSS comprises of double metallic layers with a metallic cavity, the ECM contains two LC tank circuits with similar lumped elements which are cascaded back-to-back. The impedance offered by the

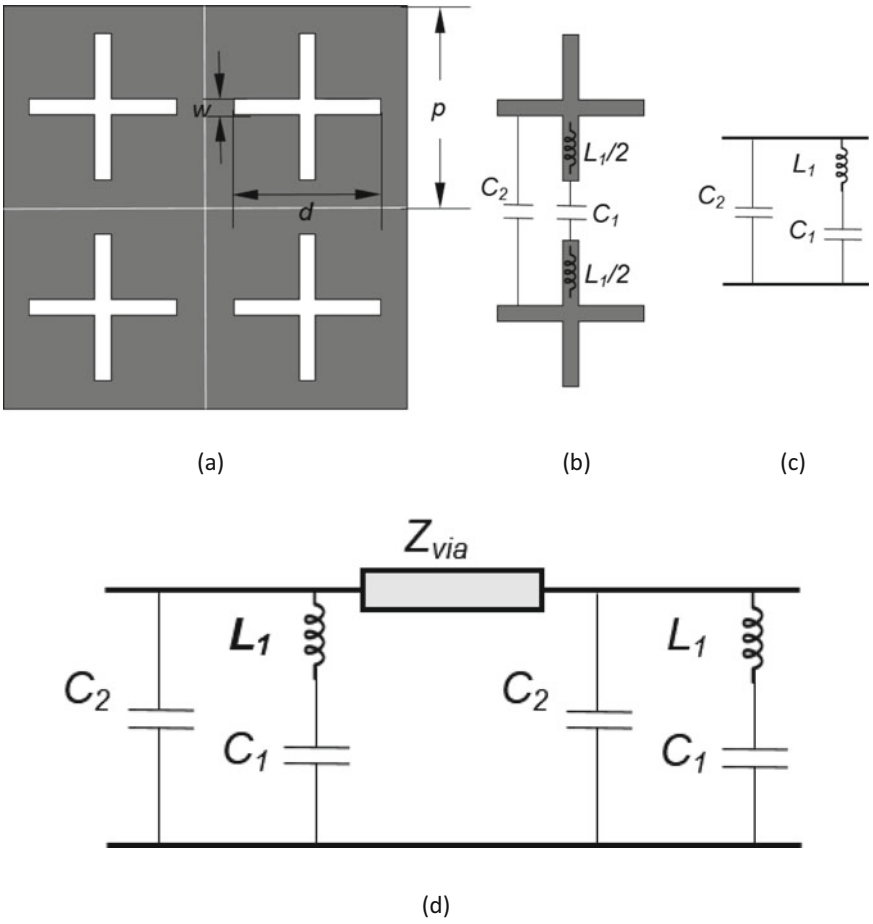

Fig. 4 Circuit modeling: (**a**) 2 × 2 array of the single layer cross slot, (**b**) modeling of lumped elements on the cross slot FSS, (**c**) ECM of single layer FSS, (**d**) ECM of the FSS based on SIW cavity

metallic vias along with the substrate material (Z_{via}) is connected between the two LC tank circuits as shown in the equivalent circuit model. The vias in the structure can be modeled as an inductor (L) and mutual capacitance (C) and can be found in the dielectric substrate material between the two parallel periodic FSS layers. It is indeed worth mentioning that the ECM shown in Fig. 4 is valid for the patch type dipole cross elements. From the knowledge of Babinet's principle, it is known that the reflection coefficient of a patch array is equivalent to the transmission coefficient of its complementary (slot) array [Krushna Kanth V., and Raghavan S. (2018a)]. Therefore, the scattering parameters of the proposed model dipole cross slot model can be estimated from the given circuit model.

The frequency response of the SIW Cavity – 1 based cross slot FSS is shown in Fig. 3c. It is observed that the structure exhibits a transmission bandwidth better than −1 dB in the frequency range from 9.5 GHz to 10.5 GHz with a relative BW of 10.0%. Further, the structure exhibits a maximum insertion loss of 0.02 dB and a return loss of −51.0 dB at operating frequency (10.0 GHz). Nevertheless, it is observed the frequency selectivity at the center frequency is significantly improved compared with the normal single layer cross slot. This can be seen from plotting the return loss (S_{11}) response. The influence of cross slot length on the resonance frequency (on both transmission and reflection coefficients) in Cavity-I is presented in Fig. 5a, b. As it is obvious, the increase in slot dimension leads to reducing the resonance frequency leftwards. Furthermore, the transmission and reflection response of Cavity-I is studied for oblique incidence angle up to 60°, and it is observed that the structure exhibits stable resonance. The corresponding characteristics are shown in Fig. 6a, b, respectively. The discontinuity in the frequency response at 11 GHz for higher incidence angles (>45°) is due to the generation of higher-order modes in the cavity structure. However, there is no significant deviation found in the resonance frequency due to oblique angle incident EM wave on the surface.

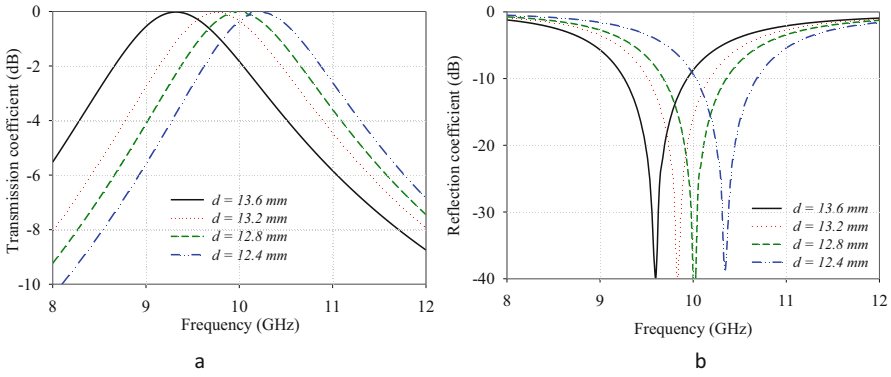

Fig. 5 Effect on frequency response w.r.to length of the dipole cross slot FSS based on SIW cavity-I. (**a**) Transmission coefficient, (**b**) Reflection coefficient

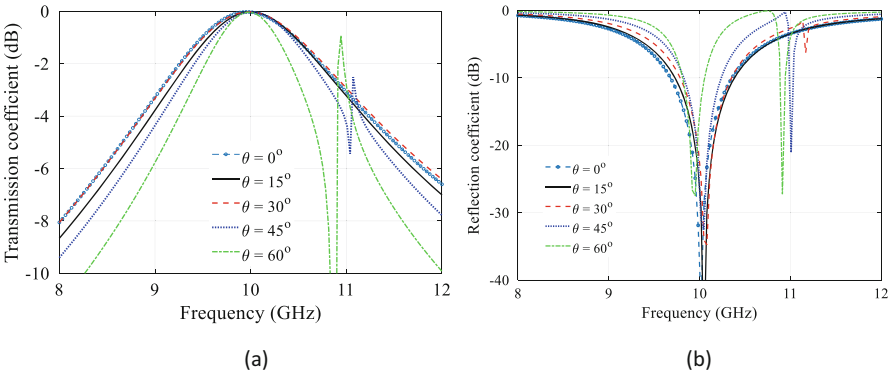

Fig. 6 Frequency response w.r.to oblique angles on FSS based WIW cavity-I: (**a**) Transmission coefficient, (**b**) Reflection coefficient

Increasing the FSS Bandwidth

To increase the BW of FSS, a square metallic cavity has been. The vias form a closed metallic cavity around the periodic FSS element as shown in Fig. 8. In this case, the cross-slot FSS element generates a single resonant passband, and second passband is generated by the effect of the SIW cavity. The cavity size is frequency-dependent ($f_{r,C}$) and operating mode dependent TM_{mn0}, which is computed from [Momeni Hasan Abadi S. M. A., and Behdad N. (2015)]:

$$f_{r,C,mn0} = \frac{0.5c_0}{\sqrt{\varepsilon_r}} \left[\left(\frac{m}{W_{eff}} \right) + \left(\frac{n}{L_{eff}} \right) \right]^{1/2} \tag{6}$$

In the above equation, W_{eff}, L_{eff} are the effective width and length of the SIW cavity. ε_r and c_0 are the dielectric permittivities of the substrate and free space velocity, respectively. Furthermore, the proposed model is a square periodic structure with unit cell element periodicity P. Hence, $W_{eff} = L_{eff} = P_{eff}$ and can be computed by [Deslandes D., and Wu K. (2006)]:

$$P_{eff} = P - 1.08 \frac{d_v^2}{d_p} + 0.1 \frac{d^2}{P} \tag{7}$$

The size of the SIW cavity of first-order mode (TM_{10}) is

$$f_{r,C,10} = \frac{0.5c_0}{\sqrt{\varepsilon_r}} \left(\frac{1}{P_{eff}} \right)^{1/2} \tag{8}$$

In the above equation, P is the periodicity of the unit cell model, and d_v and d_p are the cylindrical via diameter and pitch distance between adjacent vias, respectively.

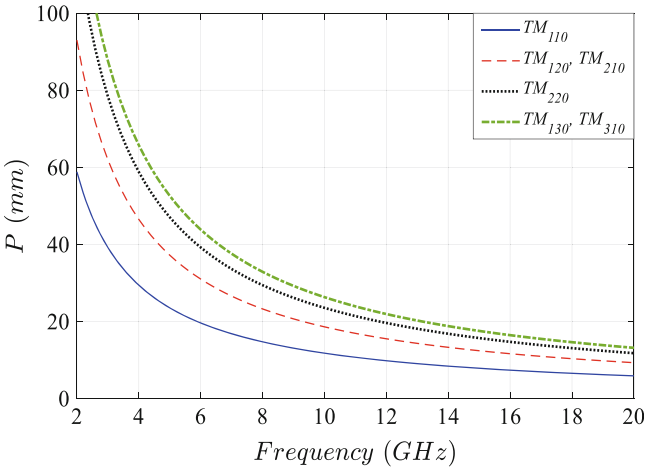

Fig. 7 Width of SIW cavity at different modes

The generation of various operating modes inside the SIW cavity model is presented in Fig. 7 as a function of the periodicity of the proposed FSS-based SIW cavity. From Fig. 7, the size of the cavity for TM_{110} is 15.0 mm at d_v, d_p are 0.6 mm and 1.5 mm, respectively, which are calculated based on Eqs. (2) and (3). The dimension of the slot is remaining unchanged as given in section II to see the effectiveness of the SIW cavity on the BW response. The Cavity-II with cross slot FSS is presented in Fig. 8a, b. In this case, based on the size, the SIW cavity will resonate at an $f_{r,C}$, and the slot will resonate at $f_{r,S}$ based on the design dimensions of the slot. The broadband frequency response is achieved by making both resonance frequencies resonate close to each other. As an effect, flat–top resonance is achieved. In other words, the EM wave perturbations from the incoming wave caused in the SIW cavity leads to a wider bandwidth. Figure 8c shows the frequency response of the Cavity-II model combined with the FSS slot. The obtained sharp roll-off at the higher cutoff frequency and flat-top response over the band of the frequency region is evidently presented in the figure. The structure shows a $-$ 1 dB transmission bandwidth from 9.24 GHz to 10.72 GHz with a relative bandwidth of 14.82% in the X-band region. The structure has a maximum IL of 0.02 dB and a maximum RL of -43.0 dB at 10.0 GHz. Furthermore, the structure exhibits return loss below -15 dB from 9.64 GHz to 10.65 GHz. Further, understanding the design parameter's influence, the effect of slot length in the SIW cavity is studied using parametric analysis. The effective results of transmission and reflection response are shown in Fig. 9. The slot length of the FSS is directly proportional to the bandwidth. It is due to the EM perturbations inside the SIW cavity. With a smaller slot length, the resonance is occurred at a lower frequency and shifts to a higher frequency with reducing slot length. Consequently, less coupling between the resonance frequency of the SIW cavity and the resonance frequency is offered by the slot. Accordingly, the bandwidth of the structure is reduced to a significant value. The higher slot width results

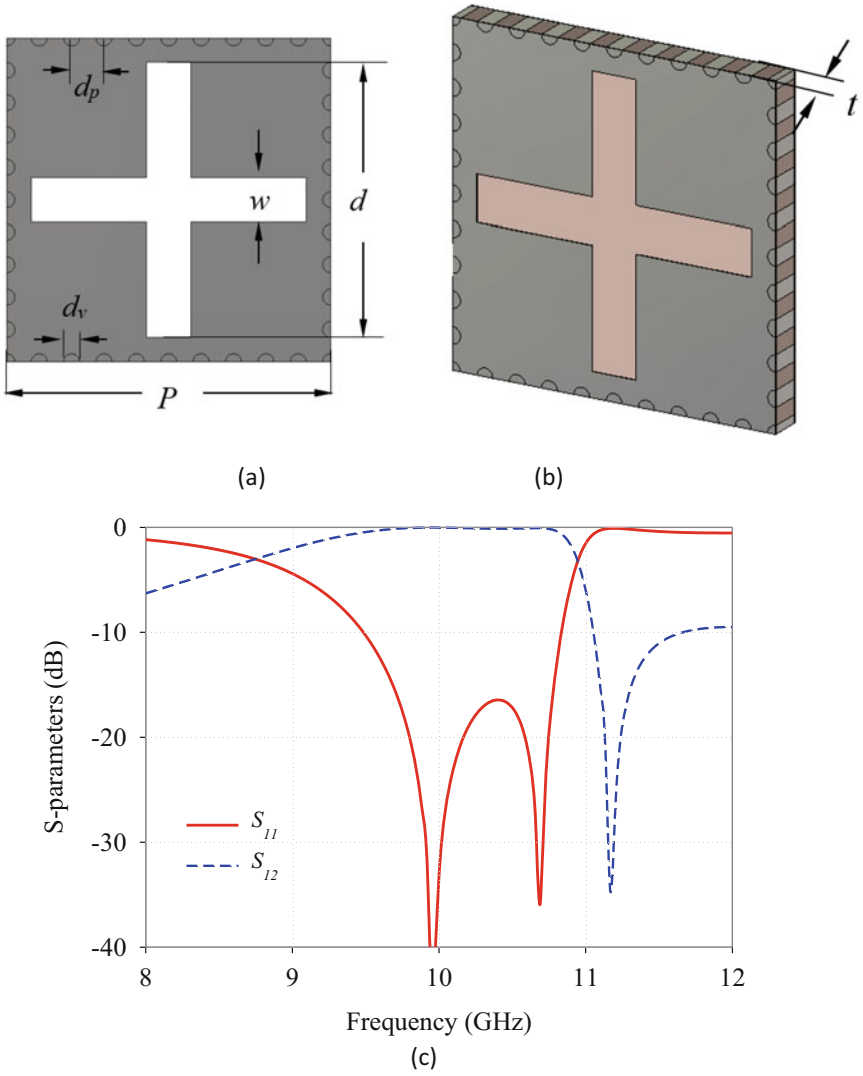

Fig. 8 SIW based cross slot FSS element Cavity-II. (**a**) unit cell model, (**b**) perspetive view, (**c**) scattering parameters

in shifting the resonance frequency to the lower band and increases its coupling with the SIW cavity. This leads to higher bandwidth. Also, the flat-top response in the transmission response is achieved by increasing the slot length. Next, the stability of the SIW Cavity-II model is investigated at a higher oblique incident angle of EM wave and the results are presented in Fig. 10. The design exhibits a very stable EM performance till 45° EM wave impingement. Whereas, due to the occurrence of higher-order modes at higher incident angles, the response starts to deviate beyond >45° in the Cavity-II.

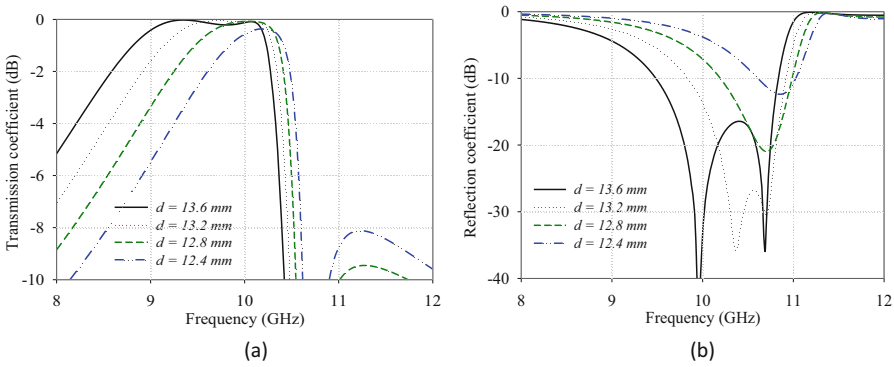

Fig. 9 Frequency response w.r.to length of the slot of periodic FSS with SIW Cavity-II. (**a**) Transmission coefficient, (**b**) reflection coefficient

The presented parametric analysis helps in understanding the frequency behaviors of the unit cell at different dimensional values. The parametric analysis gives the better understanding on the sensitivity of the design parameters on frequency response. This knowledge of sensitivity analysis helps in designing the FSS for the desired frequency of application in the designated frequency band of application.

To show the efficacy of the proposed concept, EM performance characteristics of the cross slot is studied in three different forms, viz. single-layer form, SIW Cavity-I and SIW cavity-II model. From the study, it is observed that the performance characteristics of the structure are improved by using the SIW cavity. Also, the previous studies on SIW-based cavity models improve the power handling capabilities of FSS Belenguer A., Esteban H., and Boria V. E. (2014)]. The transmission and reflection response FSS slot obtained from the analysis of three cases in planar form has been compared in Fig. 11a, b, respectively.

Design of Metamaterial-FSS Unit Cell

The center connected FSS elements are most popularly used for the isolated resonance and reduced bandwidths. The length of cross is for the resonance frequency will be ~ $\lambda_r/2$ where λ_r relative wavelength with the effect of the dielectric substrate. A center connected cross-type metamaterial inspired FSS unit cell has been designed to operate at the center frequency of x-band 10.0 GHz. The design evolution of MTM inspired FSS unit cell is illustrated in Fig. 12. The response of the unit cell is shown at each stage of the evolution process. A full-wave simulation with help of commercially available CST Microwave Studio (MWS) is used to estimate the frequency response of the proposed MTM-FSS unit cell elements.

The design parameters of the cross slot at operating frequency 10.0 GHz is as follows: the size of the FSS unit cell, p is 15.0 mm which corresponds to $\lambda_r/2$. The

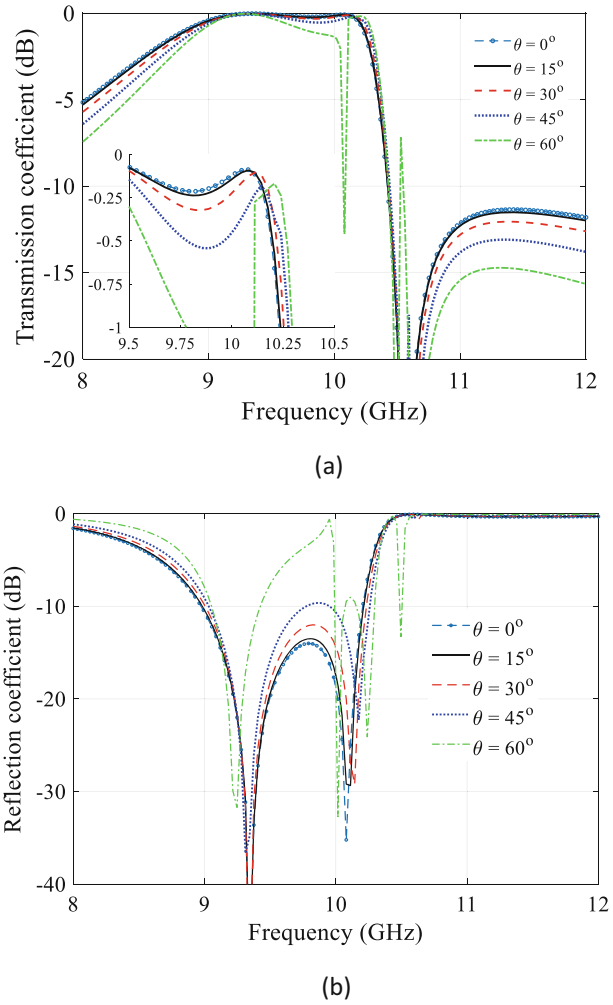

Fig. 10 The frequency with respect to the oblique incidence angle of periodic FSS based on SIW cavity-II. (**a**) Transmission coefficient, (**b**) reflection coefficient

length of the cross slot, L is 12.0 mm and the slot width of the cross, S is 1.2 mm in Fig. 12a. A dielectric substrate Rogers RT 5880 (lossy) with dielectric constant, $\varepsilon_r = 2.2$ and loss tangent, $\tan\delta = 0.0009$ with substrate thickness, $t = 1.57$ mm has been used for all the design cases of the MTM-FSS and further designs based on SIW. A thin pure copper with a thickness of 35 μm with a conductivity of 5.96×10^7 S/m is used as a top and bottom metallic layer to etch the FSS slots. The resonance frequency of the slot at which the maximum incident EM energy coupled with the structure depends on the total length of the cross dipole slot and it can be found at half of the wavelength [Munk (2000)], the approximated expression for f_r is:

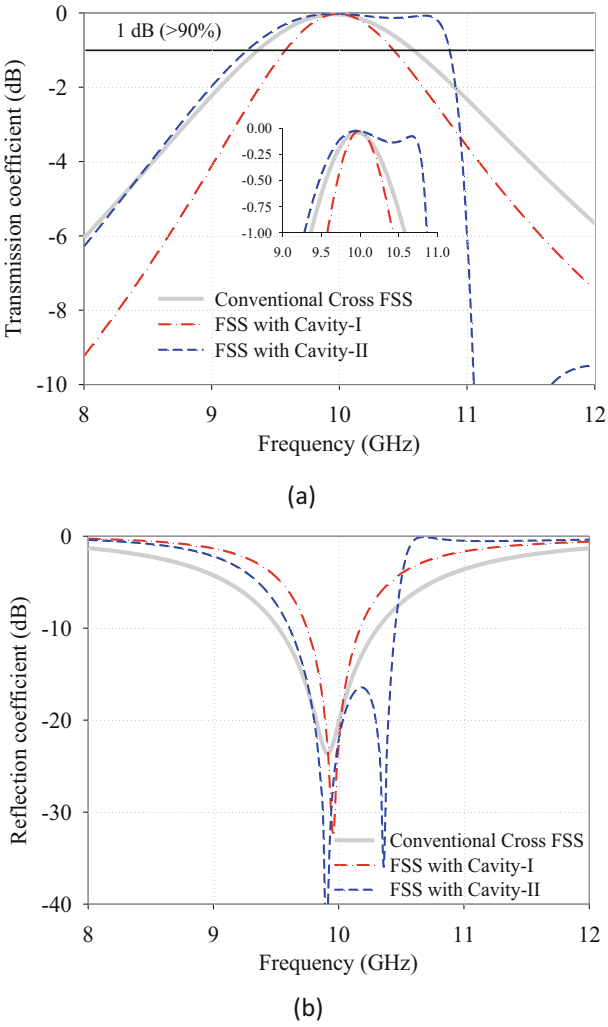

Fig. 11 Frequency response comparison of FSS topologies. (**a**) Transmission coefficient, (**b**) reflection coefficient

$$f_r = \frac{c_0}{2L\sqrt{\varepsilon_r}} \tag{9}$$

where c_0 is the velocity of light in free space, L is the length of the cross element, and ε_r is the dielectric constant of the substrate material used.

The resonance frequency of the modified cross element is shown in Fig. 1b.

As shown in Fig. 12c, structures are showing a single resonance frequency at 10.0 GHz with a bandwidth of 680.0 MHz below −10 dB from 9.62 GHz to 10.3 GHz. The transmission bandwidth of the structure better than −1 dB is

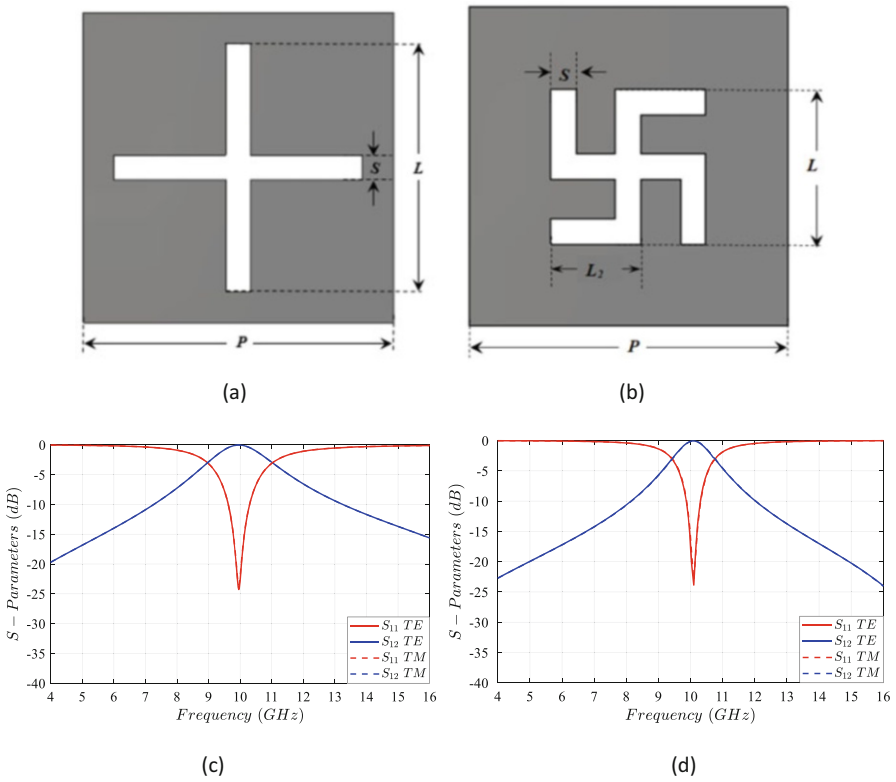

Fig. 12 Design of FSS unit cell: (**a**) cross slot, (**b**) swastika slot, (**c**) frequency response of cross slot, (**d**) frequency response of swastika slot

1030 MHz from 9.45 GHz to 10.48 GHz with insertion loss (IL) 0.045 dB at resonance frequency 10.0 GHz.

Next, the edges of the cross slot have been extended orthogonally with a length of $L_1/2$ to miniaturize the slot dimensions. The length of the cross slot in Fig. 12b, L is 8.4 mm, L_1 and L2 are L/2 and the width of the slot, S is 1.2 mm. The periodicity of the unit cell is kept remain unchanged.

The resonance frequency of the proposed swastika-shaped cross slot can be given as:

$$f_r = \frac{c_0}{2(L_1 + L_2)\sqrt{\varepsilon_r}} \tag{10}$$

From the frequency response shown in Fig. 12d, the modified structures is showing a single resonance frequency at 10.0 GHz with a bandwidth of 440.0 MHz below −10 dB from 9.88 GHz to 10.32 GHz. It is observed the bandwidth of the modified FSS slot has been reduced. The transmission bandwidth

of the structure netter than −1 dB is 650 MHz from 9.77 GHz to 10.42 GHz with insertion loss (IL) 0.05 dB at resonance frequency 10.0 GHz.

The EM performance of the proposed multi-slot unit cell is estimated using the full-wave analysis with the help of the commercial simulation package tool CST MWS. This tool has the advantage of the ease of designing, accuracy, and simplicity in optimization. The unit cell boundary conditions along x- and y-directions in conjunction with Floquet mode excitation are used in the z-direction to simulate the structure. Symmetry boundary conditions may apply to the structure to reduce the simulation time.

Material Characteristics

Nicolson–Ross–Weir (NRW) method is particularly interesting as a homogenization method for metamaterials and for studying their element characteristics. Although this approach has some functional difficulties, it has been popularly used to great effect to characterize metamaterial elements. The NRW technique has been employed here to extract the material behavior of the structure [Edward et al. (2016)]. The S-parameters used for the material characteristics extraction has been generated from the full-wave simulations. The computed material characteristics of the proposed swastika-shaped FSS are given in Fig. 13.

The structure is observed to exhibit negative permeability (μ-negative) and negative permittivity (ε-negative) characteristics at 10 GHz. Thus, the proposed Swastika-shaped tightly coupled unit cell shows double negative (DNG) metamaterial behavior in distinct microwave frequencies. The design of SIW based MTM-FSS is described in the following section.

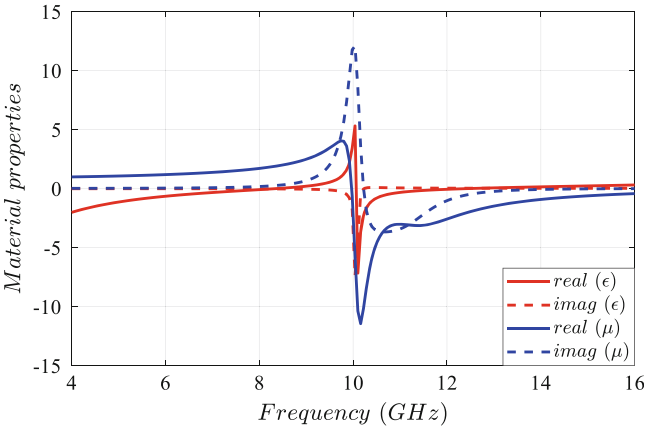

Fig. 13 Material characteristics of the MTM-FSS (swastika) unit cell

Design of MTM-FSS Unit Cell Based on SIW

The functional design of the FSS based on SIW is involved in the metallic cavity around the FSS slot in a unit cell. The SIW cavity is designed around the swastika-shaped MTM-FSS unit cell given in Fig. 12b. To form the metallic cavity, the MTM-FSS slot has been etched on either side of the dielectric substrate. Then, the metallic cavity with cylindrical vias has been formed around the unit cell as shown in Fig. 14a, b. The diameter (d_v) and pitch distance between the two adjacent vias

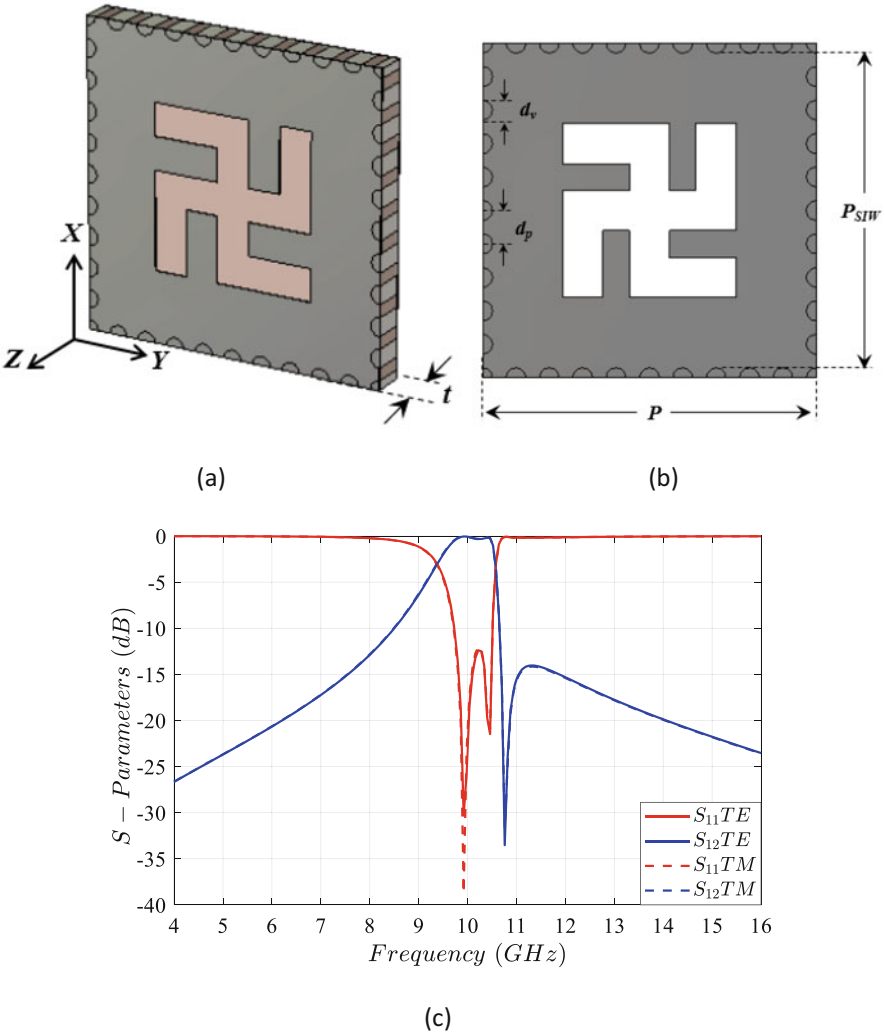

Fig. 14 Design of MTM-FSS unit cell based on SIW: (**a**) unit cell perspective view, (**b**) unit cell with notations, (**c**) frequency response

(d_p) is 0.9 mm and 1.5 mm, respectively. The resonant frequency for TM_{mn0} mode of the SIW cavity is:

$$f_{SIW} = \frac{c_0}{2\sqrt{\varepsilon_r}} \sqrt{\left(\frac{m}{P_{eff}}\right)^2 + \left(\frac{n}{P_{eff}}\right)^2} \tag{11}$$

where P_{eff}, the effective side length of the SIW cavity unit cell. In the proposed unit cell model, the cavity is square-shaped with side length p, hence:

$$P_{eff} = P - \frac{d_p^2}{0.95 d_v} \tag{12}$$

In the above equation, d_p and d_v are cylindrical via diameter and pitch distance between the metallic vias in the SIW cavity. The optimized dipole FSS slot dimensions for SIW based cavity are $L = 8.0$ mm and $S = 1.6$ mm. The metallic cavity generates an additional resonance based on the cavity size. As a result, the structure shows broad pass-band characteristics with sharp roll-off at a higher cutoff frequency and very good impedance matching below −10 dB as shown in Fig. 14c. The MTM slot element generates a single resonance frequency and the SIW cavity will generate the second resonance frequency. Tuning both resonances close to each other will results in a wide bandwidth. The proposed MTM-FSS based on SIW shows a wideband frequency response with the effect of slot and SIW cavity. A wideband with bandwidth 800 MHz from 9.7 GHz to 10.5 GHz. The −1 dB transmission of the structure is 880 MHz from 9.63 GHz to 10.52 GHz with a flat top response. The high-quality factor (Q) SIW structure results in a transmission zero at the immediate next of passband at a higher cutoff frequency band at 10.76 GHz.

The basic performance characteristics of conventional FSS may deviate from the estimated performance when used as an infinite surface. This may be due to the inter-element coupling of the adjacent unit cells. In the proposed SIW form, the metallic vias act as a wall and provide isolation among the FSS elements. Hence the response can remain unchanged even in an infinite array. This is an added advantage of the SIW cavity over the conventional FSS structure. A sharp roll-off at the edges of the passband is another notable advantage, which was hardly found in transmission type polarization rotators proposed in the open literature. The unwanted out-of-band frequency components can be eliminated using the sharp roll-off with high selectivity. This is a most desirable performance requirement for stealth antenna systems and invisible radomes. The mode analysis of the SIW cavity is performed using the Eigenmode analysis. The cavity modes have been estimated using characteristic mode analysis through an Eigenmode solver using CST MWS. The SIW cavity at the designed dimensions is exhibiting different operating modes and their operating frequencies are given in Table 1.

The existence of operating modes inside the SIW cavity is illustrated in detail in Fig. 15. As shown in Fig. 15, the surface current distribution of the MTM-FSS based

Table 1 SIW resonance frequency different Modes at 15.0 mm cell size

Mode	Resonance frequency
TM_{110}	10.13
TM_{120}/TM_{210}	15.86
TM_{220}	22.41
TM_{320}/TM_{230}	25.47

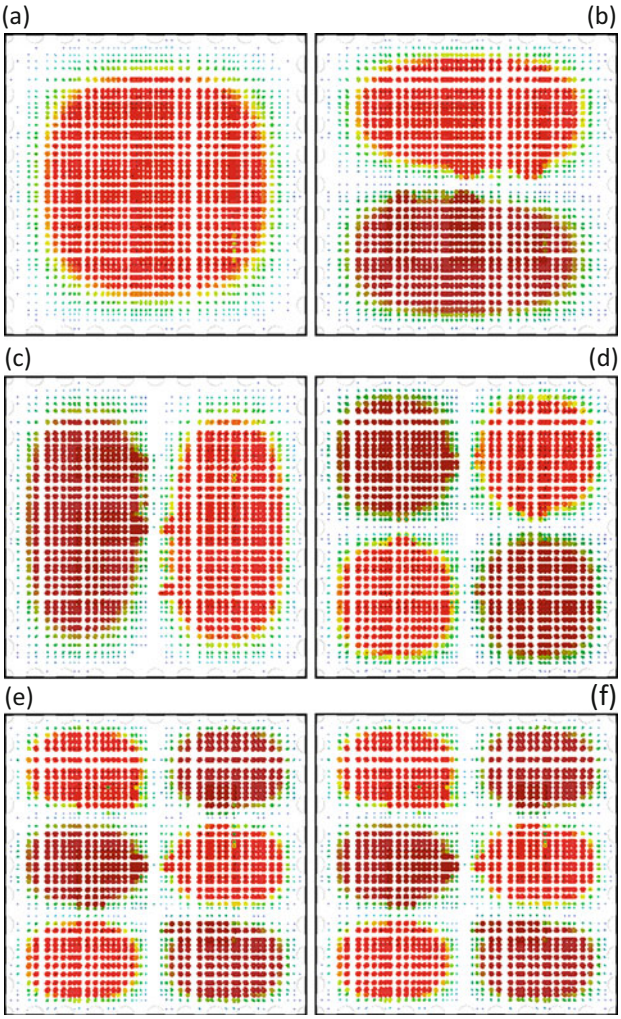

Fig. 15 Operating modes of SIW cavity: (**a**) TM_{110}, (**b**) TM_{120}, (**c**) TM_{210}, (**d**) TM_{220}, (**e**) TM_{230}, (**f**) TM_{320}

on the SIW cavity is normal at the edges of the SIW cavity. The etched swastika-shaped slot on the top and bottom surface can couple within the incident EM wave and allow the perturbations inside the SIW cavity. Also, the MTM-FSS shape

enables the unit cell to operate for all the polarization of incoming EM waves because of its horizontal and vertical arms, where the conventional cross elements are generally used for linearly polarized waves. Figure 15b, c, e, f) show the current distribution of degenerated modes, i.e., TM_{120}/TM_{210} and TM_{230}/TM_{320} are similar modes. At the designed unit cell cavity size, the TM_{120}/TM_{210} operating mode exists. The maximum field distribution occurs at the arms of the proposed swastika-shaped MTM-FSS element. Through which maximum EM energy is coupled through the FSS cross slots and undergo multiple wave perturbations inside the metallic SIW cavity and increases the overall bandwidth of the proposed cavity design.

By taking the superiority in the frequency response of the proposed MTM-FSS unit cell (see Fig. 14), a radome wall construction has been carried out. The MTM-FSS has been used as a core of the radome wall. The detailed description of the proposed radome wall is explained in the following section.

Radome Shape and Design

The radome shapes of an aircraft are mainly because of the structural and aerodynamic requirements of the aircraft. A peculiar blunt nose cone exposes a lesser surface area in the radome structure than the sharp nose cone structures. On the other hand, long conical shapes experience lesser skin air friction drag at supersonic and hypersonic speeds. The detailed wind flow and drag on the radome body are presented in Fig. 16. The aforementioned is the reason for having a blunt nose cone radome in commercial aeroplanes. However, fighter aircraft flew at supersonic (up to 2 Mach) and ultrasonic (2–5 Mach) speeds (1 Mach is equal to the speed of sound in an elastic medium, i.e., 1 Mach = 1235 km/h, at 300 K), which increases air friction drag on it [David Anderson and John D. Anderson (2001)]. The modern aircraft flew at speeds ranging from 1.5 to 7.0 Mach. At these supersonic speeds, very-high temperatures develop on the surface of aircraft radome due to air friction.

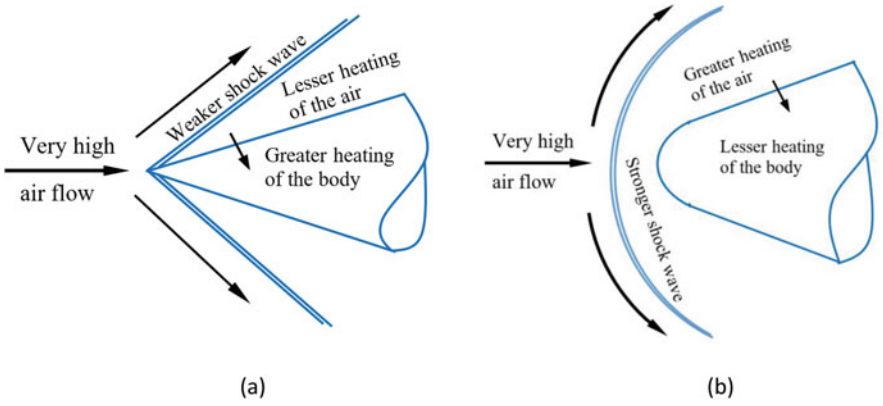

Fig. 16 Aerodynamic heating on aircraft: (**a**) slender reentry body, (**b**) blunt reentry body

The conventional fighter aircraft radomes are made with ceramic materials which are highly suitable to withstand up to 1500 K. Most of the FSS based FSS radomes presented in the open literature are designed with single-layer metallic arrays stacked in the dielectric walls. Also, when lightning strikes on the radome, it induces very high temperatures on the metallic layers inside the radome wall. The induced temperature of the body may be high enough to melt the thin metallic arrays in the embodiment structure. Hence, a metallic breakdown may occur in the FSS structure, and the system loses its basic purpose of filtering EM waves (Chap. 10 in Munk B A (2000)). Furthermore, the manufacturing process of conventional FSS elements is easy using a printed circuit board or metallic disposition method. However, the single-layer has limited in their performance in terms of low bandwidth, sensitivity to oblique incidence angle, selectivity, and high-power handling capabilities. The stacked, multilayered, and 3D structures may overcome these functional limitations, but makes the overall system heavy and bulkier, which is not a suitable solution for airborne platform.

The shape of conventional airborne radomes is described by a two-dimensional super-quadratic equation revolved around its main axis to obtain a body of revolution family of geometric shapes called super spheroids [Shavit, (2018)]. The pictorial representation of two-dimensional super-quadratic equation is shown in Fig. 17 and can be described mathematically by:

$$\left(\frac{x'}{B/2}\right)^{v} + \left(\frac{z'}{L}\right)^{v} = 1 \tag{13}$$

In the above equation, L is the radome length from base to tip and B is its base diameter ($L > B/2$). ν is a positive real number and $D = 2a$ is the antenna aperture diameter.

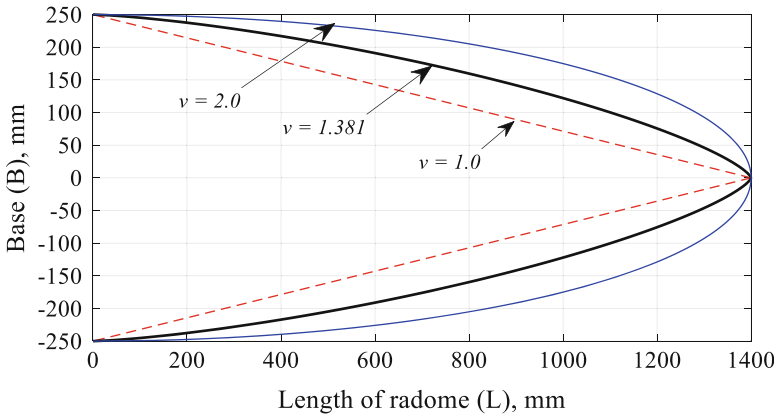

Fig. 17 Shape of the radome using super spheroidal equations

By revolving the above equation about the z-axis, a body of revolution (BOR) can be obtained and it is given by

$$x'^2 + y'^2 = \frac{(B/2)^2}{L^2}\left(L^v - z'^v\right)^{2/v} \tag{14}$$

The radome's surface $f(\rho', \phi', z')$ in cylindrical coordinates is not dependent on ϕ' due to symmetry and can be described by:

$$\begin{aligned} f(\rho', \varphi', z') &= x'^2 + y'^2 - \frac{(B/2)^2}{L^2}\left(L^v - z'^v\right)^{2/v} \\ &= \rho'^2 - \frac{(B/2)^2}{L^2}\left(L^v - z'^v\right)^{2/v} = 0 \end{aligned} \tag{15}$$

The value of 'ν' *is a* variable and interesting conical radome shapes can be obtained. In consequence, if $\nu = 2$, the above equation generates a prolate spheroid. The value of $\nu = 1.381$ yields the traditional von-Karman radome geometry and $\nu = 1$ generates the conical radome shape. The wide range of positive real numbers $1 < \nu < 2$ produces different types of radome shapes for aircraft and missile systems. The main advantage of this analytical cross-sectional model is the fact that all geometry parameters can be easily obtained mathematically.

Planar Radome Wall Design

The proposed MTM-FSS based on SIW has been further investigated for the planar radome application. A planar 5-layer radome wall structure is constructed by using the proposed unit cell as the center core of the radome as shown in Fig. 18. It consists of a core layer sandwiched between the two low dielectric spacer materials with dielectric constant, $\varepsilon_{r1} = 1.16$ and thickness, $t_1 = 2.0$ mm. A honeycomb or high-density foam material is preferred for the spacer layer. Then, the dielectric skin is applied on either side of the structure to form the complete 5-layer model. The dielectric parameters of the skin layer are: dielectric constant, $\varepsilon_{r2} = 3.5$ and loss tangent, $\tan\delta = 0.0026$. The thickness of dielectric skin, t_2 is 1.57 mm.

The effective thickness of the spacer and skin dielectric material can compute using Eqs. 1 and 2 [Liu N et al. (2018)]:

$$h_1 = \frac{\lambda 0}{4} \cdot \frac{(1 + \cos\theta_i) - \varepsilon_{r2}}{(\varepsilon_{r1} - \varepsilon_{r2})\sqrt{1 + \cos\theta_i}} \tag{16}$$

$$h_2 = \frac{\lambda 0}{4} \cdot \frac{\varepsilon_{r1} - (1 + \cos\theta_i)}{(\varepsilon_{r1} - \varepsilon_{r2})\sqrt{1 + \cos\theta_i}} \tag{17}$$

where λ_0 is the free space wavelength, θ_i is the angle of incidence. ε_{r1} and ε_{r2} are corresponding to the dielectric constants of spacer and skin dielectric layers.

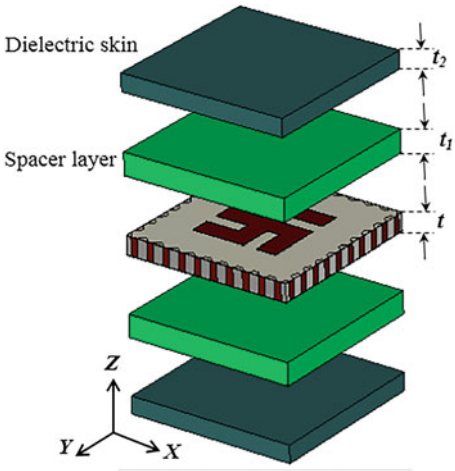

Fig. 18 Planar radome wall configuration (layers are expanded for better understanding)

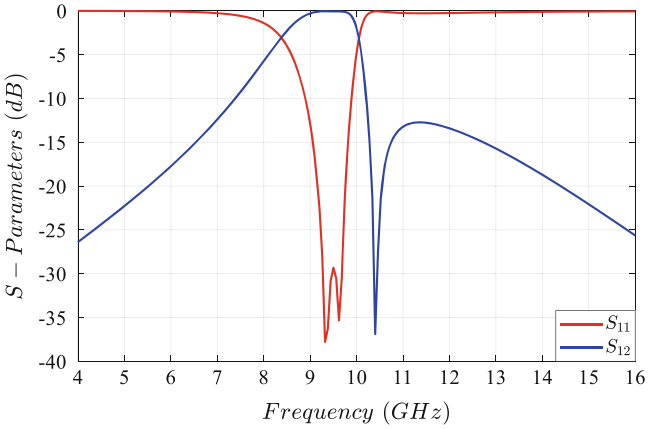

Fig. 19 Frequency response of 5-layer radome wall

The resonance frequency of the overall structure is optimized to operate at 9.6 GHz. The optimized design parameters of the unit cell are $P = 15.0$ mm, $L_1 = 7.8$ mm, $L_2 = 3.9$ mm, and $S = 1.8$ mm. The total thickness of the radome wall is $(h + 2 \times h_1 + 2 \times h_2)$ 8.71 mm.

The proposed radome wall simulations are performed using the full-wave analysis with periodic boundary conditions along x- and y-directions and open boundary conditions along the z-direction. A plane wave is excited using the Floquet mode to be carried out on the unit cell to analyze the reflection and transmission properties. The estimated frequency response of the proposed radome wall is shown in Fig. 19. The radome wall operates at a wide frequency range in the X-band regime. The response clearly shows the advantage of dual-mode resonance concerning

bandwidth and sharp roll-off at the edges of the transmission band which is the effect of SIW metallic cavities used in the design and flat top response in the transmission band. The dual resonance with wideband characteristics is happened due to the combined effect of TM_{120} mode and cross slot. The structure exhibits the reflection coefficient better than −10 dB in the X-band region from 8.8 GHz to 9.88 GHz with a maximum insertion loss better than 0.5 dB in the passband (at −1 dB) having a relative bandwidth of 12.84%.

EM Performance Analysis

Design Dimensions

The cross-slot dimensions of the structure define the operating frequency. The length and width of the slot are important parameters of FSS to influence the resonance frequency and bandwidth. The variations in the frequency response and bandwidth with change in cross-slot width are analyzed and the same is shown in Fig. 20. It is observed that as the width increases, the transmission zero shifts to higher frequencies due to the matching with wavelength. This will increase the bandwidth of the structure significantly. The slot width is varied from 1.4 mm to 2.0 mm with a step size of 0.2 mm and response is monitored respectively. Correspondingly, >90% transmission efficiency is achieved with a relative bandwidth of 10.2%, 11.5%, 12.6%, and 13.7%, respectively. Similarly, the variation in the bandwidth with changing slot length of the cross slot is shown in Fig. 21. As the length increases, the transmission zero shifts higher frequency.

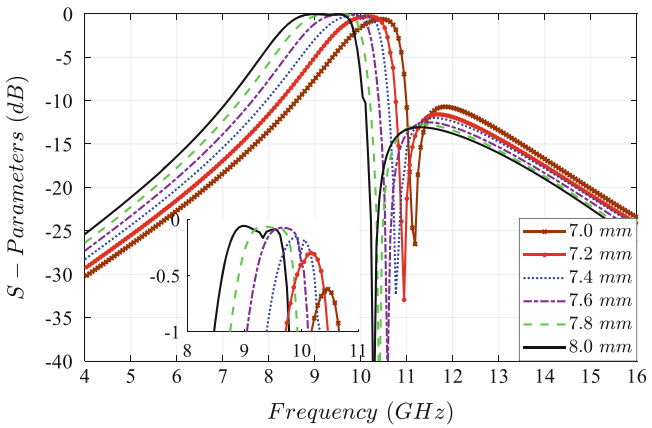

Fig. 20 Frequency response of 5-layer radome wall at $S = 1.4$ mm and different L_1

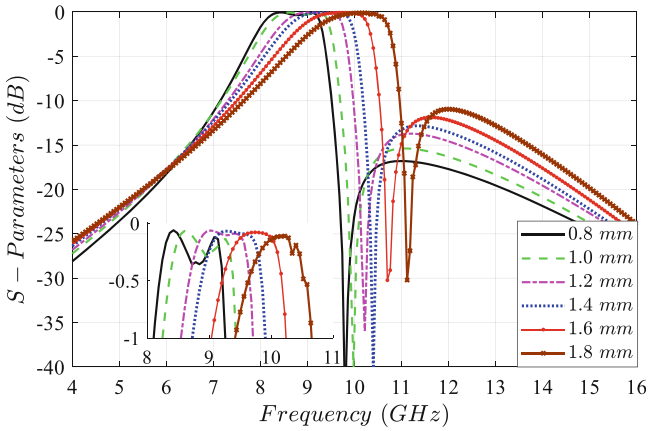

Fig. 21 Frequency response of 5-layer radome wall at $L_1 = 7.8$ mm and different 'S'

Oblique Incident Angles

In the practical usage of FSS, it can experience impinging EM waves in various angles from different directions. In addition, the frequency response suffers at higher impinging angles. So, the FSS designer needs to estimate the performance of FSS for oblique incidence angles. In this section, the frequency response of the proposed planar radome wall is studied at different polarizations for both vertical (TE) and horizontal (TM) polarizations.

The co-polarized reflection and transmission coefficient response is given in Fig. 22. It is observed that the response is more promising and acceptable till up to 40-degree incident plane wave. The higher-order grating lobes are found at a higher frequency because of the inclined wave impingement on the surface.

Fabrication and Experimental Verification

An FSS can be fabricated using a conventional printed circuit board (PCB) process for fabricating planar FSS. Any commercially available dielectric substrates can be used to create the metallic layers over them to achieve the desired EM performance with acceptable loss levels. Once, the FSS is fabricated using the PCB process, it can be embedded into the radome layers. Few fabricated prototypes of the FSS based on SIW have shown in Fig. 23. The planar radome wall fabrication is the easiest process by adding the laminates of layers. A 5-layer planar radome wall is shown in Fig. 24a. A free space measurement setup was used to measure the transmission coefficient of the radome wall shown in Fig. 24b. It consists of two horn antennas to transmit and receive the plane wave and a positioner to place the FSS. A network analyzer can be used to analyze the received signal through the radome wall. The center layer consists of the FSS core layer followed by two low

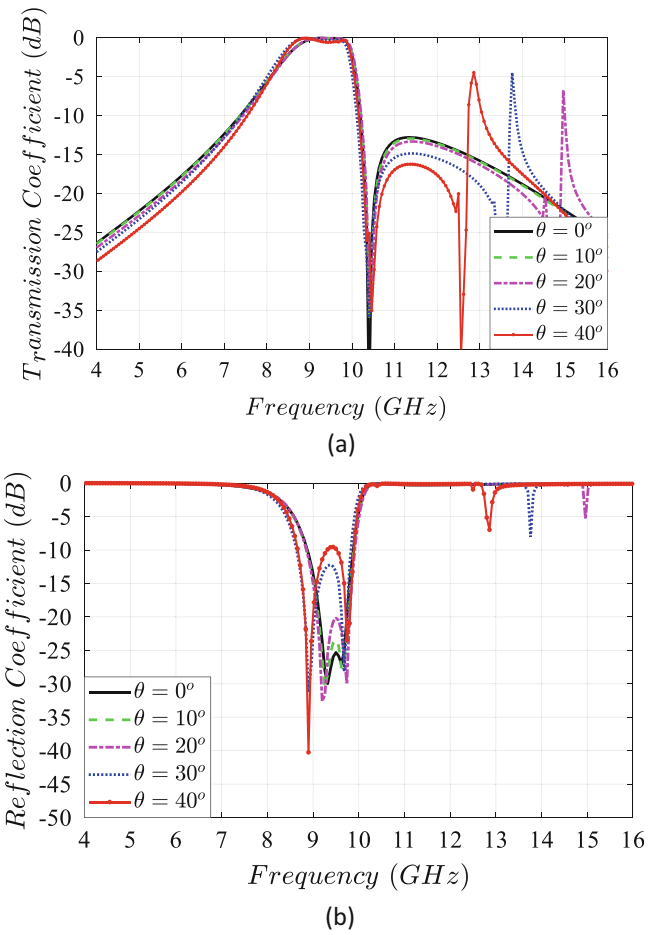

Fig. 22 Frequency response of 5-layer radome wall at oblique incident angles

dielectric constant layers (dense foam or honeycomb) followed by high dielectric constant skin layers, whereas the real challenge occurs with fabrication on conformal or conical radomes. Some of the most used radome substrate materials are given in Table 2. The conformal radome molds must be fabricated first and the FSS and other dielectric layers have to be added with utmost accuracy. Maintaining the thickness is the most important parameter to be considered while fabricating the conformal radome.

In practice, the fabrication of radome is generally done by a molding process. For the specific design of radome (including variable thickness), the molders are prepared first (masterpiece). The molders are filled with radome materials such as metal grids, ceramics, and dielectrics to get the shape of the radome. The additional layers may be added to the main design as required.

Fig. 23 Prototypes of fabricated FSS based on SIW

The fabrication of the FSS radome consists of the fabrication of the bandpass FSS pattern as per the design specifications and requirements of angular stability. Then, these metallic layers are inserted into the dielectric layers. Care should be taken care for the dielectric thickness of reinforced layers. The padding of metallic FSS based SIW layers to the homogeneous radome increases the total weight of the radome and this also includes the fabrication complexities. The technology to design and develop such complex-shaped radomes demands a sophisticated and dedicated manufacturing environment. This is a process that included huge capital and skilled designers and manpower. However, the EM performance of such radomes is superior in many ways, which is most demanded by modern airborne radars and electronic warfare systems.

The radome is equipped in the front of the aircraft; it has a higher tendency for physical damage. It is often oblivious that airborne and aircraft radomes are subject to unavoidable damages due to hail impact, heavy rains/winds, acid rains, lightning strikes, bird strikes, and bumping hangar doors or some maintenance equipment. The affected radome immediately becomes susceptible to water ingress, which badly impedes the radar performance. The damage in the radome leads to reflecting most of

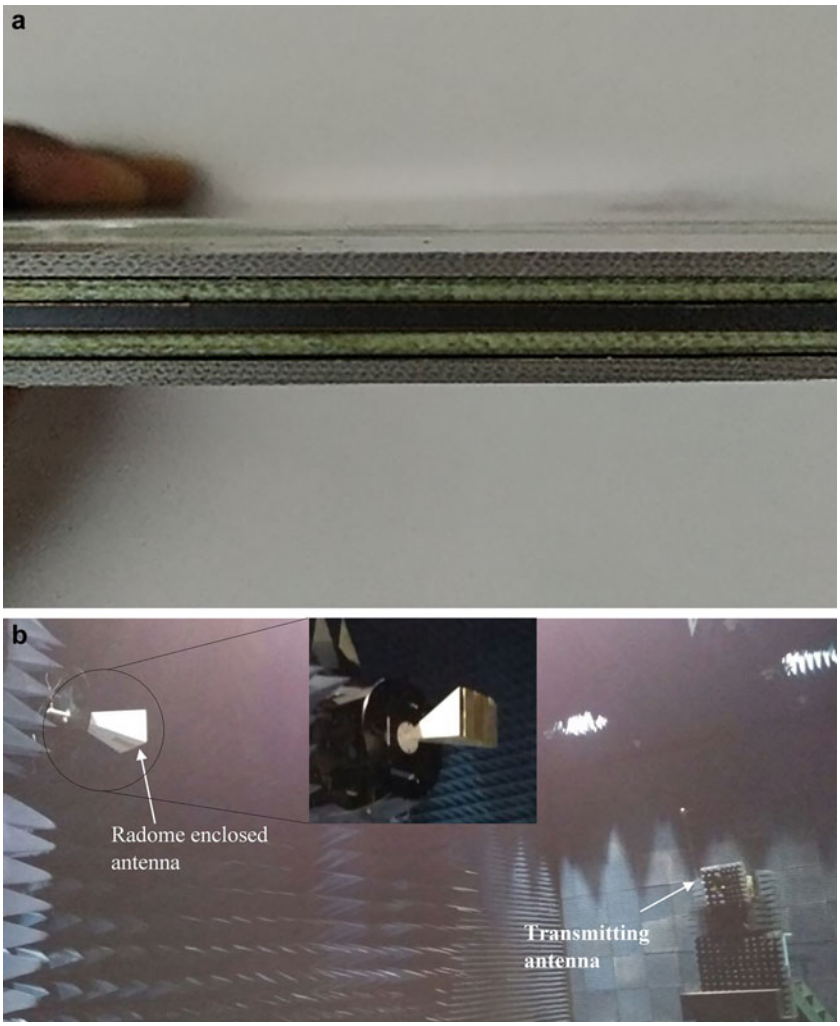

Fig. 24 (**a**) 5-layered radome wall with FSS based on SIW core layer at centre layer, (**b**) experimental setup of transmission coefficient measurement

Table 2 Dielectric properties of some plastic materials

	X-band		Ka–band	
Material	ε_r	tanδ	ε_r	tanδ
Rexolite	2.54	0.0005	2.6	0.002
Nylon	3.03–3.21	0.014–0.02	3.6	0.02
Teflon	2.1	0.0003	2.08–2.10	0.0006–0.001
Polystyrene	2.55	0.0004	2.54	0.00053–0.001
IparPlexi glass	2.59	0.015	2.61	0.02
Polyethylene	2.25	0.0004	2.28	0.007

the energy of radiating antenna due to the change in radome wall properties with induced water quantity into the damaged radome wall. Ultimately, no energy passes through the radome, effectively rendering the weather radar blind to anything that might be hiding in the shadow of that patch of water. Subsequently, false forecasting is made and confuses the piolet with wrong target information and in addition, the pilot cannot see any real targets. This renders the aircraft radar useless and complete damage to the aircraft and risks the life of the pilot.

The fabricated radome must undergo several EM, mechanical, and environmental tests before the actual usage to make it airworthy. The EM test includes Transmission loss, Boresight error, Side lobes, and back lobes. Mechanical tests include vibration, pressure, and rainwater leakage at the joints, etc. The Rain Erosion, Thermal shock, Transient Analysis, Fungus growth, Lightning and Bird strike, etc., will come under the environmental tests. The parameters should satisfy the Military standards to pass the tests and to use for real environmental conditions.

Conclusion

This chapter presented a concept of the frequency selective surfaces based on substrate integrated waveguide technology cavity in detail, a technique to control the bandwidth of FSS without disturbing design parameters and by applying a SIW cavity. Further, an MTM inspired FSS (MTM-FSS) is designed based on SIW technology for radome application. The design procedure and methodology of the MTM-FSS element are presented in detail. Further, the modeling of multilayer radome wall is discussed in brief. The important design parameters and oblique incidence are studied. Also, comments have been made on fabrication methodology and testing of the airworthy airborne radome.

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Metamaterial-Based High-Performance Engineered Surface for RCS Reduction

6

Fereshteh Samadi

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Abstract

In the past few years, the implementation of periodic structures for RCS reduction applications has been attracting attention. By engineering the geometric configuration of the unit-cell, the reflection phase can be manipulated toward achieving destructive interference in the far-field. Thus the unit-cell design with desired electromagnetic characteristics is one of the main factors toward achieving high performance-low RCS engineered surface. Another factor that significantly impacts the RCS reduction results is the arrangement of the unit-cells. Up to now, different types of unit-cells and arrangement methods have been introduced to enhance the RCS reduction bandwidth and minimize RCS in all directions. In this chapter, we will discuss and illustrate some design examples to validate the efficiency of implementing surface engineering for RCS reduction applications through experimental results.

Keywords

RCS reduction · Polarization conversion · Artificial magnetic conductors · Optimization algorithms

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Introduction

The radar cross-section (RCS) of an object refers to an efficient area that can be detected by radar. It depends on several significant factors of frequency, the angle of incidence, the material property, and its geometry (Jenn 1995). The RCS of an object is specified by two cases of monostatic and bistatic. The monostatic case refers to the condition that the same antenna is used to transmit and receive the signal. However, for the bistatic case, there are two antennas, one is specified as the transmitter, and another is the receiver (Skolnik 1990). From the RCS properties, several information such as the size, material, velocity, and distance of the object can be extracted. Therefore it is mainly implemented for detecting the targets, especially in military applications. In response to preventing the exposition of the target to the detecting radars, the RCS of the target should be minimized. Using the RCS reduction technology, the reflected signal from the target becomes very low, which can not be easily distinguished with adversary radars. Thus reducing RCS becomes a crucial subject in stealth technology.

RCS reduction has been receiving attention since World War II, where the missiles were covered with low reflective materials such as wood in order to reduce their detectability by radar. Subsequently, methods of shaping and applying radarabsorbing materials (RAM) have emerged and improved day by day. These methods, to some extent, were efficient in terms of reducing RCS, however; they suffer from unavoidable disadvantages. For instance, the method of shaping involves the alteration of the shape of the target to redirect the reflected wave toward the direction other than the receiver antenna, which dramatically reduces the efficiency of the target from an aerodynamic point of view. Moreover, the application of RAM usually is bulky and limited to narrow bandwidth due to its working principles. To overcome these shortcomings, scholars have proposed new approaches such as plasma-based and active RCS reduction. Although these methods contribute to wide bandwidth, their applications are constrained since applying plasma gives rise to infrared signatures and the method of active RCS reduction is difficult to implement. An alternative method is to engineer the reflected fields from a surface to create destructive interference, and hence reduce RCS. This can be achieved by engineering the reflection phase of metamaterials (Li et al. 2016), linear polarization converters (Qi et al. 2020), artificial magnetic conductors (AMCs) (Libi Mol and Aanandan 2018; de Cos et al. 2010), electromagnetic band gaps (EBGs) (Chen et al. 2015), and so on. This method is low cost and easy to implement; also, it could be properly designed to efficiently broaden the bandwidth and increase the oblique incident angle stability. The basic idea of the method is to design two reflection type unit-cell with in-phase and out-of-phase reflection phase properties, where by using the array theory, their superposition becomes zero at boresight. Mainly the RCS reduction results normalize compared to the RCS of perfect electric conductor (PEC) plate to eliminate the effect of distance (see (1)) (Chen et al. 2015).

$$\mathrm{RCS}_{\mathrm{reduction}} = 10\log\left[\frac{-\lim_{r\to\infty} 4\pi r^2 \frac{|E^x|^2}{|E^i|^2}}{-\lim_{r\to\infty} 4\pi r^2 (1)^2}\right] = 10\log\frac{|E^s|^2}{|E^i|^2}$$

if we use two types of unit-cells the equation of (1) can be simplified to (2) (Chen et al. 2015):

$$\mathrm{RCS}_{\mathrm{reduction}} = 10\log\left(\frac{\Gamma_A\left|e^{j\phi A}+\Gamma_B\right|e^{j\phi B}}{2}\right)2$$

where $|\Gamma_A|$ and $|\Gamma_B|$ are reflection amplitudes for the two types of unit-cells, respectively, meanwhile, 0_{A} and 0_{B} are associated with the reflection phases of these unit-cells. Since the unit-cells are reflection type and grounded with PEC plate, $|\Gamma_A|$ and $|\Gamma_B|$ are ~1. Therefore, in order to design a low RCS surface, the reflection phases of 0_{A} and 0_{B} are of importance and should follow (3) to obtain RCS reduction less than −10 dB (Chen et al. 2015).

$$180^0 + 37^0 < \phi_A - \phi_B < 180^0 + 37^0$$

In this chapter, we specifically focus on linear polarization converter and AMC-based unit-cells to satisfy the phase properties required for RCS reduction application. We also discuss some arrangement methods and compare their efficiency in terms of maximum RCS reduction in normal direction and uniform distribution of the scattered pattern.

RCS Reduction Using Polarization Converter Metamaterials

In recent years, the application of polarization converter unit-cells in RCS reduction has become a research interest (Jia et al. 2017; Li et al. 2016). This is due to the fact that the reflection phase difference between a polarization rotator unit-cell and its 90° rotated one is 180° which satisfies the phase cancellation criteria for the method of destructive interference. Up to now, numerous linear polarization converters based on anisotropic metasurfaces and chiral metamaterials have been presented (Xia et al. 2017; Feng et al. 2013; Wei et al. 2011). In (Gao et al. 2015), a reflective type V-shaped co-to-cross linear polarization converter with a high conversion efficiency is introduced. Also, to achieve broad bandwidth and high polarization conversion efficiency, several designs such as multi-resonance (Feng et al. 2013) and multi-layer (Wei et al. 2011) polarization converters have been proposed. Moreover, anisotropic metamaterials are introduced as polarization converters, where they are characterized by simple configuration and near-complete cross-polarization conversion (Xu et al. 2018). Usually, the efficiency of the polarization conversion is evaluated

through polarization conversion ratio (PCR), where in most applications PCR more than 90% are considered as an efficient polarization conversion. Since the unit-cell of reflective type- anisotropic polarization rotator has a symmetric patch along the diagonal axis, the PCR result for y-polarized incident wave is the same as the one for x-polarized incident wave. The value of the PCR can be calculated using (4):

$$\mathrm{PCR} = \frac{r_{xy}^2}{r_{xy}^2 + r_{yy}^2}$$

where $r_{yy} = E_{ry}/Ei_y$ and $r_{xy} = E_{rx}/Ei_y$ are co- and cross-polarized reflection coefficients in case the structure is illuminated by y-polarized electric field. Figure 1 shows some examples of polarization converter unit-cells with wide bandwidth and high PCR value. All examples are shown in Fig. 1 (Example #1, Example #2, and Example #3) consist of a diagonal patch that is backed by uniform ground plain for ideal reflection. In Example #1 (Samadi et al. 2018b), an air foam is placed between substrate and patch to increase polarization conversion efficiency. For this structure, a bandwidth of about 112% at the center frequency of 10 GHz for PCR more than 90% is achieved. In the second design, Example #2, the bandwidth and PCR ratio is improved by implementing five diagonal vias that connect PR patch to its uniform ground plane. Moreover, a plain superstrate is loaded on the top of the metallic patch

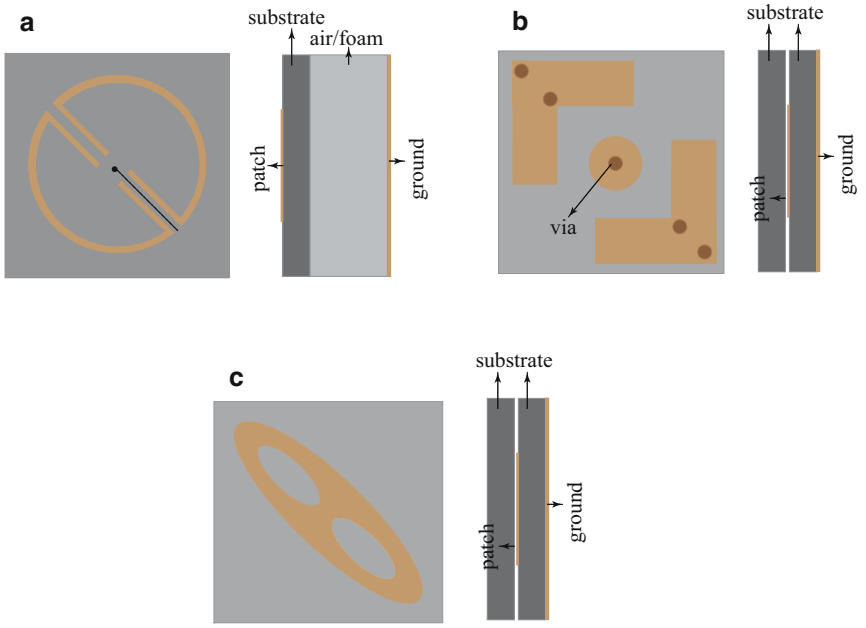

Fig. 1 Examples of some recent polarization converter unit-cells (**a**) Example #1, from Ref. (Samadi et al. 2018b). (**b**) Example #2, from Ref. (Samadi et al. 2018a). (**c**) Example #3 with permission from Ref. (Samadi et al. 2019). Copyright @ 2019 by the IEEE

(Samadi et al. 2018a) to improve PCR results. This design configuration contributed to 107% bandwidth for PCR more than 90%; also, it showed wide incident angle stability up to 40°. Finally, in the last design, Example #3 (Samadi et al. 2019), a diagonal ellipse shape geometrical patch with two slots loaded with a dielectric superstrate is introduced, which resulted in multiple resonances and subsequently wide bandwidth of about 115% with large incident angle stability up to 40°. Figure 2 depicts the co- and cross-reflection coefficients and PCR results for the last example shown in Fig. 1c (Samadi et al. 2019). These results demonstrate that the PR unit-cell design proposed in (Samadi et al. 2019) converts the co-polarization to the cross-one with high efficiency through a wide frequency range of 5.5–20.5 GHz. In this frequency range, the reflection phase difference between the PR unit-cell and its mirror is 180° which makes these types of unit-cells suitable for RCS reduction based on destructive interference. Figure 3a depicts the PR unit-cell of Example #3

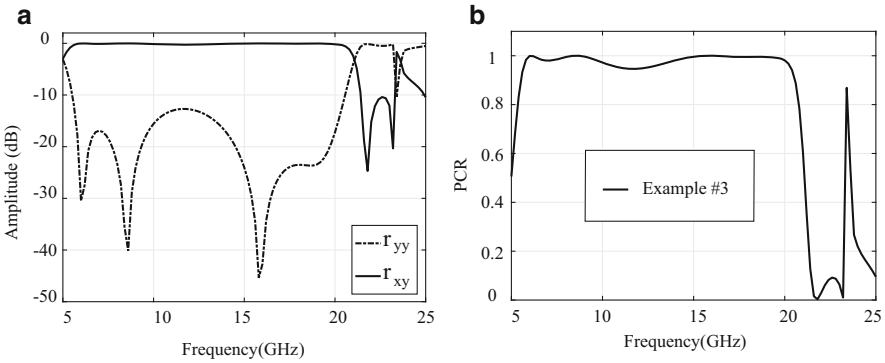

Fig. 2 The results of polarization conversion efficiency for Example #3 (**a**) Co and cross-reflection amplitudes, (**b**) PCR. With permission from Ref. (Samadi et al. 2019). Copyright @ 2019 by the IEEE

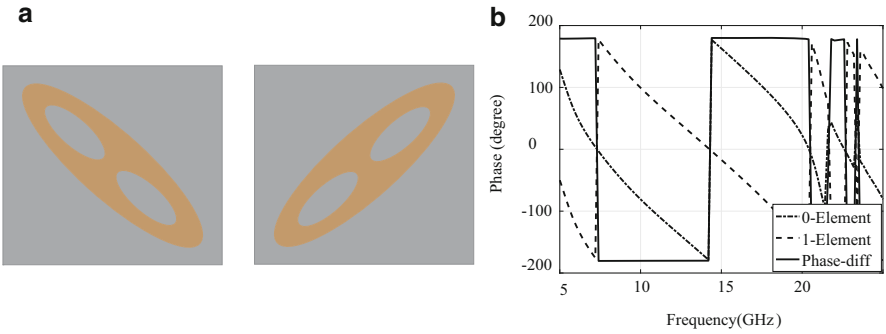

Fig. 3 The concept of reflection phase difference for Example #3 (**a**) schematic of PR unit-cell and its mirror, (**b**) The results of reflection phase for PR unit-cell, its mirror, and the reflection phase difference between them. With permission from Ref. (Samadi et al. 2019). Copyright @ 2019 by the IEEE

and its 90° rotated one. Their reflection phase coefficients and the phase difference between them are also illustrated in Fig. 3b. As it can be seen from Fig. 3b, the reflection phase difference between PR unit-cell and its mirror is 180° which satisfies the criteria required for destructive interference of electric field.

The unit-cell characteristics mainly control the bandwidth and incident angle stability of the low RCS surface. While the arrangement of the unit-cells mostly controls the level of RCS and the distribution of its scattering pattern. Meanwhile, the arrangement configuration also has a great impact on bandwidth. In order to find the appropriate arrangement of unit-cells, first n x m number of unit-cells are considered as antenna elements (here we call them super-cells), then the array theory is applied to find the proper arrangement that contributes to the maximum power cancellation. Thus far, several studies have been carried out on the arrangement of the super-cells to efficiently reduce RCS and uniformly distribute the scattered pattern (Chen et al. 2015; Khan et al. 2019; Yang et al. 2016a; Liu et al. 2017). The initial design arrangement was a checkerboard-like pattern (Paquay et al. 2007), which remarkably minimizes RCS in the boresight direction, however; the level of RCS was high in other directions. Toward reducing RCS in other directions, several design arrangements such as hexagonal (Chen et al. 2015), triangular (Khan et al. 2019) and, random (Yang et al. 2016b) have been introduced. Furthermore, optimization algorithms are implemented to intelligently find the arrangement that contributes to the lowest RCS in all directions (Zhao et al. 2016; Yang et al. 2016a; Liu et al. 2017; Wang et al. 2014; Samadi et al. 2019). In (Samadi et al. 2019), the effect of arrangement on RCS reduction and scattering pattern is investigated for three arrangements illustrated in Fig. 4, where the design configuration depicted in Fig. 4c is obtained through group search optimization (GSO) algorithm (Samadi et al. 2019) subject to achieving the minimum RCS in all directions.

Figure 5 demonstrates the normalized RCS reduction results for these three arrangements. To be noted that in all these three designs, 3 × 3 number of unit-cells of example #3 is implemented. Considering RCS results of Fig. 5, it can be realized that optimized arrangement results in further RCS reduction and wider bandwidth. Figure 6 shows the distribution of scattered pattern for the three mentioned arrangements at the frequency of 15 GHz. The scattered pattern of the PEC plane is also depicted for comparison purposes. As it can be seen from Fig. 6, structure #1 distributes the scattered pattern to two major lobes, structure #2 distributes it to four major lobes, while the coded structure distributes the power to several lobes with very uniformed scattering pattern. Therefore, a high level of RCS reduction in both monostatic and bistatic cases can be achieved by implementing efficient optimization algorithms toward finding optimal arrangements.

RCS Reduction Using Reflection Type Metamaterials

The RCS of an object can be minimized by efficiently engineering the EM properties of the state-of-the-art metamaterials, FSSs, AMCs, EBGs, to name a few. As it is discussed in the introduction, one of the efficient approaches to achieve low RCS

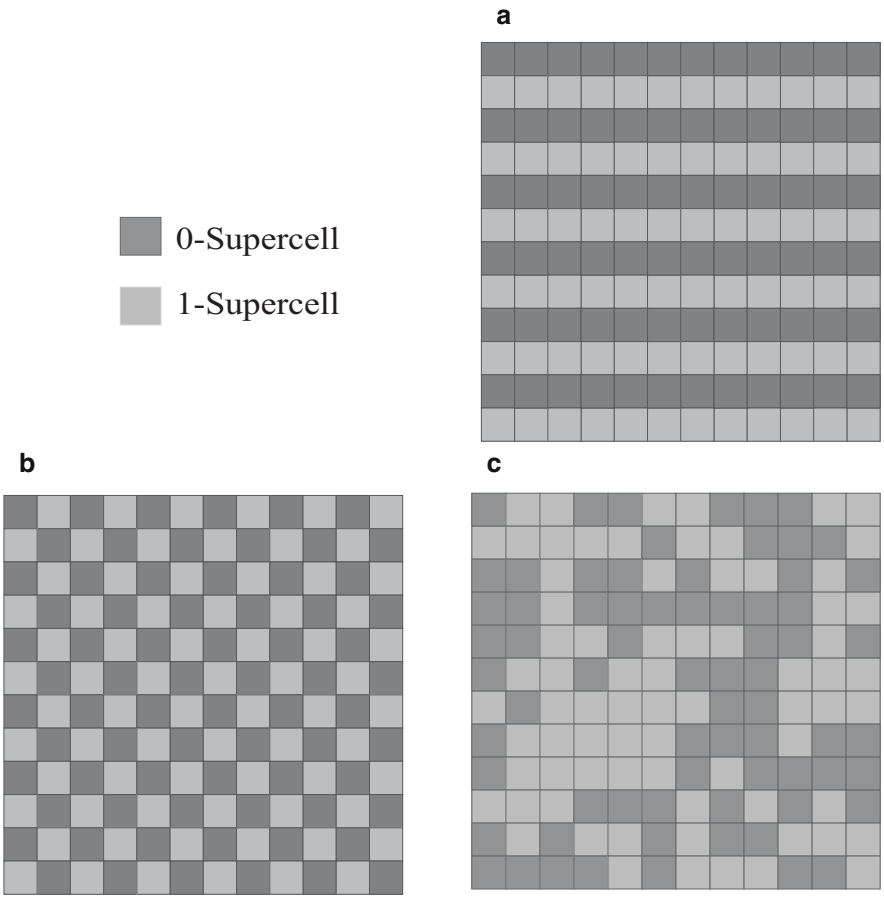

Fig. 4 Examples of different arrangement types (**a**) Arrangement #1, (**b**) Arrangement #2; checkerboard arrangement, and (**c**) Arrangement #3; coded arrangement, with permission from Ref. (Samadi et al. 2019). Copyright @ 2019 by the IEEE

surfaces is to design at least two reflection type unit-cells where their reflection coefficients of $|\Gamma A| < \phi A^{\circ}$, and $|\Gamma B| < \phi B^{\circ}$ satisfy the criteria in (3), with $|\Gamma A|$ and $|\Gamma B|$ are almost equal to one. To design two unit-cells satisfying the phase criteria mentioned in equation (3) within a wide bandwidth, numerical method available by full-wave-based commercial software, optimization-based methods (Chen et al. 2015; Wang et al. 2014), or recent machine learning-based approaches (Samadi and Sebak 2020; Panda et al. 2013) can be implemented. Thus far, several studies have been carried out aiming at increasing the reflection phase difference bandwidth criteria by introducing various geometrical configurations and design techniques (Chen et al. 2015; Samadi et al. 2017; Samadi and Sebak 2020a, b; Samadi et al. 2020). In (Chen et al. 2015; Samadi et al. 2017) geometrical pattern of the metallic patch is intelligently engineered to create multiple resonances and subsequently

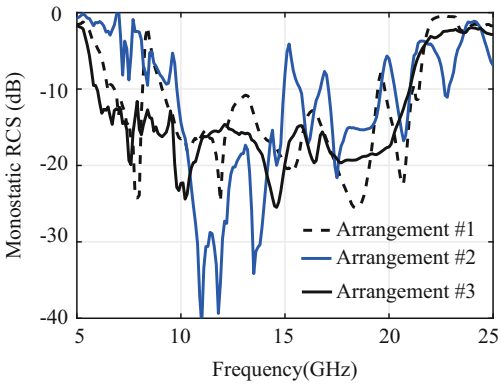

Fig. 5 Normalized RCS reduction results for (**a**) Arrangement #1, (**b**) Arrangement #2, and (**c**) Arrangement #3, with permission from Ref. (Samadi et al. 2019). Copyright @ 2019 by the IEEE

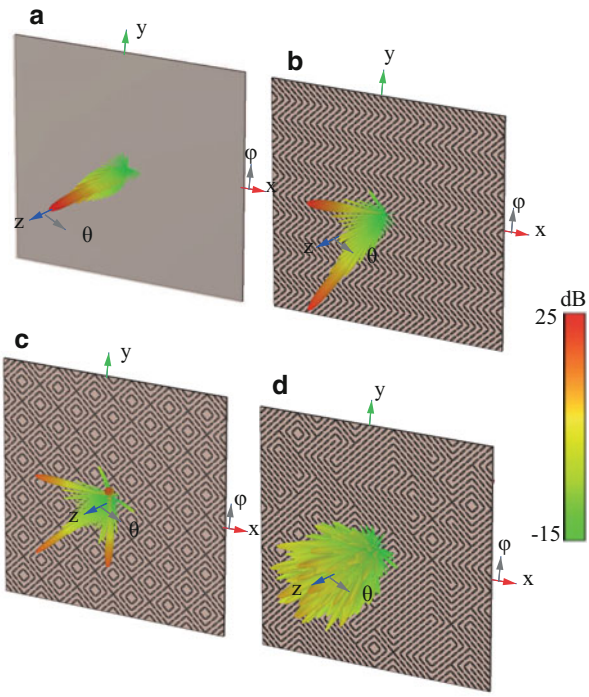

Fig. 6 3D scattering pattern at 15 GHz for the (**a**) PEC, (**b**) design with arrangement #1, (**c**) design with arrangement #2, and (**d**) Design with arrangement #3. (With permission from Samadi et al. 2019). Copyright @ 2019 by the IEEE

increase the bandwidth. Furthermore, dielectric-based unit-cells are reported in (Samadi et al. 2020; Taher Al-Nuaimi and Hong 2016), where by adjusting the relative permittivity of the unit-cell, the reflection phase difference criteria are effectively satisfied in a wide bandwidth. Most recently, a combination of the dielectric resonator and multi-resonance metallic patch design is proposed to effectively control the reflection phase difference property through a wide frequency

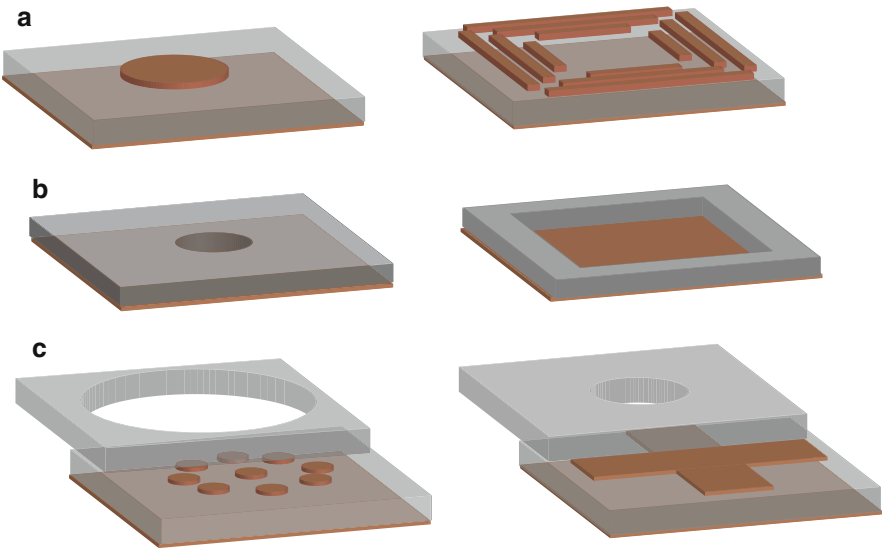

Fig. 7 Three different design examples of AMC unit-cells for RCS reduction purposes (**a**) Example #1, Ref. (Samadi et al. 2017). (**b**) Example #2, Ref. (Samadi and Sebak 2020a, b) (**c**) Example #3, with permission from Ref. (Samadi et al. 2020). Copyright @ 2020 by the IEEE

range (Samadi et al. 2020). Figure 7 shows some examples of reflection type unit-cells designed for RCS reduction application.

In the first example, that is, Example #1, Fig. 7a (Samadi et al. 2017), a circular patch and a multi-resonance square were properly designed to achieve wide bandwidth of 51.3% for the reflection phase difference criteria of RCS reduction. To implement these unit-cells for RCS reduction application, a 6×6 numbers of them are considered as super-cell and arranged in checkerboard-like format. For this surface, the normalized monostatic RCS is investigated, where 10 dB RCS reduction through the bandwidth of 51.4% is achieved. However, the 3D bistatic RCS showed that the scattering pattern has four major lobes, which results in a high RCS level in other directions. Furthermore, this structure was not analyzed for oblique incidence, which is an important consideration for real-life low-RCS surface design.

In the second example, Example #2, Fig. 7b (Samadi and Sebak 2020a, b), dielectric resonators are implemented where by engineering the dielectric substrate, the reflection phase coefficient is adjusted. The unit-cell was designed to achieve wide incident angle stability at mmWave frequencies. For the two unit-cells of Example #2, the reflection phase difference criteria are satisfied for the frequency range of 59–73 GHz with the oblique incident stability up to 40°. For this design, the unit-cells were arranged in triangular type format to form a super-cell. Afterward, the triangular type super-cells were arranged in a conventional checkerboard-like configuration. For this design configuration, the scattered pattern is distributed to six major lobes, which results to further RCS reduction in other directions.

Finally, Example #3, Fig. 7c is proposed in (Samadi et al. 2020) to improve the bandwidth and incident angle stability. In this new configuration, a metallic- patch-based unit-cell is loaded with perforated superstrate to have control over relative permittivity during the design process. A reflection phase difference bandwidth of 111.5% at the center frequency of 37.25 GHz is achieved for Example #3. Meanwhile, the reflection phase difference bandwidth is also investigated under oblique incidence where a bandwidth of more than 54% is achieved for the maximum oblique incidence of 60°. Figure 8 depicts the reflection phase difference versus frequency under oblique incidence up to 60°, which verifies the wide angle stability of Example #3.

In order to use the unit-cells of Example #3 for RCS reduction applications, different sizes of rectangular-type super-cells are considered, which is first investigated in (Edalati and Sarabandi 2014) with a random selection of super-cell sizes. In this study, through simulation and measurement results, it is verified that the various sizes of super-cells result in efficient distribution of the scattered pattern in different directions. However, in (Edalati and Sarabandi 2014), the super-cell sizes are selected randomly. To improve the design concept of (Edalati and Sarabandi 2014), the super-cell size selection is optimized in (Samadi et al. 2020), where the concept of coded arrangement is also included during the arrangement process instead of simply arranging them in checkerboard-like format. It is shown that optimizing super-cell size along with the conventional coded arrangement (Samadi et al. 2019) results in better RCS reduction results for both monostatic and bistatic cases. Figure 9 illustrates the configuration pattern for the two models, coded with optimized supercell sizes (Structure #1) and conventional coded arrangements (Structure #2). The first arrangement is obtained from the algorithm presented in (Samadi et al. 2020) and the second one is optimized using the one developed in (Samadi et al. 2019). The normalized monostatic RCS reduction results for the two models of arrangement are depicted in Fig. 10. As it can be realized from Fig. 10, the first structure consisting of the optimized super-cell sizes results in further RCS reduction level and wider bandwidth compared to Structure #2 which consists of predetermined same super-cell sizes. The results of the bistatic RCS reduction are also investigated as illustrated in Fig. 11 (Samadi et al. 2020). By analyzing the results of

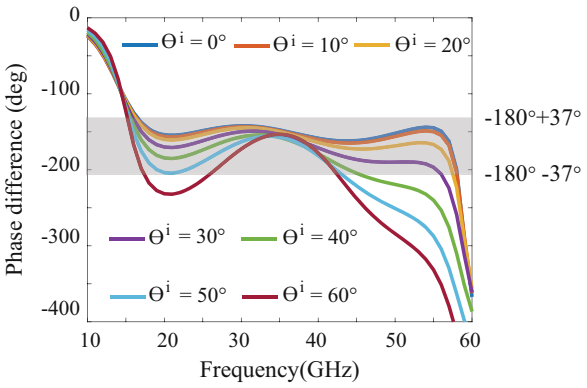

Fig. 8 Reflection phase difference versus frequency under different oblique incidence, for the unit-cells of Example #3 in Fig. 7c, with permission from Ref. (Samadi et al. 2020). Copyright @ 2020 by the IEEE

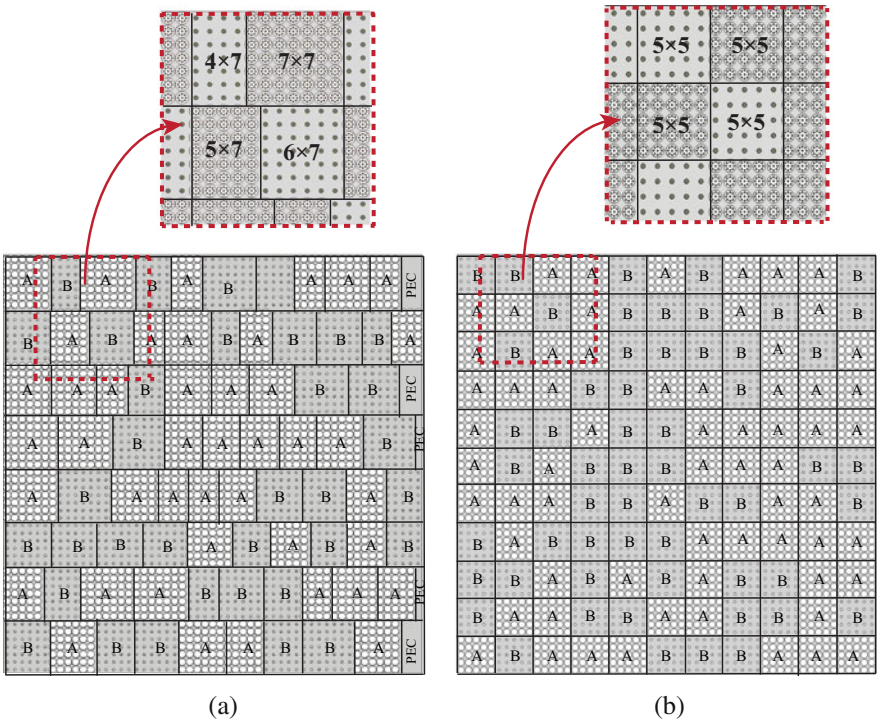

Fig. 9 Schematic of the configuration pattern of two low RCS structures consists of the unit-cells of Fig. 7c. (**a**) Structure #1: Coded with optimized super-cell sizes, and (**b**) Structure #2: Conventional coded, with permission from Ref. (Samadi et al. 2020). Copyright @ 2020 by the IEEE

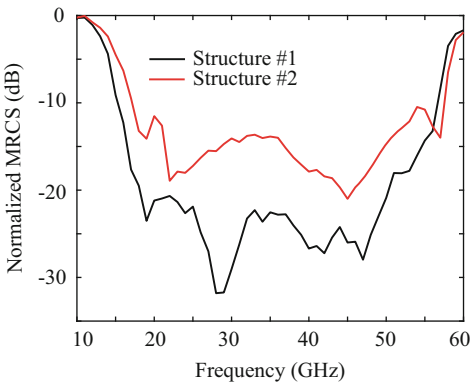

Fig. 10 Normalized monostatic RCS reduction results versus frequency for Structure #1 and #2. Ref. (Samadi et al. 2020). Copyright @ 2020 by the IEEE

Fig. 11 it can be comprehended that the coded structure with optimized super-cell size, Structure #1, results in a more uniformly scattered pattern compared to the conventional coded one, Structure #2. The bistatic scattering pattern of the same size PEC is also

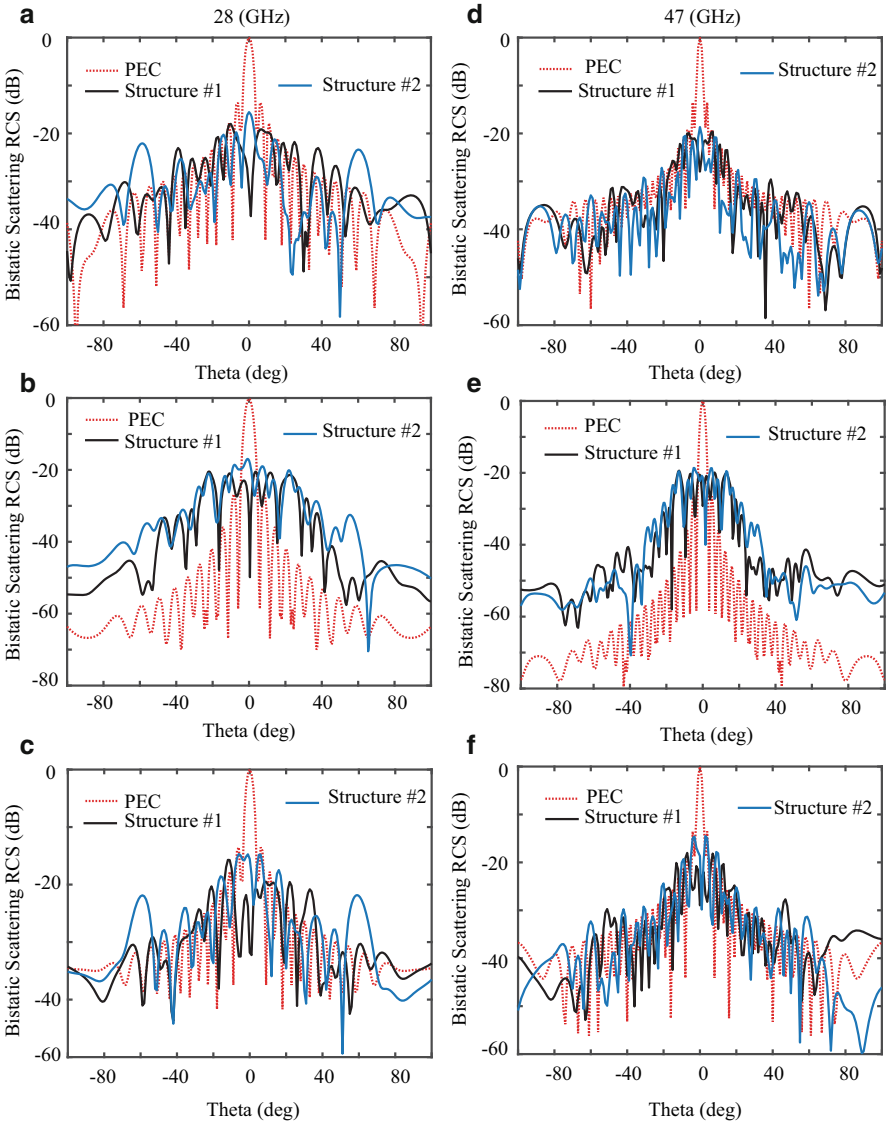

Fig. 11 Bistatic RCS patterns versus theta: (**a**) $\phi = 0°$ at 28 GHz, (**b**) $\phi = 45°$ at 28 GHz, (**c**) $\phi = 90°$ at 28 GHz, (**d**) $\phi = 0°$ at 47 GHz, (**e**) $\phi = 45°$ at 47 GHz, and (**f**) $\phi = 90°$ at 47 GHz (From Samadi et al. 2020). Copyright @ 2020 by the IEEE

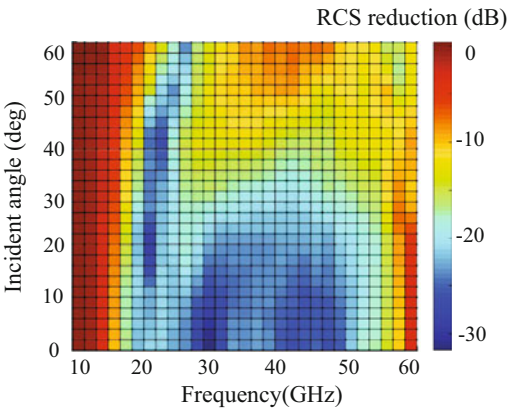

Fig. 12 Normalized RCS reduction versus frequency and oblique incidence, with permission from Ref. (Samadi et al. 2020). Copyright @ 2020 by the IEEE

included in Fig. 11 to clearly compare the level of RCS reduction of the two structures. The RCS reduction results of structure are also investigated under oblique incidence. Figure 12 depicts the 10 dB normalized RCS reduction of Structure #1, versus frequency and incident angle. As it can be seen from this figure, the Value of RCS is less than −10 dB for large frequency range up to 60°. The results for 3D scattering pattern of Structure #1 are also depicted in Fig. 13 to evaluate the bistatic RCS reduction results. Figure 13 shows that the 3D scattering pattern distributes very uniformly for the incident angles of 0°, 20°, 40°, and 60°.

Summary

In this chapter, we introduced high performance engineered surface for **RCS** reduction. By properly manipulating the EM characteristics of the metamaterials, in phase and out-of-phase electric fields are generated, and by appropriately arranging them, destructive interference is achieved, which minimizes the **RCS** of the surface. This method offers an efficient and low-cost design of low **RCS** surface with wide bandwidth and wide incident angle stability. Here, we discussed several design approaches, where different types of unit-cells and arrangement configurations were implemented to increase RCS reduction bandwidth and minimize the level of monostatic and bistatic **RCS.** Through the literature review that has been provided here, we conclude that implementing metamaterials and periodic structures can efficiently reduce the **RCS** of a structure with low cost.

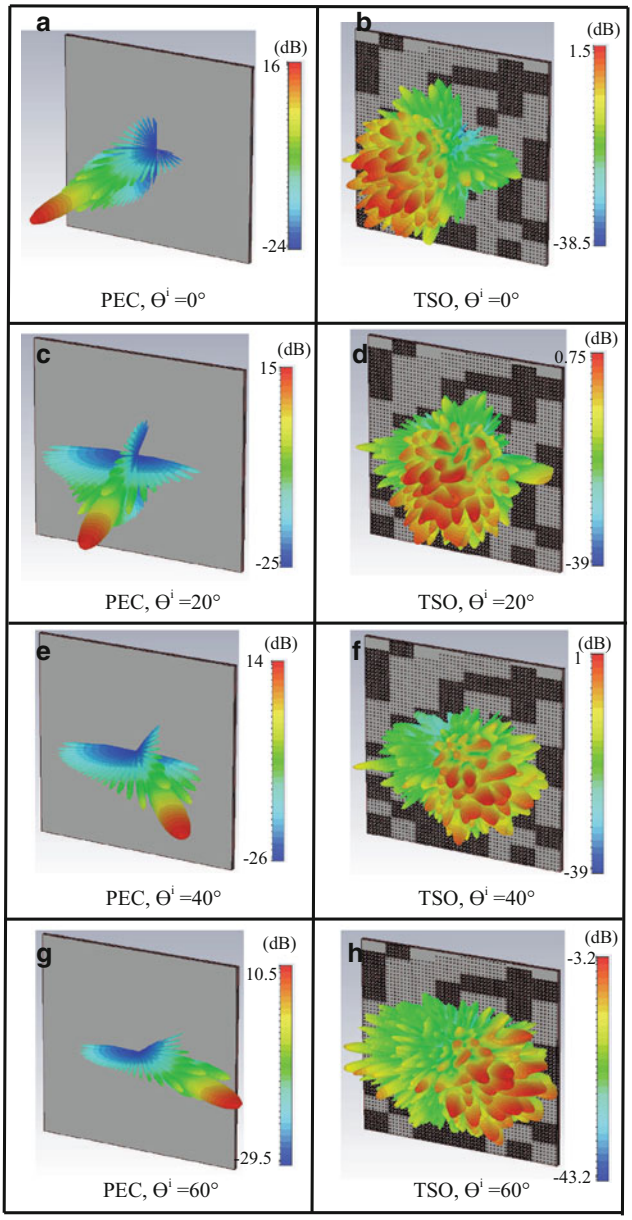

Fig. 13 3D scattering pattern of PEC and structure #1 in Fig. 9a for different incident angles, θs (**a**) at θ = 0° for PEC, (**b**) at θ = 0° for structure #1, (**c**) at θ = 20° for PEC, (**d**) at θ = 20° for structure #1, (**e**) at θ = 40° for PEC, (**f**) at θ = 40° for structure #1, (**g**) at θ = 60° for PEC, and (**h**) at θ = 60° for structure #1. (With permission from Samadi et al. 2020). Copyright @ 2020 by the IEEE

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Increased RCS for Enhanced Detection by Radars for Road Safety Applications

7

Mohammed Kalaagi and Divitha Seetharamdoo

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Abstract

Metasurfaces are a well-known interesting topic nowadays for many researchers and for various applications. The idea of anomalous reflection is proposed to demonstrate the concept of retrodirectivity. Retrodirectivity can be very important property particularly for radar applications at microwave and millimeter wave frequencies where the enhancement of the radar cross section of poorly scattering targets is needed. Many conventional topologies for such feature are presented in the literature such as corner dihedrals and Van Atta arrays; however, these topologies can be bulky and difficult to fabricate. Indeed metasurfaces can be compact and of remarkable physical properties, the main challenge is to design retrodirective metasurfaces that can operate for multiple incident angles simultaneously. In this chapter, we introduce various techniques to design multi-angle retrodirective metasurfaces from cascading, surface impedance modulation to Fano resonance. The design procedure is given following the generalized law of reflection and measurements have been carried out for several prototype designs to validate the simulation results. From the applicative aspect, this project is in the framework of project CYCLOPE, which is dedicated to protect vulnerable road users such as cyclists and pedestrians.

To demonstrate the potential of retrodirective metasurfaces for such applications, measurements have been carried out using a mannequin alongside the metasurface designs. The metasurface design showed remarkable results by increasing the radar cross section of the mannequin when implemented alongside the model.

Keywords

Metasurfaces · Retroreflection · Multiple angles · Cascading · Impedance modulation · Fano resonance · Dual polarization · Radar cross section · Road users

Introduction

Metasurfaces (Chen et al. 2016; Glybovski et al. 2016) have gained exceptional interest for their capability to control the propagation of an electromagnetic wave in unconventional ways (Sun et al. 2012; Holloway et al. 2012). Several functional designs have been introduced based on metasurfaces for different applications in terms of electromagnetic wave manipulation, including beam guiding (Pors et al.

2013b; Pors and Bozhevolnyi 2013; Pors et al. 2013a), conversion of propagating waves into surface waves Zhang et al. (2016), multifunctional metasurfaces (Elsakka et al. 2016), polarization control (Pfeiffer and Grbic 2013), hologram (Zheng et al. 2015), both at optical and at microwave frequencies. Different metasurface designs have also been developed for dual polarization applications (Kalaagi and Seetharamdoo 2018; Yurduseven and Smith 2017; Cala'Lesina et al. 2015), as well as for broadband frequency designs (Chen et al. 2018; Liu et al. 2014; Li et al. 2015). In terms of wave control and manipulation based on metasurface designs, retrodirective metasurfaces (Kalaagi and Seetharamdoo 2018) have proved to be of interest for applications such as radar cross section (RCS) enhancement (Knott 2012; Guo et al. 2008), specifically for radar targets for which the retro-scattered power level is low. This phenomenon is established according to Snell's law of reflection and refraction for a standard metallic flat plate or a mirror at normal incidence.

Previously, Yu et al. have demonstrated that using Snell's generalized law of reflection and refraction (Yu et al. 2011), i.e., by implementing a phase gradient on a flat plate, a higher control of an EM wave in a chosen direction is obtained (Sun et al. 2012). Retrodirective reflectors have been developed by (Doumanis et al. 2013) following the reflect array design methodology. For the proposed super-cell, retrodirectivity is achieved for a single angle of incidence at a subwavelength periodicity. The reciprocal property of the design obtained validates the theory of the generalized law of reflection with respect to the well-known grating equation (Loewen 1983).

Considering the applicative potential of such retroreflective metasurfaces, the scattering of an EM wave in a retrodirective way can be an exceptional feature for the design of radar targets in order to increase their radar cross section (RCS) (Knott 2012) in particular when these targets have low inherent backscattering properties. This property has been achieved in the past using engineered topologies such as corner dihedrals (Guo et al. 2008) and Van Atta array technologies (Guo et al. 2008). Despite the fact that these topologies demonstrate good retrodirective properties at specific range of incident angles, they are inappropriate for implementation in applications where compactness is essential. These applications include automotive radar applications where vulnerable road users (VRU) such as pedestrians and cyclists need to be equipped to enhance their probability of detection. Equivalent topologies presenting retrodirectivity, compactness, as well as ease of implementation should be proposed.

The design of a retrodirective metasurface can be achieved by using the Snell's generalized law of reflection. However, the challenge is to design a retrodirective metasurface operating for multiple incident angles simultaneously. In this chapter, several concepts and designs are provided from a technical aspect using various super-cell superposition and cascading methods to theoretical following the impedance modulation technique and Fano-resonance super-cell designs in order to achieve the desired phenomenon. A cascaded retrodirective metasurface is designed operating at a wide range of incident angles using various super-cell element structures. Following the impedance modulation technique, a more compact design is achieved using a single super-cell design with a high periodicity with respect to

lambda λ, thus operating at multiple oblique incident angles simultaneously for retroreflection. The designs are fabricated and tested in an anechoic chamber to validate their performances in terms of retroreflection compared to that of the simulation results. A mannequin model is also given to measure its RCS when implementing the metasurface along its side; this is to show the applicative aspect of metasurfaces specially that it is presented in the context of project CYCLOPE, which aims to protect vulnerable road users such as cyclists and pedestrians.

Metasurfaces and Anomalous Wave Control

An electromagnetic wave propagating between two media of different refractive indices usually results in a reflective and refractive waves in both media (Hou-Tong Chen and Yu 2016). However, for a homogeneous media, the wave scattering obeys to the conventional Snell's law of reflection and refraction. This can differ for engineered topologies such as metasurfaces, which are characterized by a non-homogeneous media, thus giving rise to unprecedented physical properties for anomalous control of the reflected and refracted waves. The principle is shown in Fig. 1.

An EM wave is shown propagating through a metasurface: it is reflected and refracted in anomalous directions, in a way which is different if compared to the conventional physical laws. Gradient metasurfaces (Sun et al. 2012), which we will focus on in this chapter, are acknowledged for their capability of controlling of an

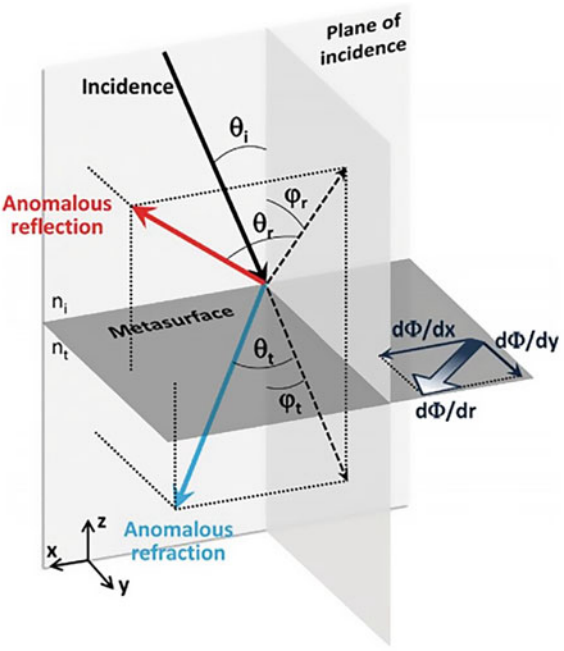

Fig. 1 Concept for a metasurface where an incident plane wave is controlled in arbitrary directions (Hou-Tong Chen and Yu 2016)

EM wave, such as beam guiding and abnormal reflection. The generalized law of reflection and refraction is used in the design of a gradient metasurface. Yu et al. (2011), when contrasted to the traditional Snell's law (Eqs. 1 and 2), an additional term known as the phase gradient is introduced. This corresponds to the well-known grating equation (Loewen 1983).

$$\sin\theta_r - \sin\theta_i = \frac{\lambda_0}{2\pi}\frac{\partial\phi}{\partial x}, \tag{1}$$

$$n_t \sin\theta_t - n_i \sin\theta_i = \frac{\lambda_0}{2\pi}\frac{\partial\phi}{\partial x}, \tag{2}$$

Equation (1) shows the generalized phase law of reflection where θ_r is the angle of reflection, θ_i is the angle of incidence, λ_0 wavelength in free space, and $\partial\phi/\partial x$ is the phase gradient. Equation (2) shows the generalized phase law of refraction where θ_t is the angle of transmission, θ_i is the angle of incidence, λ_0 the free space wavelength, $\partial\phi/\partial x$ is the phase gradient, and n_t and n_i are the refractive indices of media 1 and 2. However, particularly for reflection where transmission is neglected, an electromagnetic wave is reflected to a chosen anomalous direction by establishing a phase gradient between the element scatterers on top of a designed metasurface design.

From the Generalized Phase Law of Reflection to Retrodirectivity

For metasurfaces, retrodirectivity is a desirable characteristic. It is characterized as bouncing a wave back in the same direction it came from (Asadchy et al. 2017). The standard Snell's equation of reflection can readily be used to determine retrodirectivity at normal incidence from a metallic surface. Figure 2 shows that after introducing a phase gradient δ between unit cells, the wave is reflected back from a virtual surface where the initial reference is tilted by an angle θ_i reflecting back in the same direction as the conventional property for a metallic flat plate.

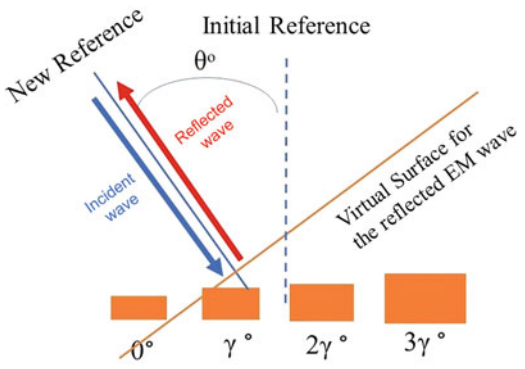

Fig. 2 The concept of a wave reflecting in the same direction of incidence from an engineered gradient retrodirective metasurface

For the design of a gradient metasurface for wave control, we use the generalized phase law of reflection defined in Eq. (1). However, for the case of retroreflection, the angle of incidence $\theta_i = -\theta_r$ where θ_r is the reflection angle. The integration (from 0 to 2π) of the generalized law of reflection is used to control the complete wavefront of the propagating wave. The gradient super-cell's periodicity determines whether or not the incoming wave is fully controlled in the desired direction. Therefore, for retrodirectivity, the periodicity L_x is calculated as follows:

$$L_x = \frac{\lambda}{2 \sin \theta_i} \tag{3}$$

L_x in Eq. (3) defines directly the principle to design a retrodirective metasurface where the efficiency of retroreflection for the metasurface design is proportional to the periodicity L_x and the wavelength λ, with θ_i varying according to the desired retroreflection incident angle.

Design of a Retrodirective Super-Cell: Principle of Operation

In this section, we show briefly and in details the design procedure for a retrodirective super-cell. We have published this work in the Journal of Applied Physics entitled "Multi Angle Retrodirective Cascaded Metasurface" (Kalaagi and Seetharamdoo 2019). The same procedure and principle is used for all the designs which will follow but only by changing the desired angle of incidence θ_i. As illustrated in Fig. 3, after obtaining periodicity for a retrodirective super-cell L_x, N unit cells are derived from the super-cell with a phase gradient $d\phi/dx$ between the scatterers using the generalized law of reflection.

Thus using the generalized law of reflection and after achieving the periodicity for a retrodirective super-cell L_x, the super-cell is divided into N unit cells with a phase

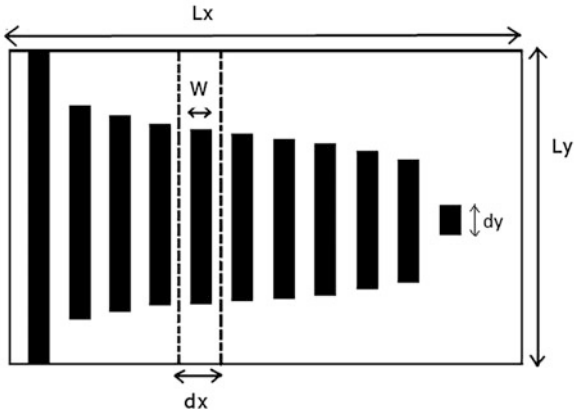

Fig. 3 Retrodirective super-cell with subwavelength patch strip-lines of dimensions L_x and L_y

gradient $d\phi/dx$ between the scatterers, as shown in Fig. 3. In this case, we give a first example of a super-cell design.

Because of their subwavelength structure and low periodicity between the scatterers, as a unit cell, the strip-lines metallic patches were chosen. As a result, a magnetic connection is formed, allowing for high efficiency performance (Sun et al. 2012). To manipulate the wavefront, the unit cells are calibrated to generate a gradual phase shift across the entire range of the phase of the reflection coefficient. The phase gradient (Chen et al. (2016) is given in Eq. (4).

$$\phi_{N+1} - \phi_N = 2K_0 d_x \sin\theta_i \tag{4}$$

d_x is the periodicity of the unit cell which is given as $d_x = L_x/N$ and K_0 the wave number. For retrodirectivity, Eq. (3) is put into Eq. (4) to provide a retrodirective super-cell's phase gradient, as shown in Eq. (5).

$$\phi_{N+1} - \phi_N = \frac{2\pi}{N} \tag{5}$$

The number of subwavelength unit-cells N directly impacts the phase gradient, as given in Eq. (5). The reflection phase coefficient and Eq. (5) are used to calculate the size of the unit cells. An example of a super-cell design is given for $\theta_i = 30^\circ$. For a unit cell of periodicity d_x and length L_y, the reflection phase coefficient is calculated using Floquet harmonic boundaries, with the width w and length d_y for a single strip-line, as shown in Fig. 4. An example in this case is shown for a design with $\theta = 30^\circ$ and $\lambda =$ 20 mm (frequency 14.7 GHz) *(Note that the frequency chosen at 14.7 GHz is just for better demonstration and ease of fabrication, where the concept can be transposed to any frequency depending on the concept),* $L_y = \lambda/2$ with $N = 12$, $w = 0.71$ mm and. The metallic strip-lines are designed on a permittivity $\varepsilon = 2.2$, $\tan(\delta) = 1 \times 10^{-3}$, and thickness 1.57 mm grounded substrate.

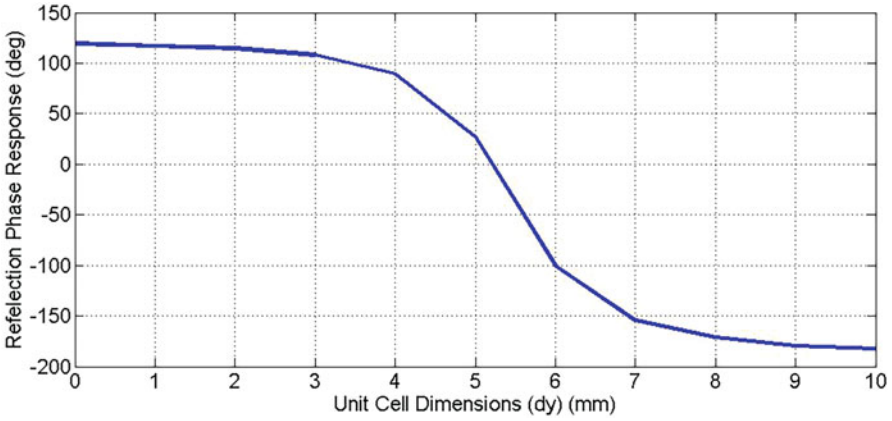

Fig. 4 At a frequency of 14.7 GHz, the phase of the reflection coefficient ϕ with regard to the dimensions of the strip-lines d_y

Table 1 Reflection phases with regard to the strip-lines' dimensions d_y and a phase gradient of 30°

Phase coefficient (degrees)	Dimensions d_y (mm)
−182.6812	10.2
−152.6812	7.1
−122.6812	6.5
−92.6812	6.06
−62.6812	5.82
−32.6812	5.6
−2.6812	5.34
27.3188	5.1
57.3188	4.6
87.3188	4.12
117.3188	1
147.3188	0

Using the reflection phase coefficient curve in Fig. 4 and Eq. (5) as a guide, the phase gradient $d\phi/dx$ is chosen as 30°. Table 1 shows the dimensions of the strip-line unit cells. Eleven strip-lines are depicted, with the last strip-line's phase equal to zero.

Condition of Lossless Reflection

A high efficient retrodirective wave is verified if the lossless reflection condition is confirmed. It has been shown (Asadchy et al. 2016) that in order to achieve a high efficient reflection, the Poynting vector's normal component should be less than or equal to zero $P_n \leq 0$ which is attained when the reflected electric field E_r is equal to that of the incident electric field E_i ($E_r = E_i$). P_n is defined as follows:

$$\mathbf{P}_n = \frac{E_i^2}{2\eta_1}(\cos\theta_r - \cos\theta_i)[1 + \cos\phi_r(z)] \tag{6}$$

where ϕ_r is the reflection phase. P_n has been calculated for the design at 30°, as shown in Fig. 5.

The result in Fig. 5 shows that P_n is less than zero for the given band of incident angles, and this signifies that the condition of Eq. (6) is validated and the retro-reflector is lossless.

Designs for Retrodirective Metasurfaces Corresponding to Different Incidence Angles

Using the provided design technique and general operating concept for a super-cell design at 30° that we have just defined, various metasurfaces with different periodicities are developed at 25°, 20°, 15°, and 10° using the same approach and principle.

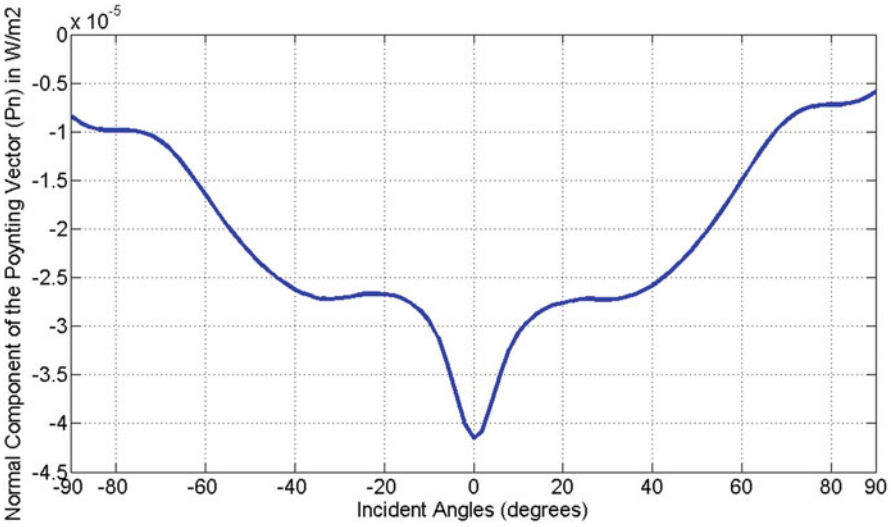

Fig. 5 Poynting vector normal component with respect to θ_i for the metasurface design at 30° angle of incidence

To determine the retrodirectivity performances of each metasurface from the angular aspect, we calculate the monostatic radar cross section (RCS) of each design. The monostatic RCS is defined as the wave's energy which is sent and redirected back to the same angle of incidence, which is basically a similar case to that which is used in WPT as a pilot signal for tracking and localization (Aziz et al. 2019). Furthermore, the efficiency of each metasurface in terms of retrodirectivity is determined by comparing it to that of a flat metallic plate.

Retrodirective Metasurface at 30° Monostatic RCS Performance

An array periodic structure of 7 × 9 super-cells is given with a Teflon substrate of relative permittivity $\varepsilon = 2.2$, as illustrated in Fig. 6, based on the design of a retrodirective super-cell and the basic principle of operation at 14.7 GHz. Table 1 shows the dimensions of the strip-lines. Using Eq. (3), the periodicity L_x is equal to λ for $\theta = 30°$. Figure 7 shows the monostatic RCS for the design, compared to a normal metallic flat plate.

Figure 7 shows that at the given angles, as well as at $-30°$ and normal incidence, there is a large value of retrodirectivity. The metasurface's symmetrical reaction is expected, owing to the equal response in both directions of the phase of the reflection coefficient from 117° to −182.68° and from −182.68° to 117°, as shown in Fig. 4.

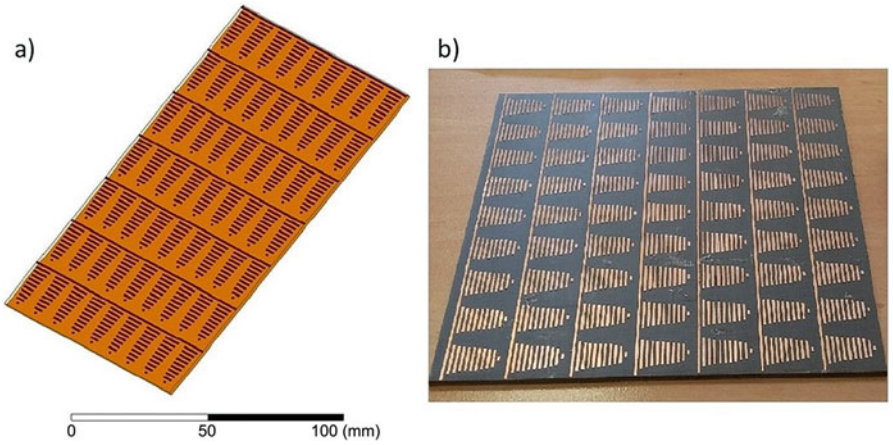

Fig. 6 (**a**) Retrodirective metasurface designed at 30° using Ansys HFSS. (**b**) Metasurface fabricated prototype of 7 × 9 super-cells with a Teflon substrate of dimensions 14.3 cm × 9.2 cm and thickness 1.57 mm

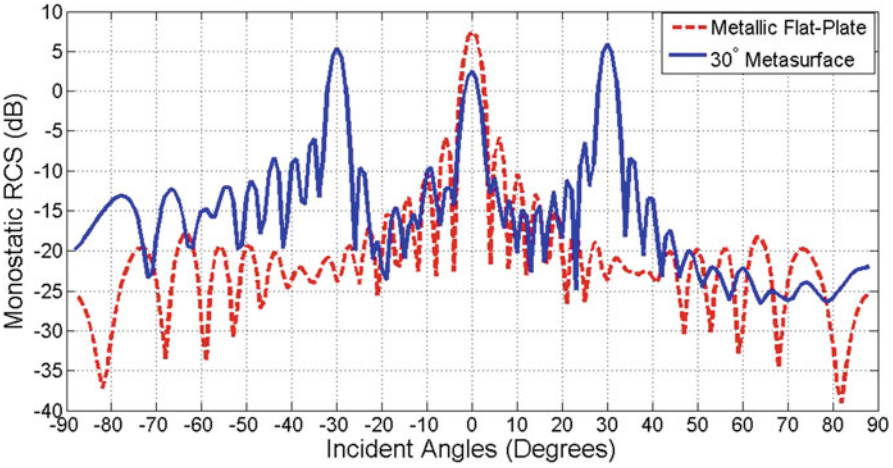

Fig. 7 In compared to a flat metallic plate of comparable size, the monostatic RCS (dB) for a 30° retrodirective metasurface design of 7 × 9 super-cells and dimensions 14.3 cm × 9.2 cm

Retrodirective Performances for Metasurface Designs at 25° and 20°

In comparison to the design at 30°, a similar design approach is used for the designs at 25° and 20°. The periodicity L_x is determined to be 1.18λ and 1.462λ at 25° and 20°, respectively. The dimensions of the unit cells for 25° and 20° are shown in Tables 2 and 3, respectively. Figures 8 and 9 show the monostatic RCS of the designs, respectively.

Table 2 Reflection phases and the strip-lines d_y with a specific phase gradient ϕ_N equal to 30° for the retrodirective super-cell at 25°

Phase coefficient (degrees)	Dimensions d_y (mm)
−182.6812	10.2
−152.6812	7.12
−122.6812	6.618
−92.6812	6.12
−62.6812	5.87
−32.6812	5.63
−2.6812	5.38
27.3188	5.13
57.3188	4.63
87.3188	4.12
117.3188	1.56
147.3188	0

Table 3 Reflection phases and the dimensions of the strip-lines d_y with a given phase gradient ϕ_N equal to 30° for the retrodirective super-cell design at 20°

Phase coefficient (degrees)	Dimensions d_y (mm)
−182.6812	10.2
−152.6812	7.63
−122.6812	6.78
−92.6812	6.25
−62.6812	5.93
−32.6812	5.68
−2.6812	5.43
27.3188	5.17
57.3188	4.7
87.3188	4.16
117.3188	2
147.3188	0

In Fig. 8, a high value of retroreflection is given at 25° and at −25° as a result of the metasurface design's symmetrical behavior, a loss at normal incidence on the other hand for the 30° design, is due to the periodicity $L_x > \lambda$. As the periodicity rises over λ, higher orders of diffraction arise (at −58°), as predicted by Floquet-Bloch analysis (Epstein and Eleftheriades 2014). Figure 9 shows the result of retroreflection for the desired metasurface at 20° establishing a significant value at 20° and − 20° with high retroreflection at a diffracted order of −44°.

Performances of Retrodirective Metasurface Designs at 15° and 10°

In this scenario, where the periodicity Lx is equal to 2λ and 2.88λ at 15° and 10°, respectively, a retrodirective metasurface is designed. The unit cell's dimensions of the given designs are given in Tables 4 and 5. The monostatic RCS of the designs is calculated in Figs. 10 and 11, respectively.

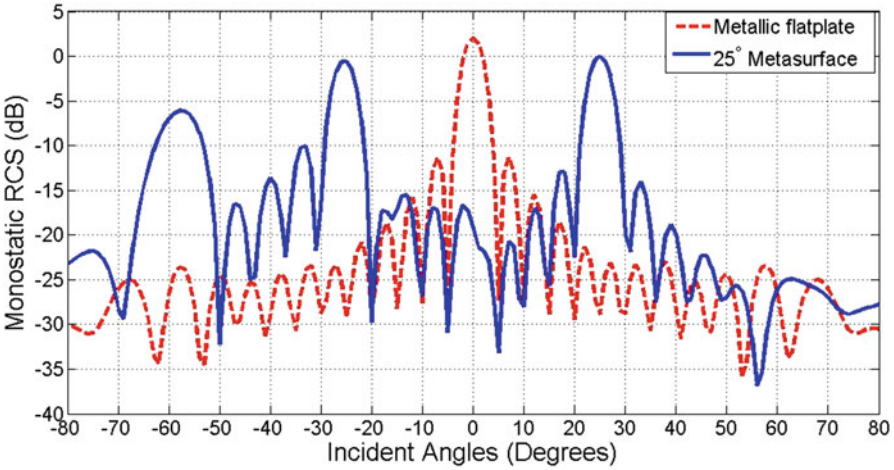

Fig. 8 In comparison to a flat metallic plate of equivalent dimensions, monostatic RCS (dB) for a 25° retrodirective metasurface design of 6 × 5 super-cells and dimensions 6.1 cm × 12 cm

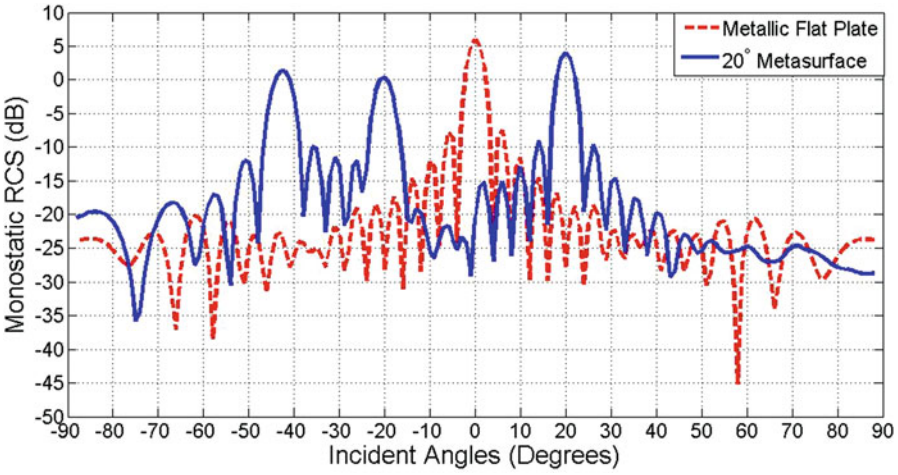

Fig. 9 In comparison to a flat metallic plate of equivalent dimensions, monostatic RCS (dB) for a 20° retrodirective metasurface design of 7 × 5 super-cells and dimensions 7.1 cm × 15 cm

The results in Fig. 10 reveal that monostatic RCS has a substantial value at 15°. Although the symmetrical reaction in the opposite direction is absent at −15°, the phase of the reflection coefficient stays the same when compared to the responses at 30°, 25°, and 20°. At −32° and − 50°, two diffracted orders with a high monostatic RCS value are observed, as seen in Fig. 10. Due to the high periodicity, the symmetrical response in regards of retrodirectivity for the metasurface design is lost. L_x being $\geq 2\lambda$ and the presence of the higher orders of diffraction (Epstein and

Table 4 Reflection phases and strip-line dimensions d_y for the retrodirective super-cell design at 15° with a specified phase gradient ϕ_N equal to 30°

Phase coefficient (degrees)	Dimensions d_y (mm)
−182.6812	10.2
−152.6812	7.8
−122.6812	6.88
−92.6812	6.39
−62.6812	6
−32.6812	5.75
−2.6812	5.5
27.3188	5.24
57.3188	4.83
87.3188	4.29
117.3188	2.22
147.3188	0

Table 5 Reflection phases and the dimensions of the strip-lines d_y with a chosen phase gradient ϕ_N equal to 30° for the retrodirective super-cell design at 10°

Phase coefficient (degrees)	Dimensions d_y (mm)
−182.6812	10.2
−152.6812	7.83
−122.6812	6.9
−92.6812	6.46
−62.6812	6.05
−32.6812	5.8
−2.6812	5.54
27.3188	5.29
57.3188	4.93
87.3188	4.33
117.3188	2.76
147.3188	0

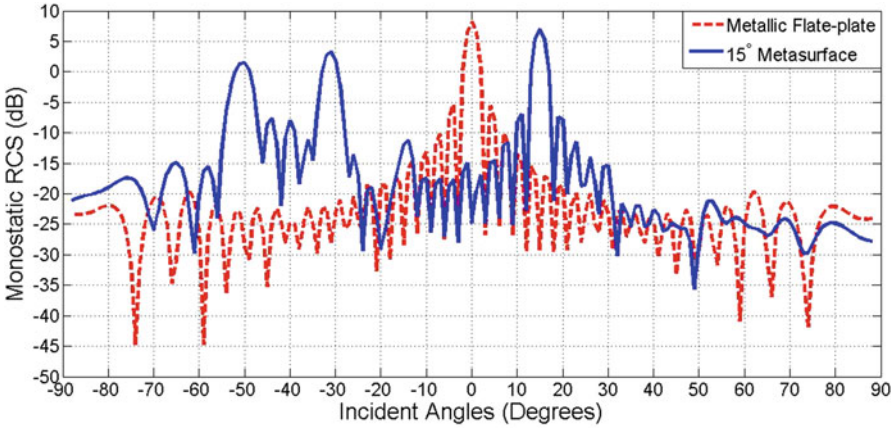

Fig. 10 In comparison to a metallic flat plate of equivalent size, monostatic RCS (dB) for a 15° retrodirective metasurface of 7 × 5 super-cells and dimensions 7.14 cm × 19.7 cm

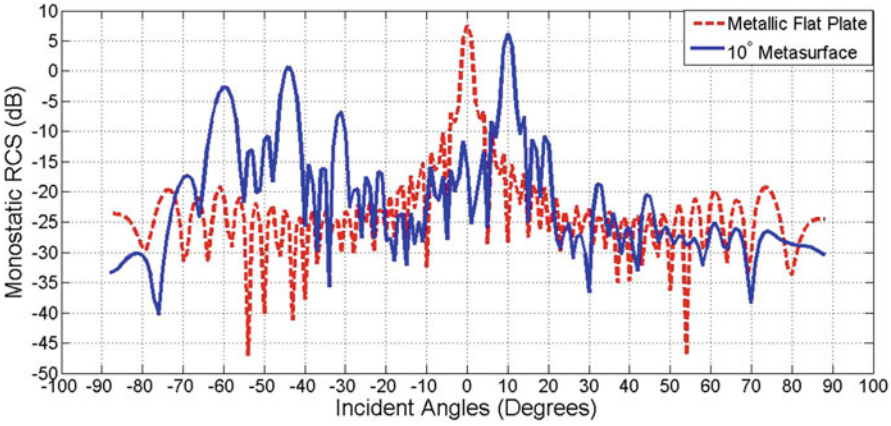

Fig. 11 In comparison to a metallic flat plate of equivalent size, monostatic RCS (dB) for a 10° retrodirective metasurface of 6 × 4 super-cells and dimensions 6.12 cm × 23.5 cm

Eleftheriades 2014). The monostatic RCS result for a 10° retrodirective metasurface with a periodicity $L_x = 2.88\lambda$ is shown in Fig. 11. With no retrodirectivity in the symmetrical direction at −10°, a high value of monostatic RCS is produced at the required angle as well as the two diffracted orders −44° and − 60°.

Symmetrical Responses at −15° and − 10° for Retrodirective Metasurface Designs of Periodicities Greater than 2λ

The retrodirective metasurface for a periodicity $\geq 2\lambda$ signifies a loss in the symmetrical behavior of the metasurface. As a result, for the metasurfaces at 15° and 10°, an alternate design structure is provided by constructing a metasurface with two opposite orientations, as illustrated in Fig. 12. Figures 13 and 14 illustrate the study results of the suggested design's monostatic RCS at 15° and 10°, respectively.

Figures 13 and 14 show a comparison of the suggested metasurface designs to a metallic flat plate at 10° and 15°, respectively. Figures 13 and 14 show the results of the retroreflection; results indicate the regain of the loss in the reciprocal directions utilizing the proposed structural design at −10° and − 15° for the chosen angles of incidence with a moderate efficiency compared to the metallic flat plate, but the loss of the metasurface's symmetrical behavior is returned.

Measurements and Experimental Verification of a 30° Retrodirective Metasurface Prototype

In order to validate the numerical data of a 30° retrodirective metasurface, a prototype is fabricated which is shown in Fig. 6b with 7 × 9 retrodirective

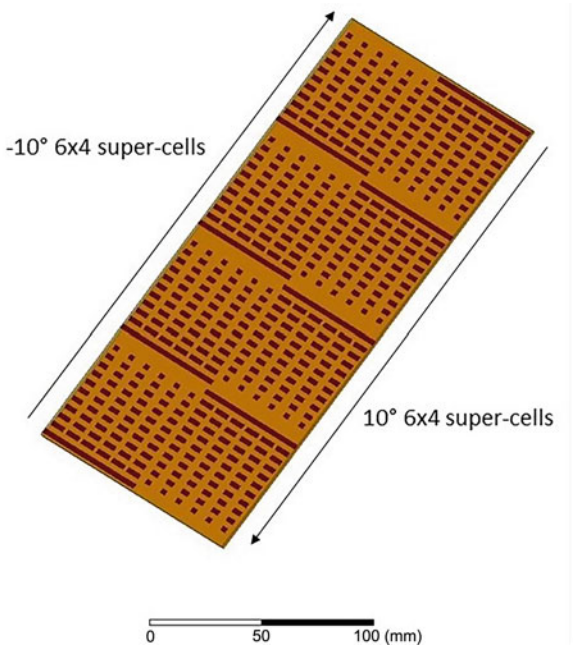

Fig. 12 A retrodirective metasurface in both symmetrical directions at 10°. The design has been carried on using Ansys HFSS

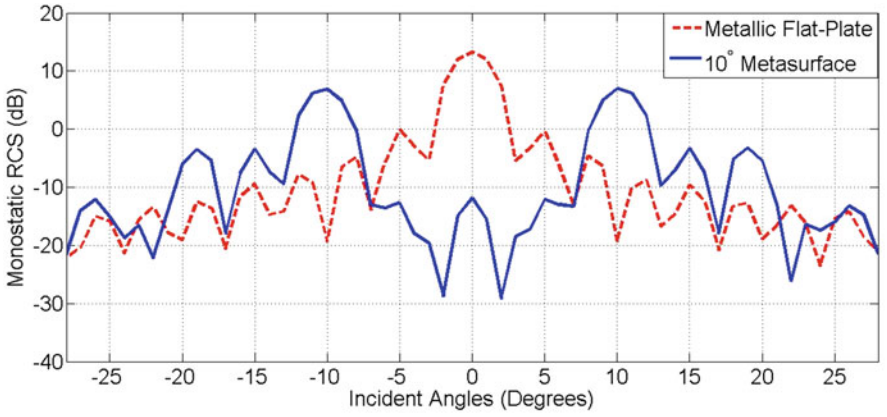

Fig. 13 Monostatic RCS of the 10° retrodirective metasurface given structure design with 12 × 4 super-cells (Fig. 12) of dimensions 12.224 cm × 23.5 cm in comparison to that of a flat metallic plate of comparable dimensions

super-cells of dimensions 14.3 cm × 9.2 cm, with a 1.6 mm thick Teflon grounded substrate of relative permittivity $\varepsilon = 2.2$. The experimental measurements were done in an anechoic chamber at an operating frequency of 14.7 GHz. Figure 16 shows the simulated data compared to the experimental results.

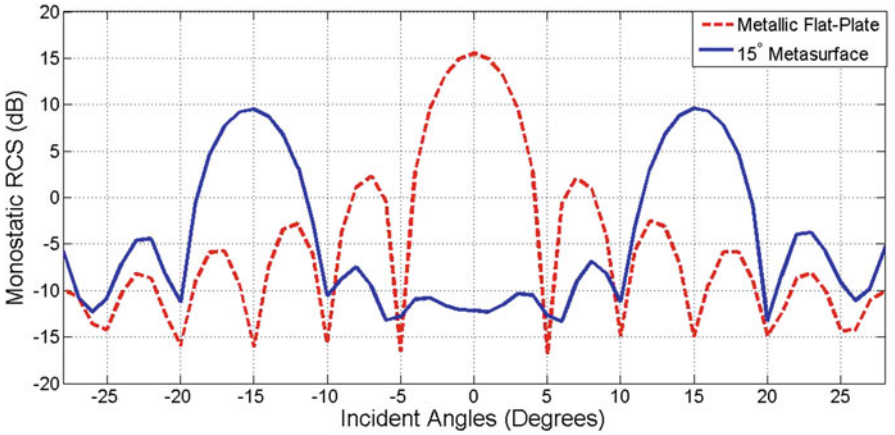

Fig. 14 Monostatic RCS of the 15° retrodirective metasurface given structure design with 14 × 5 super-cells of dimensions 14.28 cm × 19.7 cm in comparison to that of a flat metallic plate of comparable dimensions

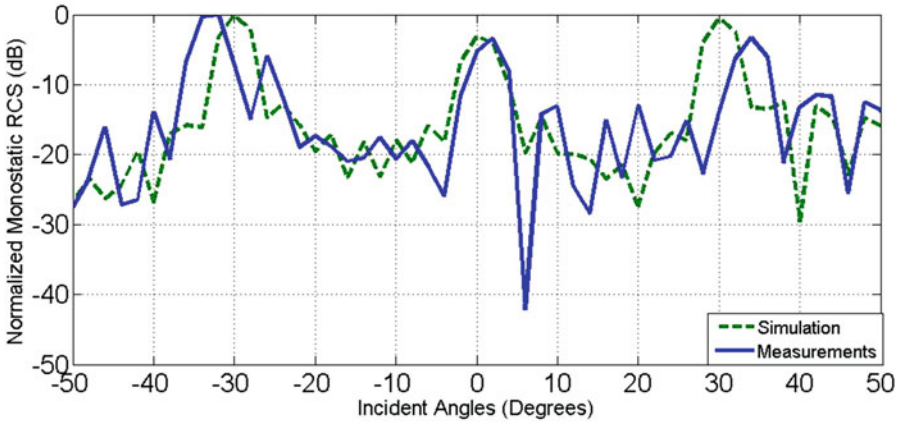

Fig. 15 Normalized monostatic radar cross section of the experimental data in comparison to the simulation analysis for a 30° retrodirective metasurface design

Figure 15 shows that the measurement findings correspond to the numerical data and are in agreement with it.

Design of Dual Polarized Retrodirective Metasurfaces

In some applications, dual polarization can be essential. In this section, we introduce the design of dual polarized retrodirective metasurfaces following the same design procedure and principle of operation explained before, the only change to be in the unit cell

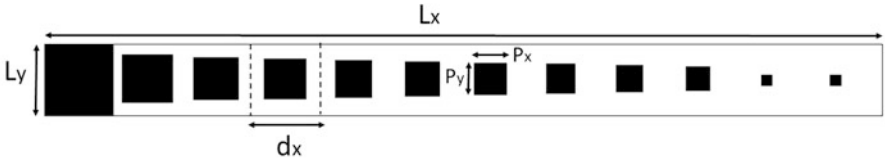

Fig. 16 Super-cell of dimensions L_x and L_y with subwavelength patch unit cell scatterers

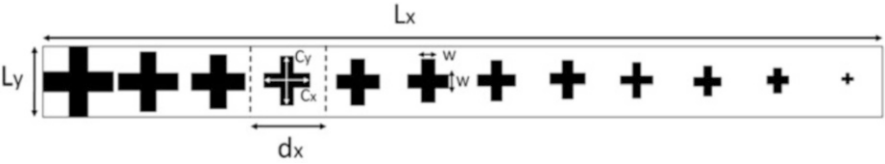

Fig. 17 Super-cell of dimensions L_x and L_y with subwavelength cross unit cell scatterers

element chosen. This section has been presented at the 2018 IEEE Radio and Antenna Days of the Indian Ocean (RADIO) Conference (Kalaagi and Seetharamdoo 2018).

A design for an incident angle at 5° is given in this section for demonstration. L_x is 5.76λ in this case, $N = 12$ unit cells at an operating frequency of 14.7 GHz, and a grounded substrate with permittivity $\varepsilon = 2.2$. The retrodirective super-cell designs are given in Figs. 16 and 17. Figure 16 displays the retrodirective patch structure with two degrees of freedom P_y and P_x, whereas in Fig. 17, the retrodirective cross is designed having three degrees of freedom C_y, C_x, and w which are calculated using the reflection phase coefficient and the phase gradient in (5.5). L_x and L_y are the dimensions of the retrodirective super-cell and d_x for that of the subwavelength unit cell in both cases.

Numerical Analysis for the Retrodirective Dual-Polarized Metasurfaces

A cross structure and patch metasurface of 4×3 super-cells are given, as shown in Figs. 18 and 19, respectively.

The monostatic RCS for both metasurfaces is calculated for an incoming wave at 14.7GHz. The results in Figs. 20 and 21 display the normalized monostatic RCS for the cross and patch metasurfaces, respectively.

The results in Figs. 20 and 21 show a significant value of retroreflection at the chosen angle 5° for both polarization TM and TE; this indicates that this results in tracking enhancement for both polarizations at oblique incidence. Measurements have been carried out for a cross structure metasurface in an anechoic chamber at 14.7 GHz. The prototype is shown in Fig. 22.

Measurements for both TM and TE polarizations have been carried out for the cross prototype metasurface. The results of both TE and TM measurements compared to the simulation results are given in Figs. 23 and 24.

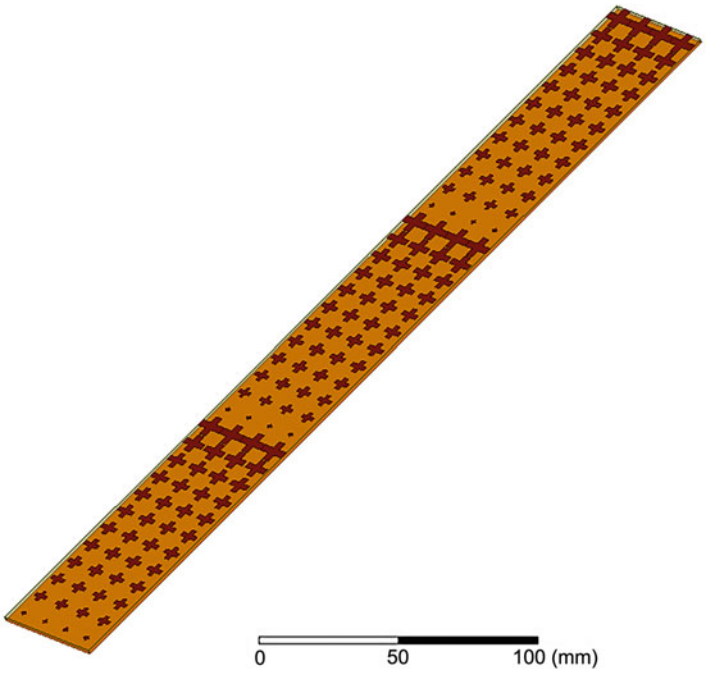

Fig. 18 Cross structure metasurface of 4 × 3 super-cells

The results in both Figs. 23 and 24 for TE and TM measurements of the cross metasurface agree almost perfectly with the simulation results, thus validating the results.

Design of a Multi-angle Cascaded Retrodirective Metasurface Based on Superposition

The aim of our study is to design retrodirective metasurfaces which can operate at wide range of incident angles. This is to enhance the potential of tracking and localization from the angular aspect of a target, especially for extreme oblique incident angles. The cascaded multiple incident angle retrodirective metasurface (C-MIARM) is designed based on superposing different designed metasurfaces at various angles. In this case, the metasurface is designed by superposing designs at 30°, 25°, 20°, 15°, and 10°, using the symmetrical metasurface designs at 15° and 10° as displayed in Fig. 12. The C-MIARM design is shown in Fig. 25.

The overall C-MIARM's dimensions are 8.163 cm × 56.23 cm. In order to establish the overall retroreflective performance of the proposed metasurface, the normalized monostatic RCS with regard to the metasurfaces at various angles is given, as illustrated in Fig. 26.

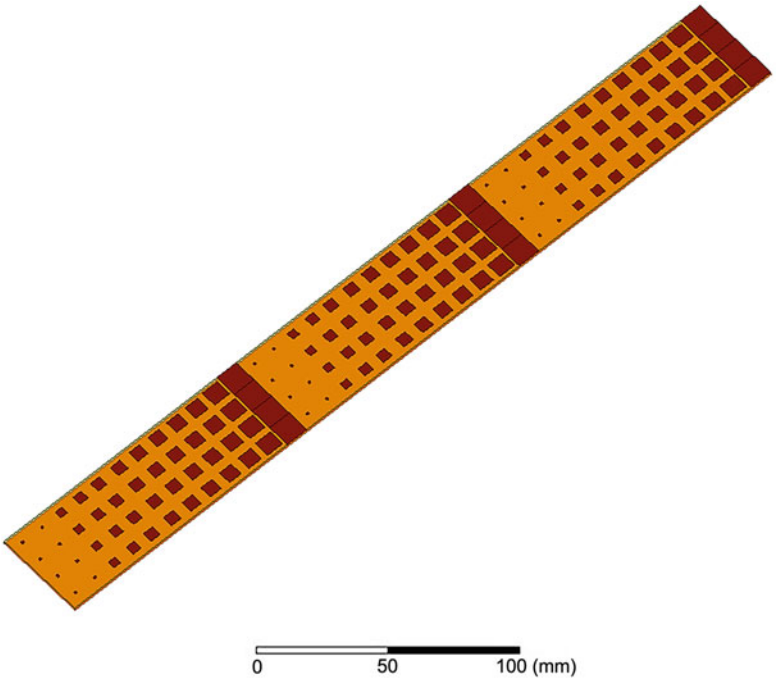

Fig. 19 Patch structure metasurface of 4 × 3 super-cells

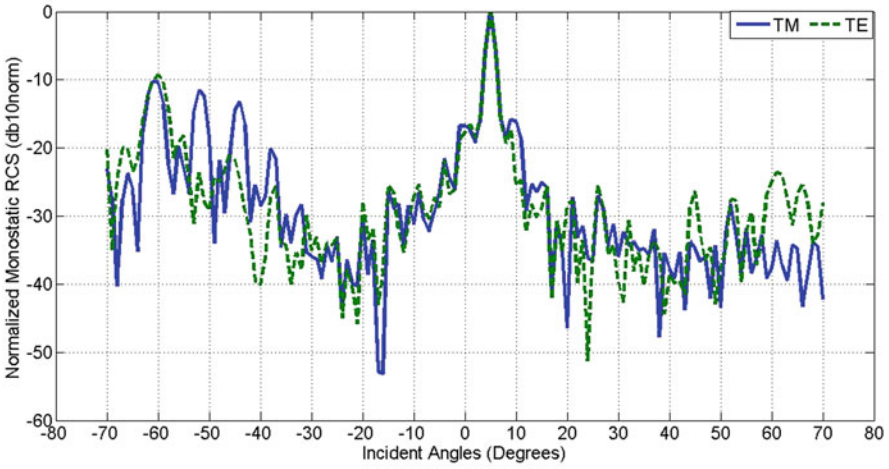

Fig. 20 Simulation results of the normalized monostatic RCS in dB10norm for a cross metasurface structure at TM and TE polarizations using ANsys HFSS

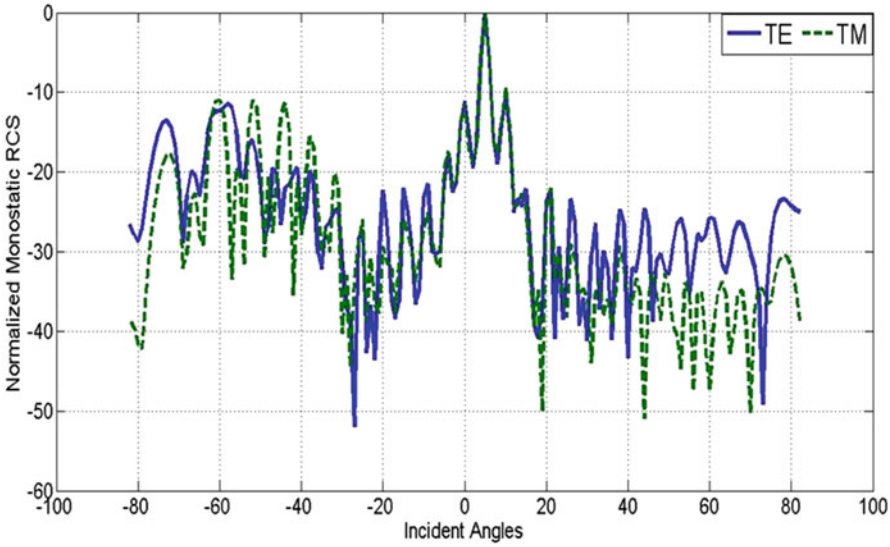

Fig. 21 Simulation results of the normalized monostatic RCS in dB10norm for a patch metasurface structure at TM and TE polarizations using ANsys HFSS

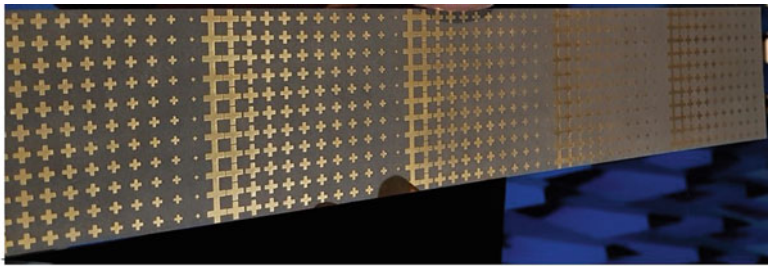

Fig. 22 Prototype of a cross metasurface structure for measurements carried on in an anechoic chamber

In the regions of $-30°$ to $-10°$ and $10°$ to $30°$, Fig. 26 shows a wide range of retrodirectivity with regard to various angles of incidence. A low value is obtained in the range of $-10°$ to $0°$, which is related to the lack of a metasurface design at $5°$ for the numerical analysis' simplicity, but a moderate value is obtained between $0°$ and $10°$.

C-MIARM Retroreflection Performance Compared to that of a Conventional Corner Dihedral

Various engineered topologies in terms of retrodirectivity are presented in the literature such as the conventional corner dihedral (Srour et al. (2018). For performance and efficiency comparison purposes, the C-MIARM performance in terms retrodirectivity is compared to a conventional corner dihedral, as given in Fig. 27.

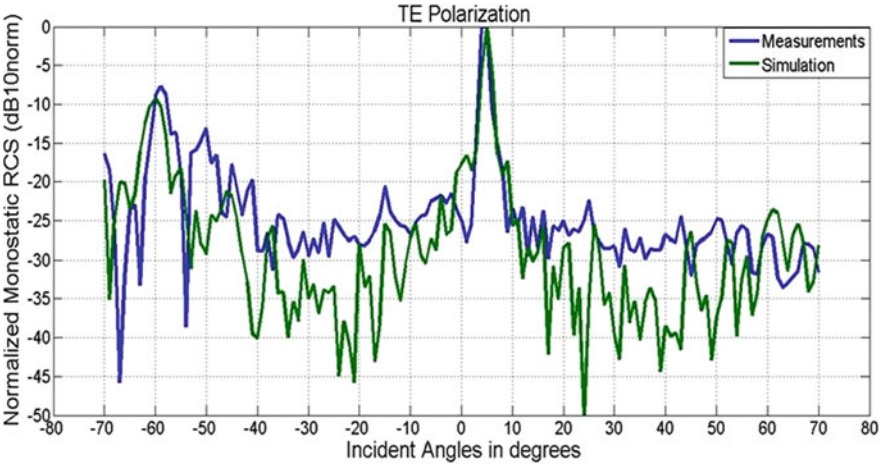

Fig. 23 Normalized monostatic RCS measurements compared to that of the simulation results for TE polarization for the cross metasurface at 14.7 GHz

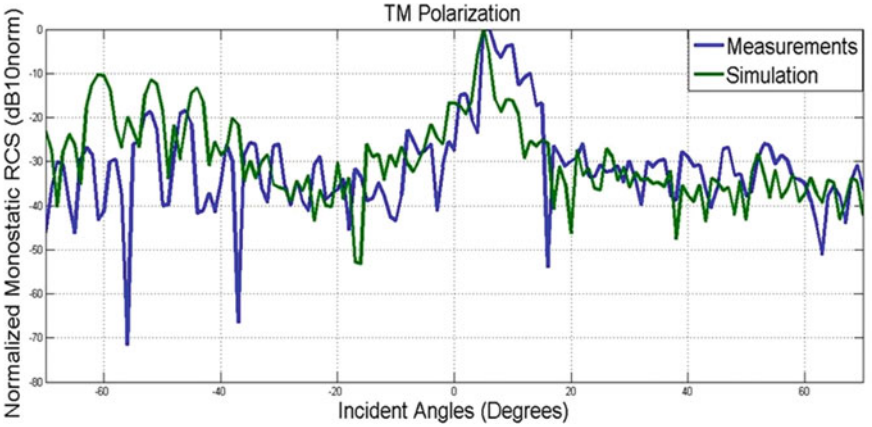

Fig. 24 Normalized monostatic RCS measurements compared to that of the simulation results for TM polarization for the cross metasurface at 14.7 GHz

When comparing the C-MIARM to the corner dihedral of dimensions 8 cm × 15.5 cm × 7.75 cm, the results in Fig. 27 show a 50 dB gain in retroreflection in the incident angle range between −30° to −20° and 20° to 30°, and the corner dihedral, on the other hand, is confined to a range of −20° to 20°, but the C-MIARM performs well in a range of −30° to −10° and 10° to 30°. By fixing a value in terms of monostatic RCS to 5 dB, the C-MIARM was capable of achieving this value with the dimensions 8.163 cm × 56.23 cm in comparison to that of a three-dimensional topology such as the corner dihedral with dimensions 8 cm × 15.5 cm × 7.75 cm. The 2D C-MIARM's structure allows it to be used in a variety of applications where

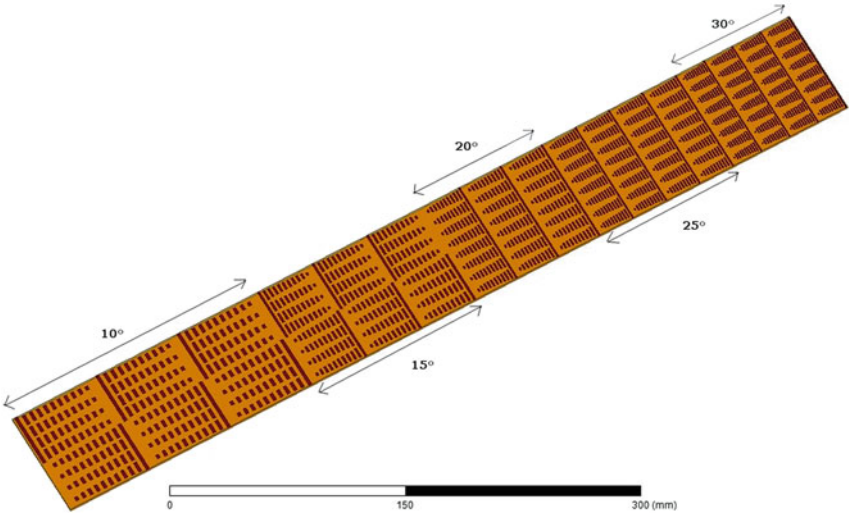

Fig. 25 The C-MIARM design superposing 4 × 8 of the metasurface super-cell designs at 30°, 25°, and 20° as well as the 15° and 10° metasurfaces using the symmetrical design, with an overall dimension of 8.163 cm × 56.23 cm and substrate $\varepsilon = 2.2$ and thickness 1.57 mm

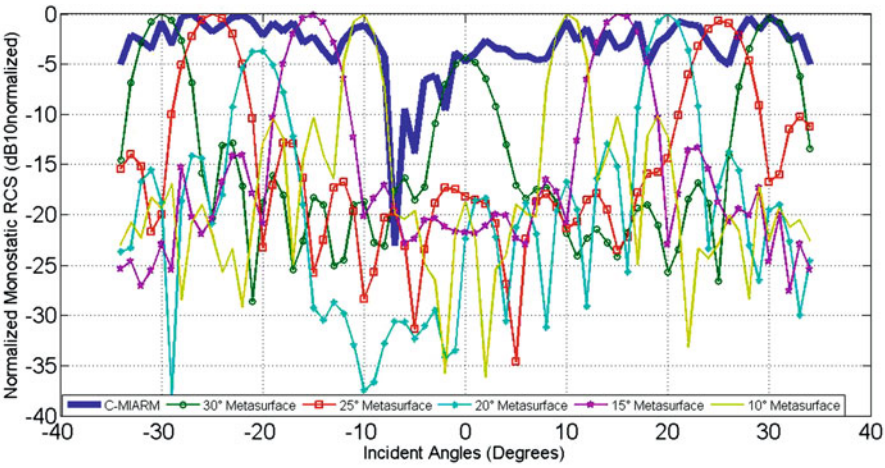

Fig. 26 Analytical studies of the C-MIARM's normalized monostatic RCS in comparison to the retroreflection performances of metasurfaces assembled at 30°, 25°, 20°, 15°, and 10°

exceptional incident angles must be addressed, and it can be a complementing solution for such topologies.

The corner dihedral is widely recognized for operating over a wide frequency range. However, at a narrow-band frequency of 14.7 GHz and a tolerance of 150 MHz over the operational frequency, the C-MIARM is sufficient. For applications such as vehicle radar systems, this functionality can be quite useful and

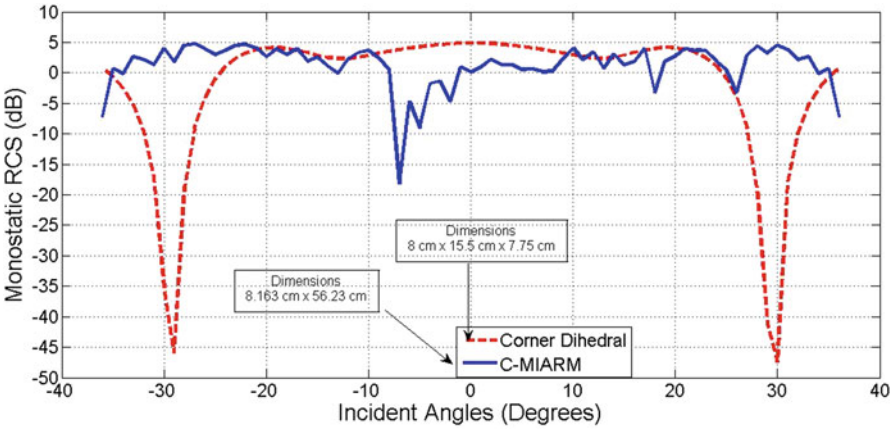

Fig. 27 Analytical results using Ansys HFSS of the monostatic RCS to determine the retroreflection performances of the C-MIARM compared to that of a conventional corner dihedral with dimensions 8.163 cm × 56.23 cm and 8 cm × 15.5 cm × 7.75 cm, respectively

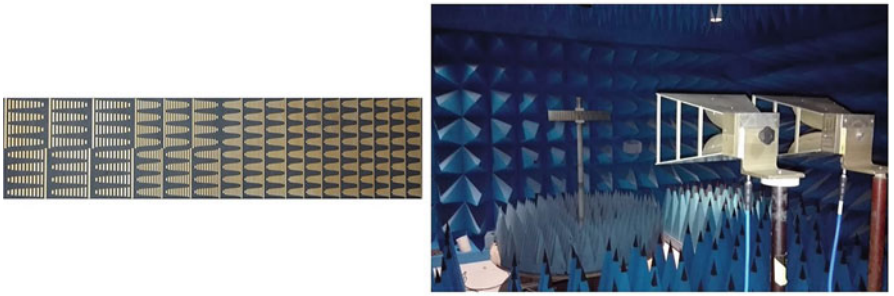

Fig. 28 (**a**) The C-MIARM prototype of dimensions 8.163 cm × 56.23 cm. (**b**) Measurements using two Vivaldi antennas operating at 14.7 GHz, one transmitter and one receiver situated at the same point 3 m away from the metasurface, which are implemented on a turntable to calculate the monostatic RCS at a wide range of angles

appealing. For applications that need a wide range of frequencies, several unit cell structures and designs are used (Chen et al. 2018) and can be taken into consideration for a stable distribution of the reflection phase at a wide band of frequencies.

Measurements and Experimental Validation of the Analytical Results for the C-MIARM Design

To validate the performances and analytical results of the C-MIARM design, as shown in Fig. 28a, a prototype is developed and tested. The prototype's measurements are 8.163 cm × 56.23 cm. The tests are conducted in an anechoic chamber with two Vivaldi antennas (transmitter and receiver) operating at 14.7 GHz. Both

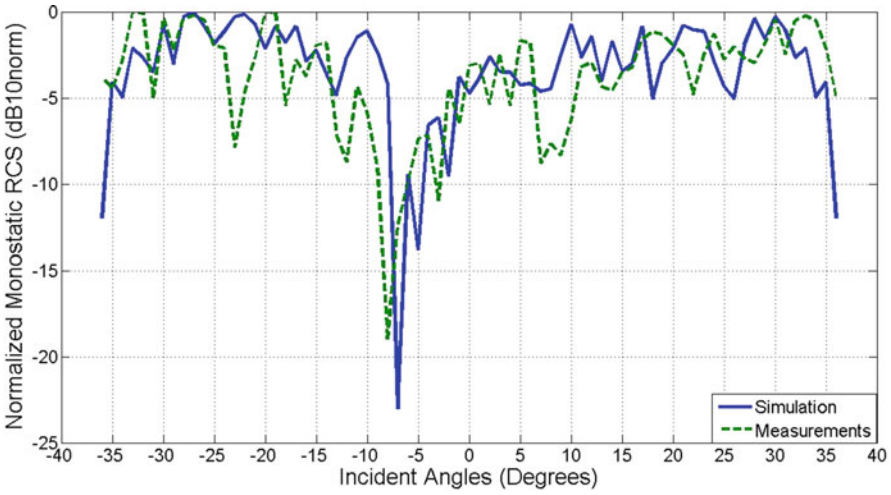

Fig. 29 Comparison of the analytical results and the experimental measurements of the C-MIARM design at the operating frequency of 14.7 GHz

antennas are situated at the same point 3 m away from the metasurface, which is implemented on a turntable to calculate the monostatic RCS at a wide range of angles, as shown in Fig. 28b. The measurements are determined and compared to the analytical results, as shown in Fig. 29.

The test results in Fig. 29 are reported to follow and match those of the simulated results, with some losses at 8° and − 23° due to mechanical imperfections, as well as the coupling and phase shift between the two Vivaldi antennas.

The C-MIARM has shown remarkable results in terms of retroreflection at a wide range of incident angles. It can be a complement solution for tracking and localization enhancement where it can be for applications such as RCS enhancement and wireless power transfer for IOT devices.

From Nonreciprocal to Reciprocal Multi-angle Retrodirective Metasurfaces: High Super-Cell Periodicity Designs Using Impedance Modulation

In this section, we introduce a different technique to design multi-angle retrodirective metasurfaces based on impedance modulation. This work has been published in the journal *Applied Physics A*, Springer entitled "Retrodirective metasurfaces from non-reciprocal to reciprocal using impedance modulation for high-super-cell-periodicity designs" (Kalaagi and Seetharamdoo 2020). Retroreflection is well known to be obtained at a chosen angle of incidence and in the reciprocal or opposite direction, as given in Eq. (3) (Sun et al. 2012; Doumanis et al. 2013; Shen et al. 2018). In the previous sections, we discussed the construction of retrodirective metasurfaces at various angles of incidence at various super-cell periodicity designs (Kalaagi and

Seetharamdoo 2019), and we proved that for retrodirective metasurfaces with sub-wavelength super-cell periodicities L_x, the metasurface performs retroreflection in the desired given direction and the opposite reciprocal direction as well as at normal incidence (Sun et al. 2012; Kalaagi and Seetharamdoo 2019; Doumanis et al. 2013; Shen et al. 2018). However, according to Floquet-Bloch harmonic analysis (Epstein and Eleftheriades 2014; Asadchy et al. 2017), higher orders of diffraction for the super-cell design can be achieved by increasing the periodicity higher than the wavelength λ, resulting in more apertures for wave propagation at varied incident angles. Choosing a design of a super-cell with a periodicity in between λ and 2λ, at normal incidence, the metasurface has no retrodirectivity (Kalaagi and Seetharamdoo 2019), which we recall in Figs. 8 and 9. The metasurface loses its classic reciprocal feature in terms of retroreflection when the frequency exceeds 2λ (Kalaagi and Seetharamdoo 2019). This is because, as seen in Figs. 10 and 11, higher orders of diffraction and apertures exist where the wave can propagate.

Asadchy et al. implemented the concept of the surface impedance modulation (Asadchy et al. 2016) to design multi-channel reflectors (Asadchy et al. 2017), with a specified periodicity of 1.064λ, using the principles of lossless and reciprocal reflectors. For an incident angle of 28°, the given proposition is proven; however, investigating the design of retrodirective metasurfaces that cover lower angles of incidence such as 10° can be significant, where the periodicity is equal to 2.88λ. The goal is to reclaim the classic reciprocal properties of a retrodirective metasurface with a high super-cell periodicity in relation to λ and be able to achieve efficient retroreflection at the higher orders of diffraction. Opposed to what we have introduced in the previous sections, where the concept of cascading has been introduced (Kalaagi and Seetharamdoo 2019) and the design of opposite super-cell designs (Figs. 25 and 12), we would like to optimize the design of a few retrodirective metasurfaces without affecting the performance, where the cascaded metasurface (Kalaagi and Seetharamdoo 2019) can be bulky and of large dimensions for some applications.

This section introduces a retrodirective metasurface design with a high super-cell periodicity of 2.88λ. After that, the surface impedance modulation approach is used to keep the standard reciprocal property at 10° and normal incidence at a high performance with a single super-cell design of unit cell patches. Furthermore, because of the large number of diffraction orders at these periodicities, the retrodirective metasurface can be extended to function at numerous incidence angles at the same time with a proper engineered surface impedance. This can be a great complementary solution for tracking enhancement and of compact dimensions for feasible implementation in different applications.

Relation between Periodicity L_x and Wavelength λ

The performance of retroreflection of the metasurface depends on the relation between the wavelength λ and the periodicity L_x, in which θ_i varies according to the chosen desired angle of incidence for retroreflection, as given in Eq. (3). L_x should be a positive value since it is the periodicity of the super-cell design. Thus,

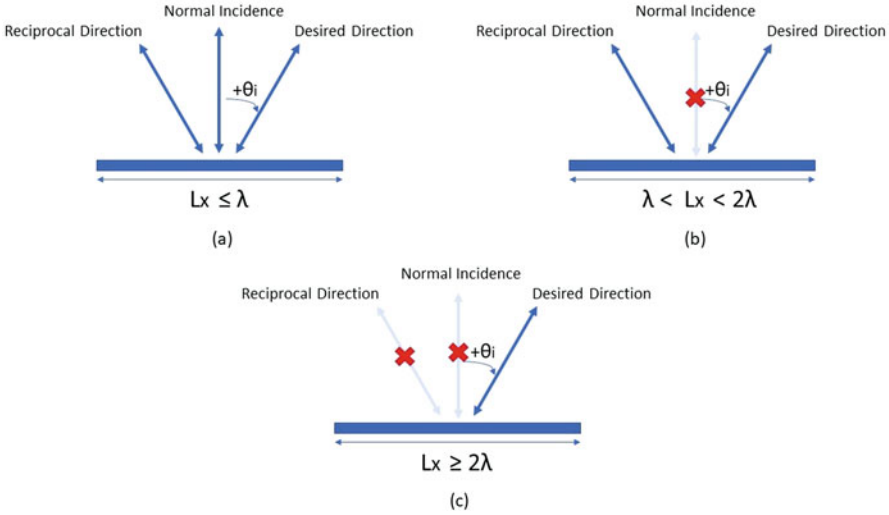

Fig. 30 (**a**) For a periodicity $L_x \leq \lambda$, retrodirectivity is achieved at the chosen angle θ_i, the symmetrical direction, and at normal incidence. (**b**) For a periodicity $\lambda < L_x < 2\lambda$, retrodirectivity is achieved in the chosen given direction θ_i and the symmetrical direction; however, for normal incidence, retroreflection is missing. (**c**) For a periodicity $L_x \geq 2\lambda$, retrodirectivity is achieved just in the chosen direction θ_i; however, the conventional properties of the retroreflective metasurface are absent in the symmetrical and normal incidence directions

in theory, the metasurface design performs retrodirectivity for the chosen angle θ_i and the angle in the symmetrical direction $-\theta_i$ simultaneously. This is proved in preceding work for subwavelength periodicities (Sun et al. 2012; Asadchy et al. 2017; Doumanis et al. 2013; Shen et al. 2018). The performance of the metasurface alters when the periodicity is greater than λ, indicating a direct relationship between the wavelength λ and the periodicity L_x, this concept is explained briefly in Fig. 30.

Giving a periodicity $L_x \leq \lambda$, retrodirectivity can be achieved at the given chosen angle, symmetrical direction, and normal incidence, as shown in Fig. 30a (proved in Fig. 7). On the other hand, for $\lambda < L_x < 2\lambda$, the performance of the metasurface is lost at normal incidence (Kalaagi and Seetharamdoo 2019), as given in Fig. 30b (proved in Figs. 8 and 9). However, by increasing the periodicity above 2λ ($L_x \geq 2\lambda$), the metasurface loses retroreflection for its standard symmetrical characteristic and it is achieved only at the chosen given angle (Kalaagi and Seetharamdoo 2019), as given in Fig. 30c (demonstrated in Figs. 10 and 11). According to Floquet-Bloch analysis (Epstein and Eleftheriades 2014), for a metasurface with a super-cell periodicity greater than λ, open apertures and higher orders of diffraction are present. The existence of open apertures that allow the propagation of the wave in other directions, where the direction of the open apertures or a higher orders of diffractions is derived in the following Eq. (7), might cause the loss of retroreflection in the symmetrical direction and normal incidence.

$$\theta_{rn} = \arcsin \frac{n\lambda}{L_x} \tag{7}$$

where n signifies the number of diffraction orders and θ_{rn} the diffraction orders' directions.

The Design of Retrodirective Metasurfaces with High Periodicities with Respect to λ Using Surface Impedance Modulation

The goal and objective is to keep the retrodirective metasurface's traditional symmetrical characteristic at normal incidence. With a given super-cell periodicity design higher than 2λ, one can achieve efficient retroreflection in the directions of the apertures and higher orders of diffraction by optimizing the metasurface's surface impedance. The design approach for a reciprocal retrodirective metasurface with a periodicity of 2.88λ operating at 14.7 GHz is presented in this section. The surface impedance is tuned to achieve a purely reactive impedance in the nominal super-cell, which is constructed using the generalized law of reflection.

In two processes, the retrodirective metasurface is created. The definition of a configuration based on the generalized law of reflection is the beginning point for designing a gradient metasurface. Figure 31 shows the theoretical super-cell design.

The periodicity L_x and L_y are the length and width of the nominal super-cell, respectively, as displayed in Fig. 31. L_x is obtained according to Eq. (3). The patch dimensions d_y are calculated following the phase of the reflection coefficient, by tuning the patches each with a specific phase gradient. In the second step, the surface impedance of the nominal super-cell design is optimized to be purely reactive. It is proved (Asadchy et al. 2016) that for an efficient lossless retrodirective metasurface, with respect to the stated periodicity, the surface impedance Z_{si} should be purely reactive at each patch element i, with the incident electric field E_i equal to the reflected electric field E_r (Asadchy et al. 2016, 2017). For a given TM polarized wave $Z_{TM} = \eta \cos \theta_i$, Z_{si} is given as:

$$Z_{si} = j \frac{Z_{TM}}{\tan\left(\frac{\varphi_i}{2}\right)} \tag{8}$$

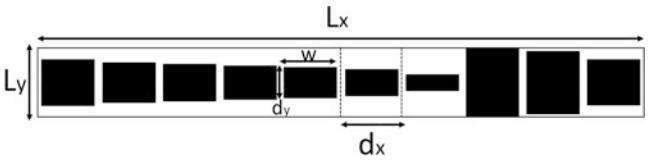

Fig. 31 Patch gradient super-cell (nominal super-cell) design divided in 10 unit cells of dimensions d_y (given in Table 6) following the generalized law of reflection

The surface impedance Z_{si} required for each unit cell element is given by Eq. (8), where ϕ_i is the phase of the reflection coefficient. Post-processing optimization using the Floquet periodic boundary condition of the nominal super-cell patch size is used to get a totally reactive impedance for the given super-cell. This is accomplished by fine-tuning the nominal patch dimensions d_y in order to create the best possible combination with a purely reactive surface impedance. The super-cell design following numerical optimization to a suitable reactive surface impedance is shown in Fig. 32.

The reactive and nominal super-cells are discretized into N unit cells where N = 10 in this example, unit cell dimension $d_x = L_x/N$, and patch widths of $w = d_x - 0.5$ mm. The nominal patches that are tweaked with regard to a certain phase gradient have a dimension of d_y. The phase of the reflection coefficient ϕ_i determined in Fig. 33 is used to calculate the dimensions of the unit cell patches.

Choosing a frequency of 14.7 GHz as a starting point and $\theta_i = 10°$, a phase gradient is introduced between each patch element equals to 40.5° in order to maintain the whole phase distribution, as given in Fig. 33 of the super-cell design. The periodicity $L_x = 2.88\lambda$ and $L_y = \lambda/2$. In Table 6, the dimensions of the nominal unit cell designs are given versus the phase of the reflection coefficient ϕ_i, with a given 1.57 mm thick grounded substrate of permittivity $\varepsilon = 2.2$, $\tan(\delta) = 1 \times 10^{-3}$, as displayed in Fig. 31.

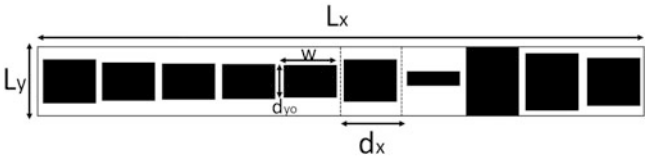

Fig. 32 Reactive super-cell after optimization of the surface impedance by tuning the unit cell dimensions d_y (given in Table 6)

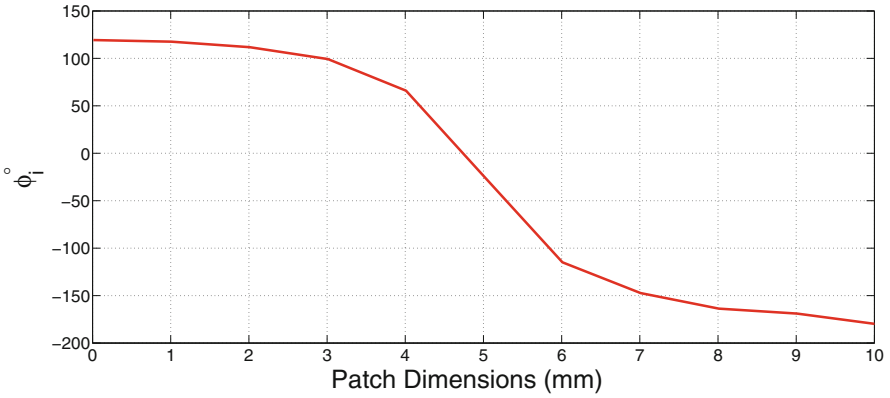

Fig. 33 The phase of the reflection coefficient with respect to the dimensions of the patches d_y in mm at an operating frequency of 14.7 GHz

Table 6 The dimensions of the nominal super-cell unit cell elements design d_y(mm) using the generalized law of reflection alongside the reactive optimized unit cell dimensions d_{yo} versus the phase of the reflection coefficient ϕ_i° calculated in Fig. 33 given a gradient phase equals to 40.5°

ϕ_i°	Nominal dimensions d_y(mm)	Optimized dimensions d_{yo}(mm)
−134.27	6.73	6.38
−93.7	5.88	5.61
−53.2	5.408	5.306
−12.7	4.969	5.102
27.8	4.489	4.898
68.3	3.979	5.915
108.8	2.306	2.04
−210.7	10.2	10.2
−170.2	9.28	8.673
−129.7	6.53	6.73

The nominal super-cell's surface impedance (Fig. 31) is derived using the Floquet periodic boundary condition for a TM polarized wave with $Z_s = 11.5 - j524.25$, at an operating frequency of 14.7 GHz and angle of incidence 10°. The unit cell patches are tuned to reach a level in which the real part of the surface impedance Z_s is close to zero; this is obtained by post-processing numerical optimization using Ansys HFSS, in order to find the finest combination between the unit cell patches of the super-cell design using Floquet boundary conditions. Table 6 shows the optimized dimensions d_{yo}. Furthermore, the surface impedance of the optimized super-cell (Fig. 32) is calculated after optimization where $Z_s = 0.9 - j267.3$ at 14.7 GHz. The real part, which defines the lossy resistive element, is lowered to almost zero, eliminating the super-cell design's resistive losses.

Multi-angle Reactive Retrodirective Metasurface Design

Following the design of the retrodirective super-cell and optimizing its surface impedance to be purely reactive, a metasurface using the nominal super-cell and another using the reactive optimized super-cell are designed with similar dimensions 5 cm × 23 cm to compare their performances using Ansys HFSS; both metasurface designs are displayed in Fig. 34.

The normalized monostatic RCS is calculated for both metasurface designs, nominal and optimized, to compare their performances in terms of retrodirectivity. The results are given in Fig. 35.

Figure 35 indicates and shows the retroreflection performances of both metasurface designs. In blue for the nominal metasurface, the results show a high level of retrodirectivity at 10° which is expected following the given angle of incidence and periodicity of the design. Retroreflection is also present at various angles such as −20°, −44°, −32°, and − 60° but with lower values and less efficient compared to the desired given angle. The presence of retroreflection at these angles in expected

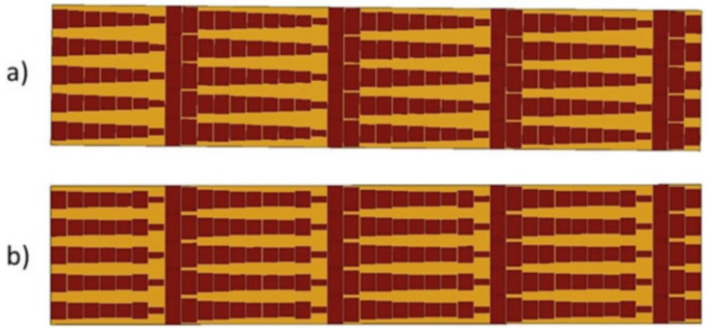

Fig. 34 (**a**) Nominal retrodirective metasurface of 5 × 4 super-cells. (**b**) Reactive optimized retrodirective metasurface of 5 × 4 super-cells

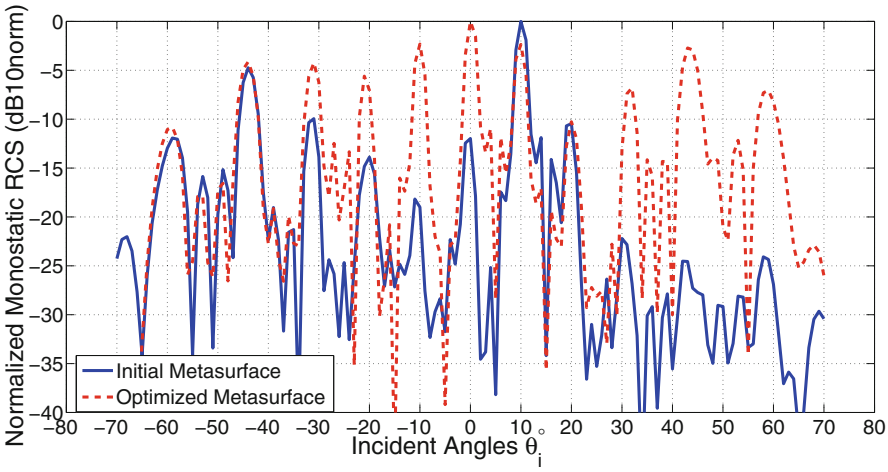

Fig. 35 The blue curve shows the normalized monostatic RCS of the nominal metasurface design (Fig. 34a) in comparison to the optimized reactive metasurface (Fig. 34b) as shown in red. The metasurfaces are of similar dimensions 5.1 cm × 23.5 cm operating at a frequency of 14.7 GHz

due to higher orders of diffraction being present at high periodicity designs. Using Floquet-Bloch analysis in Eq. (7), the directions of the open apertures can be calculated. In this case, for $L_x = 2.88\lambda$, n is limited to a certain range of values as follows $n = 0, \pm 1$, and ± 2. We conclude that five diffraction orders are expected at $\theta_i = 0, \pm 20.3°$, and $\pm 44°$. Due to the periodicity being approximately equal to 3λ, the two further higher orders of diffraction at *pm*32*circ* and *pm*60*circ* may be explained. Thus, five diffraction orders are expected at $\theta_i = 0, \pm 20.3°$, and $\pm 44°$. The two other higher orders of diffraction at $\pm 32°$, $\pm 60°$ can be explained due to the periodicity being approximately equal to 3λ. The symmetrical direction at $-10°$ and normal incidence are nonexistent and have low efficiency, as shown by the calculated values. This is due to the uncontrolled propagation of the wave in the direction of the higher diffraction orders.

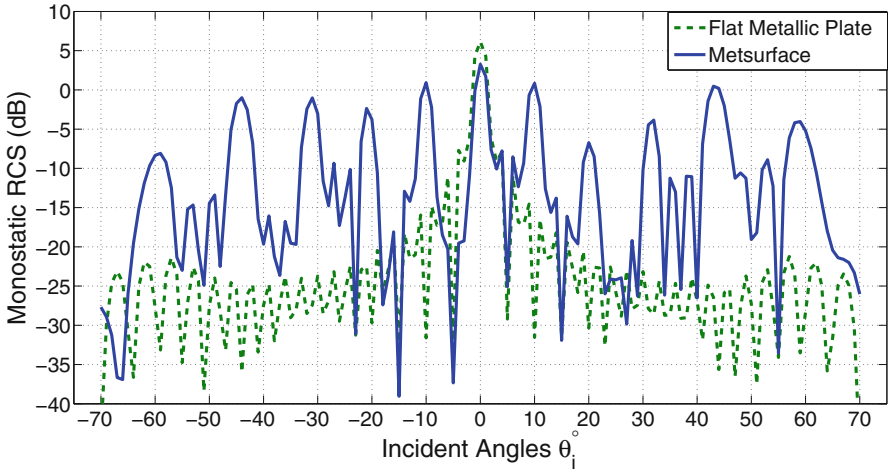

Fig. 36 Comparison between the reactive metasurface and a metallic flat plate with similar dimensions 5.1 cm × 23.5 cm by calculating the monostatic RCS to determine their performances in terms of retrodirectivity at an operating frequency of 14.7 GHz

On the other hand, for the optimized reactive metasurface shown in Fig. 34b, the results in Fig. 35 indicate that retrodirectivity is retained in the symmetrical direction at −10° as well as at normal incidence for the conventional properties with a higher efficiency compared to the nominal case. We can deduce that with the optimization of the surface impedance to be purely reactive, the resistive losses are eliminated. However, we can also notice the control of the higher orders of diffraction and achieving high-efficient retroreflection in the directions at ±20.3°, ±32°, ±44°, and ± 60°. The reactive metasurface is compared to a metallic flat plate of identical dimensions 5.1 cm × 23.5 cm in order to evaluate its efficiency in terms of retrodirectivity. The analytical study of the monostatic RCS for both metasurfaces is given in Fig. 36.

We can indicate from Fig. 36 that the optimized reactive metasurface can perform retroreflection at 11 incident angles simultaneously: normal incidence, ±10°, ±20.3°, ±32°, ±44°, and ± 60°. The metasurface shows high efficiency in terms of retroreflection at the given angles. To determine how efficient it can be, it is compared to a metallic flat plate, as shown in Fig. 36, which is highly efficient at normal incidence. The gain has been obtained for each angle of retroreflection of the reactive metasurface with respect to that of the metallic plate, as given in Table 7. However, some angles have shown some attenuation when it comes to retroreflection such as +20.3°, which can be due to some surface resistive losses. The metasurface operates at a given operating frequency of 14.7 GHz with a tolerance around the operating frequency of ±100 MHz. When it comes to broadband and dual polarization designs, one can consider other type of unit cell structures depending on the application considered (Chen et al. 2018; Kalaagi and Seetharamdoo 2018).

Table 7 The reactive metasurface gain (dB) in terms of retroreflection compared to a flat metallic plate of comparable dimensions 5 cm × 23 cm after analytical study of the monostatic RCS at an operating frequency of 14.7 GHz

ϕ_i°	Gain (dB)
−60	+17
−44	+23
−32	+22
−20.3	+21
−10	+32
0	−4
+10	+32
+20.3	+23
+32	+24
+44	+27
+60	+21

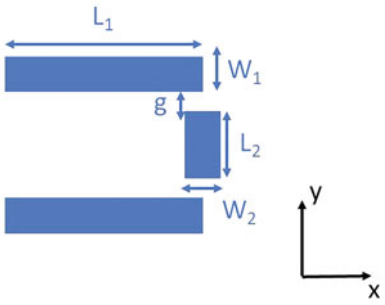

Fig. 37 The dolmen unit cell design, which is constructed of two sub-radiant modes of dimensions L_1 and W_1 given in the x-axis direction as well as a super-radiant mode of dimensions L_2 and W_2 given in the y-axis direction, whereas g represents the gap between both modes

Fano Resonance-Based Multiple Angle Retrodirective Metasurface

In this section, we use dolmen Fano-resonating structures (Yan et al. 2017; Verellen et al. 2009), these type of cells are used in applications for high-sensitive devices. The design procedure is similar to that introduced in previous sections in which a phase gradient is given between the unit cell dolmen elements of the super-cell. This is achieved by tuning the sub-radiant modes while fixing the super-radiant mode. The chosen unit cell is displayed in Fig. 37.

After choosing our unit cell element design, we fix an angle θ at 5° and an operating frequency of 24 GHz to determine the periodicity of the design, which implies that L_x is equal to 5.74λ. However, as introduced in the previous section, a high number of open channels can exist for such periodicity designs higher than λ (Epstein and Eleftheriades 2014), in which retrodirectivity can be obtained in addition to the given θ at 5° with a proper surface impedance optimization. The main advantage of Fano resonance-based unit cells is to suppress the losses and decrease the complexity in terms of surface impedance optimization, especially with the different degrees of freedom in this case. The super-cell is discretized into 8 unit

cells. The super-radiant mode is fixed, and a phase gradient is used to tune the sub-radiant modes of $\phi = 2\pi/8 = 45°$. The initial dimensions of the chosen dolmen unit cell at the given incident angle and frequency are the super-radiant mode fixed in the y-axis with $L_2 = 2$ mm and $W_2 = 1$ mm, the two sub-radiant modes in the x-axis with initial dimensions of $L_1 = 6.5$ mm and $W_1 = 1$ mm, and gap $g = 1$ mm. The dimensions after tuning of the sub-radiant modes with the given phase gradient are determined using the phase of the reflection coefficient given in Fig. 38.

The dimensions of the gradient unit cells are determined as follows: $L_{1\ (initial)} \times$ Factor X where $L_{1\ (initial)} = 6.5$ mm. Following the phase of the reflection coefficient with a phase gradient of 45°, one can determine the dimensions of the gradient unit cells which are given in Table 8.

For such high periodicity designs, we follow the same design procedure as we introduced in the previous section in terms of surface impedance optimization to be purely reactive. This is to control the retrodirectivity that can be obtained at the

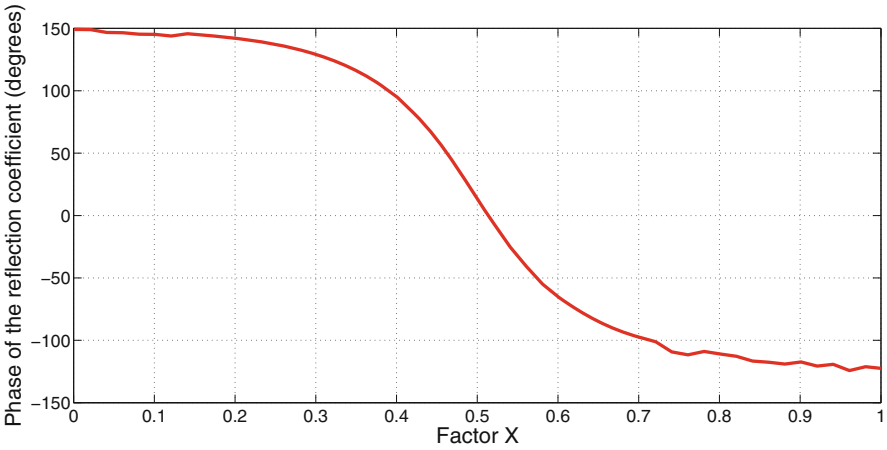

Fig. 38 Factor X used to be multiplied with the initial given dimensions of the sub-radiant modes following the phase of the reflection coefficient at an operating frequency of 24 GHz

Table 8 Factor X calculation following the phase of the reflection coefficient with a phase gradient of 45°, where L_1 (mm) the dimension of the sub-radiant modes in the x-axis are calculated by multiplying the initial value = 6.5 mm to the Factor X

$\phi_i°$	Factor X	L_1 (mm)
−122.4	1	6.5
−77.4	0.6	3.9
−32.4	0.55	3.575
12.6	0.5	3.25
57.6	0.44	2.86
102.6	0.37	2.405
147.6	0.1	0.65
−167.4	1	6.5

higher orders of diffraction. Thus we calculate the surface impedance of the nominal super-cell design which is determined as $Z_s = 15 - j326$. By optimizing and refining the initial dimensions to a solely reactive surface impedance, a suitable surface impedance can be achieved. $Z_s = 1 - j133$ for the reactive super-cell is achieved in which the resistive part is negligible. The dimensions of the reactive optimized super-cell design are given in Table 9.

After optimizing the surface impedance of the super-cell, the reactive super-cell with unit cell dimensions given in Table 9 is used to design a metasurface of 4 × 3 super-cells with dimensions 25 mm × 215 mm. A Teflon grounded substrate of permittivity $\varepsilon_r = 2.2$ and thickness 1.6 mm is used. The metasurface is displayed in Fig. 39.

In order to evaluate the Fano resonance-based metasurface design's performance, analytical studies have been performed using Ansys HFSS for the metasurface design to calculate the monostatic RCS in which the retroreflection performance of the metasurface is determined. The metasurface is excited with a propagating plane wave giving the E-field in the direction of the sub-radiant modes, at an operating frequency of 24 GHz and a range of incident angles between −70° and 70°. The results of the monostatic RCS of the metasurface design are compared to a metallic flat plate to analyze the efficiency of the metasurface in terms of retrodirectivity. The results are given in Fig. 40. High level of retrodirectivity is achieved at 17 incident angles simultaneously as we can deduce from the results in Fig. 40. The angle in which retroreflection is present are given as follows: −60°, −52°, −44°, −38°, −32°, −21°, −18°, −10°, −5°, normal incidence, +5°, +18°, +21°, +25°, +32°, +44°, and + 52°. This remarkable performance for multiple retroreflection using a single super-cell design is due to the high super-cell periodicity design given for the metasurface and the high sensitivity of the Fano resonant-based dolmen unit cell which is characterized by its high efficiency and accuracy.

Table 9 Dimensions of the reactive super-cell design after optimizing the sub-radiant modes to achieve a purely reactive super-cell compared to that of the initial dimensions

Initial dimensions L_1 (mm)	Optimized dimensions L_1 (mm)
6.5	6.5
3.9	4.095
3.575	3.705
3.25	3.38
2.86	3.055
2.405	3.9
0.65	1.3
6.5	6.5

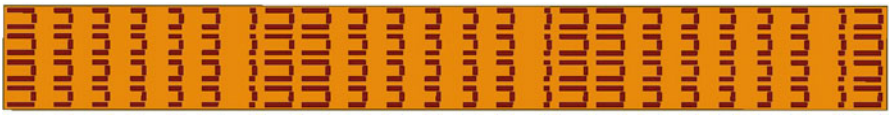

Fig. 39 The Fano resonance-based metasurface designed using Ansys HFSS with 4 × 3 reactive optimized super-cells and dimensions 25 mm × 215 mm, where the dolmen unit cells are printed on top of a Teflon grounded substrate of permittivity $\varepsilon_r = 2.2$ and thickness 1.6 mm

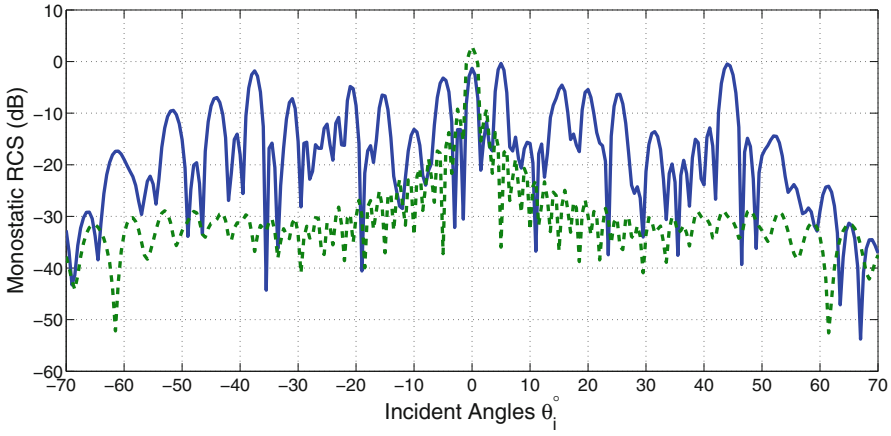

Fig. 40 Analytical study using Ansys HFSS of the Fano resonant-based metasurface design by calculating the monostatic RCS in dB in comparison to that of a metallic flat plate with a propagating wave where the E-field is given in the direction of the sub-radiant modes at 24 GHz

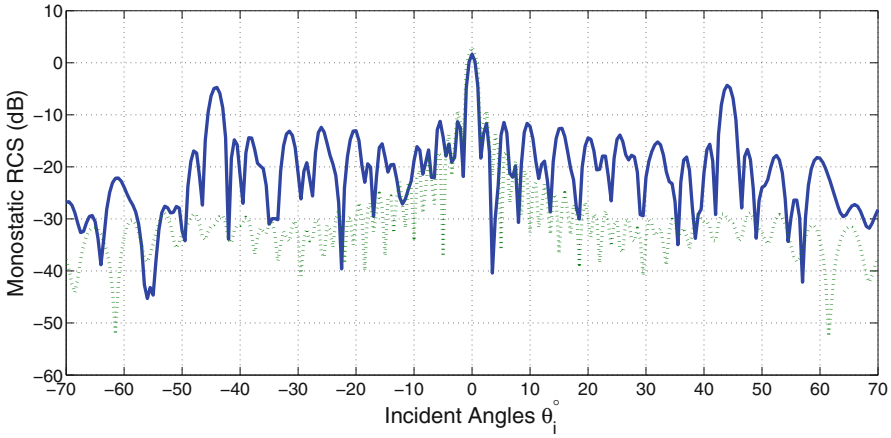

Fig. 41 Analytical study using Ansys HFSS of the Fano resonant-based metasurface design by calculating the monostatic RCS in dB in comparison to that of a metallic flat plate with a propagating wave where the E-field is given in the direction of the super-radiant mode at 24 GHz

To determine the potential of the metasurface design in terms of dual polarization performances, we excite the metasurface with a propagating plane wave with E-field in the super-radiant mode direction. Analytical study of the monostatic RCS of the metasurface for this scenario compared to a flat metallic plate is given in Fig. 41.

We can indicate from the analytical study of the monostatic RCS in Fig. 41 that retroreflection is achieved at three angles of incidence $-44°$ and normal incidence $44°$, although a phase gradient is absent in the direction of the super-radiant mode. This can be explained to be due to the present coupling for the unit cell design between the sub-radiant modes and the super-radiant mode.

Measurements and Experimental Results of the Multi-angle Reactive Metasurface Design

In this section, we present the experimental measurements of a prototype of the multi-angle metasurface (Fig. 34b) in order to validate the analytical studies and feasibility of the metasurface in terms of retroreflection. A prototype has been fabricated with 10 × 7 super-cells of dimensions 10.2 cm × 41.125 cm, with a Roger Duroid 5880 grounded substrate of dielectric constant $\varepsilon = 2.2$ and thickness 1.6 mm. The given prototype is displayed in Fig. 42. The experimental measurements are realized in an anechoic chamber, two Vivaldi antennas have been used, one transmitter and one receiver operating at a frequency of 14.7 GHz implemented at the same point 3 m away from the metasurface design, which is fixed on a turntable to analyze the monostatic RCS at a range of incident angles. Some absorbing materials have also been used between the two antennas in order to eliminate the coupling present and ensure ideal measurements. The normalized monostatic RCS of the given prototype is calculated and compared to that of the analytical results, as shown in Fig. 43.

We can indicate from Fig. 43 that the experimental results of the metasurface design are in good agreement with that of the analytical study of the metasurface using HFSS showing high performance in terms of retroreflection at multiple incident angles simultaneously, which are given as follows: ±10°, ±20.3°, ±32°, ±44°, ±60°, and normal incidence.

Radar Cross Section Measurements of a Mannequin Model With and Without a Metasurface Design

Project CYCLOPE is an European project which aims to protect vulnerable road users such as pedestrians and cyclists. With the implementation of radar systems in buses and vehicles, the detection and safety of such road users can be more efficient for drivers on the road. However, the RCS of cyclists and pedestrians can be low in terms of detection and can either scatter or absorb the transmitted signals. Retrodirective topologies and especially metasurface designs can be a

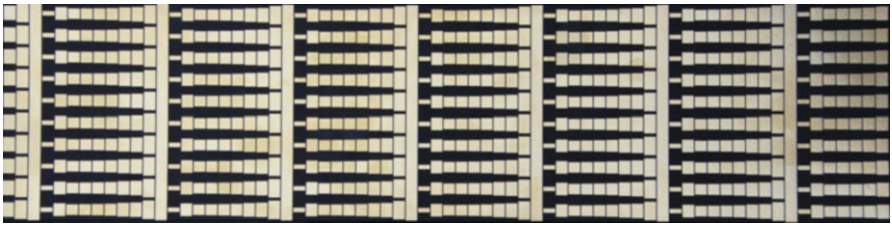

Fig. 42 Realized prototype of the multi-angle reactive retrodirective metasurface using 10 × 7 super-cells with dimensions 10.2 cm × 41.125 cm, and a Roger Duroid 5880 grounded substrate of dielectric constant $\varepsilon = 2.2$ and thickness 1.6 mm

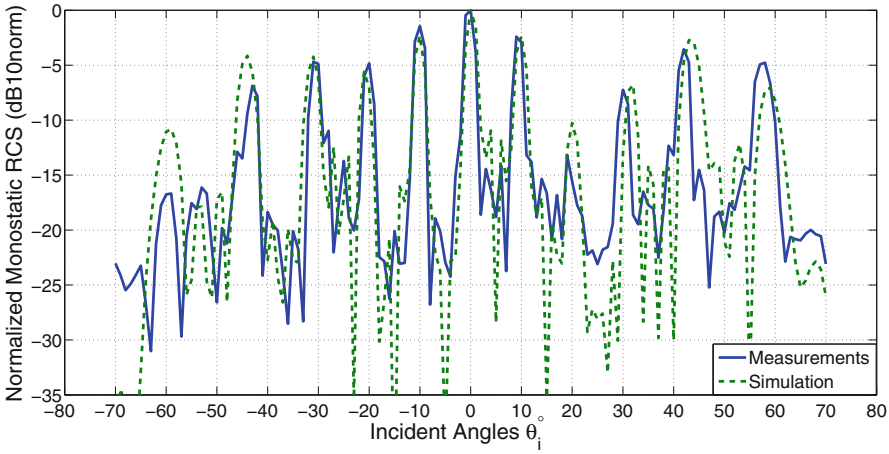

Fig. 43 Normalized monostatic RCS of the fabricated metasurface prototype (Fig. 42) compared to that of the analytical analysis of the design using HFSS (Fig. 34b) at an operating frequency of 14.7 GHz

complementary solution when implemented alongside such users to enhance their detection and localization. In this section, we implement and test one of our dual polarized retrodirective metasurface designs at 5° operating at 24 GHz alongside a mannequin model filled with water bags to imitate the structure of a human body. A radar system is set in the far field operating at 24 GHz and used to calculate the RCS of the human model with and without the metasurface design. The mannequin model scenarios are shown in Fig. 44.

Figure 44a shows the mannequin model filled with water bags where the relative permittivity is equal to 80. The metasurface (Fig. 22) is implemented on the mid part of the model. A dual polarized cross structure is used in this case, which is shown in Fig. 22, which is coherent for the application of radar systems and operating at 24 GHz for an oblique incident angle of 5°. The measurement is tested in an anechoic chamber which is said to be an ideal environment. Figure 44b shows the second scenario without the metasurface for comparison reasons. The automobile radar system used is implemented in the far field 3 m away from the given model. The radar system used is shown in Fig. 45.

To carry on with the RCS measurements, the mannequin model is set on a turntable which turns 360°. The radar transmits a constant signal during the measurements and calculates the RCS of the given model target. The RCS of both scenarios is given Figs. 46 and 47.

In Fig. 46, the RCS is calculated for the first scenario of the mannequin alone which is set on the turntable without the metasurface design. The radar detects the target giving a result shown in Fig. 46, with the target 3.6 m away from the radar at an angle of 4° and RCS equal to −10.5 dBm at an operating frequency of 24 GHz. Figure 47 shows the same measurement setup but with the metasurface being implemented alongside the human model, giving an RCS result of 5.5 dBm. This indicates that,

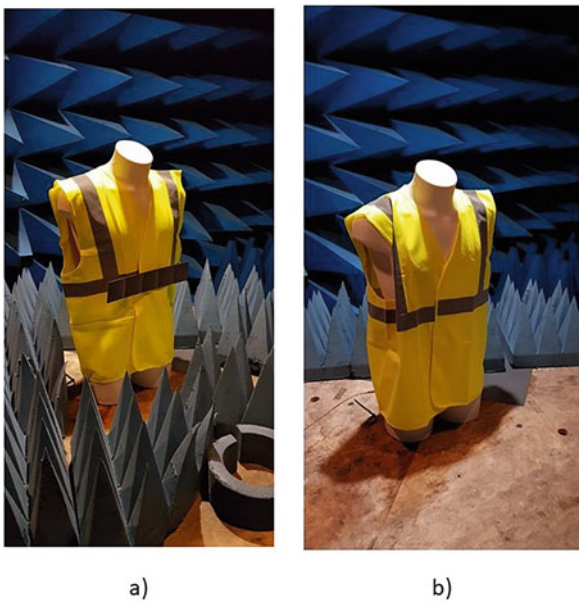

Fig. 44 (**a**) Mannequin human model with the dual polarized retrodirective metasurface structure given in Fig. 22 implemented on the mid part. (**b**) Mannequin model without the metasurface

Fig. 45 Automobile radar system operating at 24 GHz connected to a PC for RCS detection

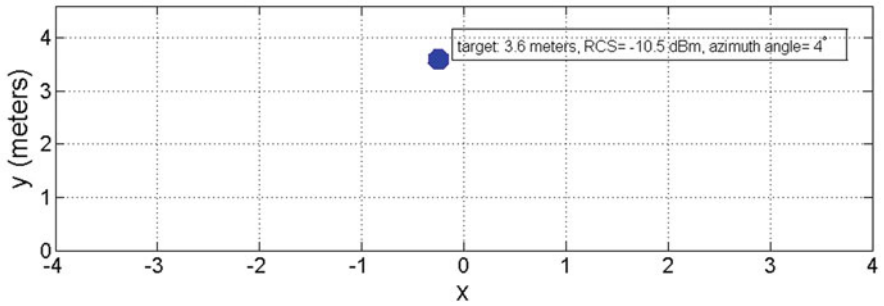

Fig. 46 Radar cross section detection of the model target by the radar system without the metasurface design

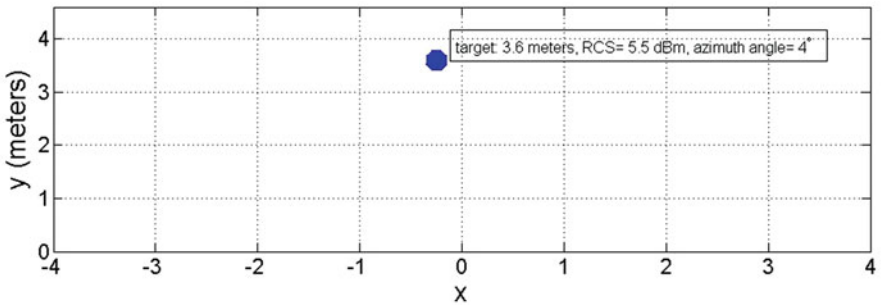

Fig. 47 Radar cross section detection of the model target by the radar system with the metasurface design

with the implementation of the metasurface design, the RCS of the target is enhanced and a gain of 15 dBm is achieved. This shows the remarkable performance of retrodirective metasurface designs from the application aspect in terms of RCS enhancement for different targets. For more information and visualization regarding the measurement setup and the application carried on, please refer to our video in the following link: https://www.youtube.com/watch?v=IZuhW-YXyvo.

Conclusion

Wave control and manipulation in the environment has gained a lot of attention by researchers recently, especially based on metasurfaces; this is due to their substantial physical properties, high efficiency, and reliability. Retrodirectivity, on the other hand, has shown to be a great feature for backscattering and localization enhancement for different targets. Moreover, in this chapter, we basically focus on investigating the potential of designing multi-angle retrodirective metasurfaces and study

their reliability in terms of retroreflection enhancement at a wide range of incident angles, where it can be of great usage for applications where tracking, localizing, and RCS enhancement is needed from the angular aspect.

We have introduced the design of retrodirective metasurfaces at various angles of incidence for retroreflection. The starting point was the generalized phase law of reflection, which states that for a nonhomogeneous surface design, an EM wave can be controlled in an anomalous direction. The main challenge was to achieve a design or metasurface topology which can operate at multiple incident angles simultaneously in terms of retroreflection. A design-based on superposition of multiple unit cells, each designed for a specific angle of incidence, was proposed. The retrodirective metasurface showed remarkable results in terms of retroreflection for a range of incident angles and was compared to a conventional corner dihedral. On the other hand, another study for multi-angle retrodirective metasurfaces was also presented to investigate more efficient, reliable, and compact design. The study was based on impedance modulation of the surface impedance of the retrodirective super-cell design. A retrodirective metasurface for 11 angles simultaneously was achieved. As well as using this technique for high periodicity designs, we were able to keep the classical characteristics of a retrodirective metasurface, especially in terms of reciprocity and retroreflection at normal incidence. Prototypes of different designs were fabricated and tested in an anechoic chamber to validate the concept and simulation results.

Retrodirective metasurfaces have shown interesting and unprecedented physical properties in terms of retroreflection, especially for multiple and a wide range of incident angles. The metasurface designs have shown remarkable results in terms of retroreflection, especially from the angular aspect compared to any conventional topology. The importance of these designs is especially in terms of application where they can be used for tracking, localization, and RCS enhancement, and a very reliable solution especially with their passive, compact, and easy implementation properties. In the context of project CYCLOPE, which aims to protect vulnerable road users, we tested a human model design using a mannequin and calculated the RCS of the model with the metasurface along its side and without the metasurface. The metasurface has shown to enhance the detection of the target's RCS with a gain up to 15 dBm compared to that when the model is set alone. This shows that metasurfaces have high potential from the applicative aspect where wave manipulation is needed.

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Part II

Microwave Circuits and Devices

Metamaterial Frequency Selective Surfaces As Polarizers

8

N. Yogesh and Zhengbiao Ouyang

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Abstract

Periodic array of dielectric/magnetic/metallic patterns also known as frequency selective surfaces (FSSs) are inevitable in manipulating and controlling the polarization state of electromagnetic radiation in the form of linear polarization converters, circular polarizers, asymmetric polarization converter, and polarization filters. FSSs with polarization controlling features are particularly useful in tailoring radiation pattern and improving gain functionalities of antennas and

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wavefront shaping. This chapter covers polarizing functionalities of FSS structures in the background of metamaterial (MTM) literature. Especially, the design method of MTM FSS polarizing structures, origin and mechanism of polarization, computational and experimental characterization methods, latest and significant developments made in MTM FSS polarizers and its various application aspects are presented at microwave frequencies. Extending MTM FSS polarizing elements for terahertz and optical frequencies is also discussed. Finally the present and future challenges in the MTM FSS polarizing structures are highlighted.

Keywords

Frequency selective surfaces · Metamaterial · Metasurface · Polarizers · Microwave devices

Introduction

Frequency selective surfaces (FSS) are the periodic array of dielectric/magnetic/metallic patterns used to control and manipulate electromagnetic radiation (e-m), based on frequency selective characteristics. For an example, a periodic array of thin vertical metallic cut-wires embedded on a dielectric substrate can act as a polarization filter for transverse electric (TE) polarization. Such an FSS can block or allow a band of e-m frequencies to pass through it depending on the scattering element's dimensions and unit cell periodicity. In RF/microwave/terahertz (THz)/infrared and visible spectra, FSS plays an integral part in antenna designs for gain enhancement, and radar cross-section reduction, bandpass filters, polarization tuning, wavefront shaping, effective ground planes, and perfect absorbers. Especially FSS technology is seminal in the design of various polarizers for RF, microwave, and optical systems, mainly because, conventionally, one needs to rely on optically active or birefringent materials for the design of polarizers. However such conventional materials are bulky and are not suitable for integrated circuits. Moreover all conventional materials won't possess optical activity at RF/microwave/THz frequencies. On the other hand, artificially synthesized FSSs can tailor the polarization state over a wide bandwidth and practically they can be designed for any e-m spectra depending on the choice of metallic patterns and dielectric substrates. Figure 1 presents different FSS geometries used in the polarization designs (Anand and Mittal 2020).

Though FSS literature at least dates back to 1980 (Munk 1980), the current research in FSS on the background of metamaterials (MTMs) opens a new avenue to control e-m radiation in unprecedented ways. Metamaterials are the artificial sub-wavelength periodic dielectric/magnetic/metallic constituents whose e-m properties such as dielectric permittivity, magnetic permeability, refractive index, and chirality can have negative/unusual values, and show exotic e-m phenomena such as negative refraction, double-focusing, cloaking, reversal of Doppler shift, giant optical activity, circular dichroism, and so on.

In general, an MTM may be considered as a kind of FSS, as both they have periodic arrangement of scattering elements, but MTMs are distinctly different from

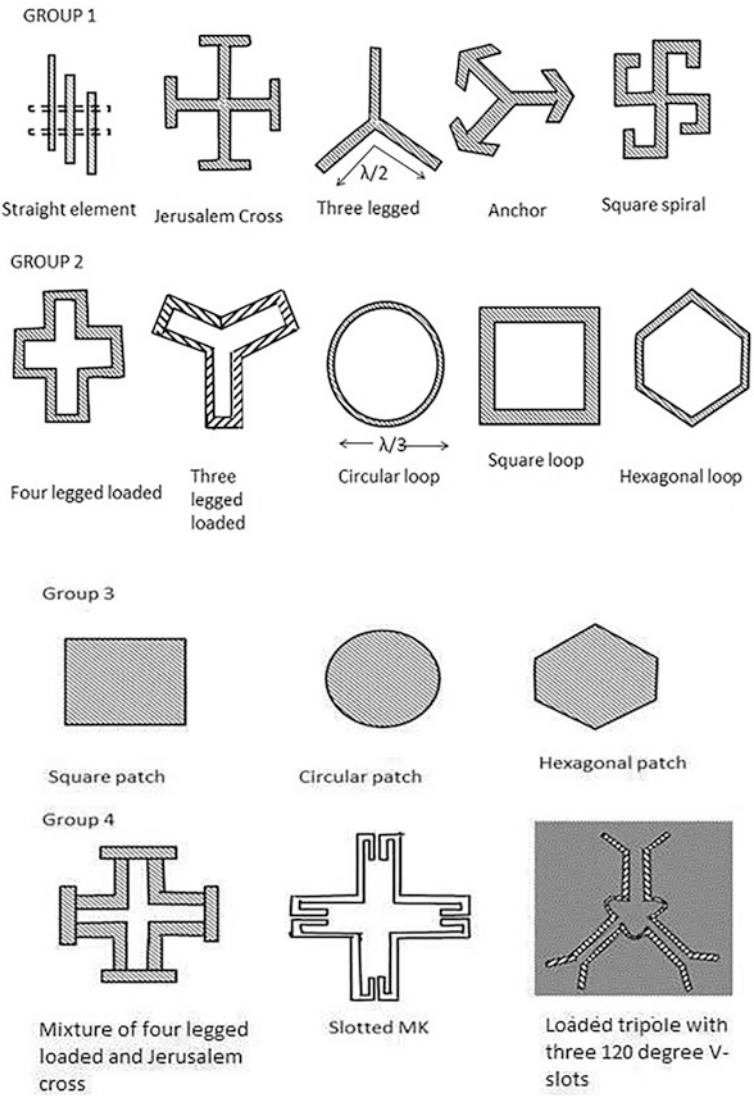

Fig. 1 FSS Geometries, Anand and Mittal (2020). (Reproduced courtesy of The Electromagnetics Academy)

FSS where MTMs are effectively homogeneous, where periodicity and size of the each unit cell in MTMs are much smaller than the wavelength of e-m wave, and the unit cell in MTM is called as meta-atom. In case of a FSS, a single unit cell may not have an effective material definition and they are strongly scattering, and their periodicity can be of the order of wavelength of e-m radiation. Apart from material definition, there exists a sub-class of MTMs known as metasurfaces (MSs). Metasurfaces (Li et al. 2018) are 2-D version of MTMs in which scattering elements are

arranged with 2-D periodicity as that of FSSs. In MSs, two classes exists, namely: (i) metafilm – An array of isolated (non-touching) scatterers and (ii) metascreen – It is a complementary structure of a metafilm, which consists of hole patterns on a metallic sheet embedded in a dielectric substrate (Holloway et al. 2011, 2019). Both metascreen and metafilm definitions already accounted in FSSs literature as aperture/slot type FSS and patch/strip type FSSs, respectively.

Depending on the transmission and reflection of e-m waves, MSs, also classified as transmission MSs and reflective MSs in which reflective MS are backed with complete metal plane and polarization functionalities, are achieved with reflected e-m waves.

Based on the prevalence of the above notions and definitions, the polarization designs based on planar 2-D structures are identified and addressed as MTM FSSs. Wherever material definition exists as per effective homogeneous response, 2-D structures are identified as MTM structure.

The design of MTM FSS polarizers are based on the introduction of chiral, anisotropic, and bianisotropic inclusions in MTM FSS structures either with transmission or reflective types. Chiral inclusions are the one in which an object and its image does not superimpose with each other. For example, in Fig. 1, the square spiral structure presented under group-1 is an example of a chiral structure, in which its image (mirror inversion) cannot superimpose with its original object. The salient feature of chiral inclusion is its optical activity, i.e., it can be able to rotate the polarization state of the e-m radiation.

Anisotropic inclusions are the one that supports either electromagnetic or magnetic–electric interactions for the incident e-m radiation. For instance, in Fig. 1, the patterns of three-legged, Jerusalem cross and anchor designs are capable of exhibiting anisotropic interactions for transverse electric (TE) and transverse magnetic (TM) polarization incidence. Bianisotropic inclusions are the one that exhibits both electromagnetic and magnetic–electric interactions for the incident e-m radiation. Bianisotropic patterns are an example of nonreciprocal structures and are useful for novel control of polarization state of e-m radiation.

This chapter presents different polarization phenomena in MTM FSS structures, the mechanism of polarization functionalities in MTM FSSs, design principle of polarizers, experimental characterization and measurement of optical activity of MTM FSS structures, applications of MTM FSS polarizers, and briefly accounts prospects and challenges at present in the field of MTM FSS at microwave, THz, and visible frequencies.

Origin and Mechanism of Optical Activity in Artificial MTM Structures

Background

In general, polarizers are designed based on waveplate concept using birefringent material. A waveplate changes the polarization state of the incident e-m radiation by producing a phase shift between the parallel and perpendicular components of the electric field vector of the incident e-m field as $\Delta\phi = \frac{2\pi L}{\lambda}\left(n_{\perp} - n_{\parallel}\right)$, where L is the

thickness of the waveplate, λ is wavelength of e-m field in air, and $n_{\perp}$, $n_{\parallel}$ are the refractive indices along the perpendicular and parallel axes of the waveplate, respectively.

A half-waveplate produces π phase difference between the parallel and perpendicular components of the E-vector of the incident e-m field so that a linearly polarized wave is rotated in the output. Hence a half-waveplate acts as a polarization rotator. On the other hand, a quarter-wave plate adds $\pi/2$ phase difference between the parallel and perpendicular components of the E-vector of the incident e-m field, so that a linearly polarized wave is converted into a circularly polarized wave. Thus a quarter waveplate acts as a circular polarizer.

Apart from natural materials, waveplates and polarization devices can be designed using MTM FSS structures based on electric and magnetic dipole interactions. Figure 2a and 2b show the first ever numerical (Svirko et al. 2003) and experimental (Papakostas et al. 2003) reports on artificial periodic patterns exhibiting strong optical activity. Figure 2a shows a bilayered cut-wire patterns which are mutually twisted each other. When TE wave is incident on such a mutually twisted pattern, inductive coupling is induced, where electric dipole moment excited in the twisted patterns oscillate antisymmetrically. It is known that antisymmetric oscillation of two electric dipoles constitutes a magnetic dipole, in which magnetic dipole moment is induced according to Ampere circuital law. This induced magnetic dipole moment interacts with the externally incident e-m field and results in the rotation of polarization state for the incident wave. To visualize more about the role of dipole interaction in optical activity of MTM FSS structures, it is first useful to classify the range of polarization effects observed in MTM FSS structures based on transmission and reflection coefficients of MTM FSSs.

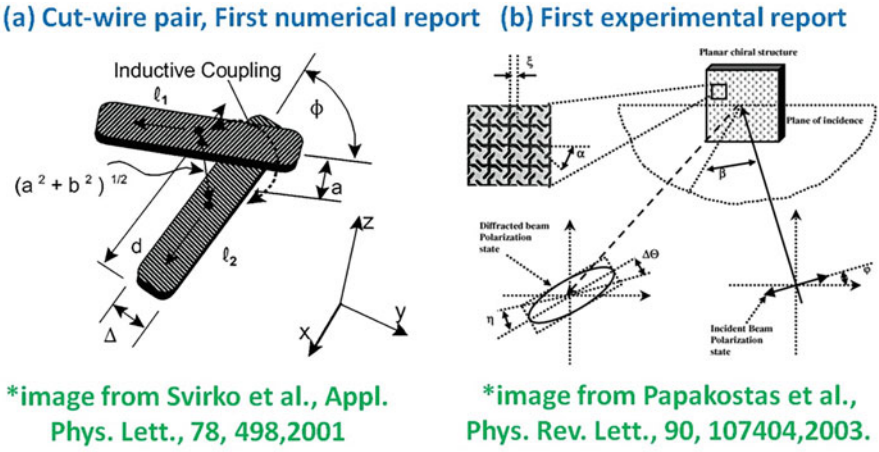

Fig. 2 (**a**) Mutually twisted metallic pattern – first numerical report on artificial chiral structure. (©American Institute of Physics, Svirko et al. (2003)). (**b**) Gammodian pattern – first experimental report on artificial chiral structure. (©American Physical Society, Papakostas et al. (2003))

Characterization of Polarization Phenomena in Metasurface FSS Polarizers

In the characterization of MTM FSS polarizer, calculation of co- (T_{xx} and T_{yy}) and cross-polarization (T_{xy} and T_{yx}) transmission coefficients are important and they are described as

$$T_{xx(yy)} = \frac{E^t_{x(y)}}{E^i_{x(y)}} \text{ and } T_{xy(yx)} = \frac{E^t_{x(y)}}{E^i_{y(x)}}$$

where $E^i_{x(y)}$ and $E^t_{x(y)}$ are the incident and transmitted electric field for TM (TE) incident polarizations, respectively, and $E^i_{y(x)}$ is the incident electric field for TE(TM) incident polarizations, respectively.

For an ideal cross-polarizer (90° pure optical rotation) in which TE(TM) is converted to TM (TE), $T_{xy(yx)} = 1$ and $T_{yy(xx)} = 0$.

Once we calculate the co- (T_{xx} and T_{yy}) and cross-polarization (T_{xy} and T_{yx}) transmission coefficients, optical activity of any FSS-MTM polarizer can be computed in the circular polarization (CP) basis as follows;

$$T_{++} = T_{yy} + iT_{xy}, T_{--} = T_{yy} - iT_{xy},$$

$$\phi = \frac{1}{2}[\arg(T_{++}) - \arg(T_{--})],$$

$$\eta = \frac{1}{2}\tan^{-1}\left(\frac{|T_{++}|^2 - |T_{--}|^2}{|T_{++}|^2 + |T_{--}|^2}\right),$$

where T_{++} and T_{--} are the transmission coefficients of right-(RCP) and left-circularly polarized (LCP) waves, respectively, ϕ is the polarization rotation angle, and η is the ellipticity of transmitted wave. For an ideal cross-polarization converter, ϕ and η should be 90° and 0°, respectively. Apart from cross-polarization, pure optical rotation, where polarization angle can be other than 90° is characterized by zero ellipticity. Nonzero ellipticity values represent the elliptically polarized wave.

Similar to TE-to-TM conversion, the linear polarization (LP) to circular polarization (CP) conversion by an MTM FSS can be characterized by Jones matrix as follows;

$$T_{\text{circular}} = \begin{pmatrix} T_{+x} & T_{+y} \\ T_{-x} & T_{-y} \end{pmatrix} = \begin{pmatrix} T_{xx} + iT_{yy} & T_{xy} + iT_{yy} \\ T_{xx} - iT_{yx} & T_{xy} - iT_{yy} \end{pmatrix}$$

$$\begin{pmatrix} E^t_+ \\ E^t_- \end{pmatrix} = \frac{1}{\sqrt{2}} \begin{pmatrix} T_{+x} & T_{+y} \\ T_{-x} & T_{-y} \end{pmatrix} \begin{pmatrix} E^i_x \\ E^i_y \end{pmatrix}$$

$$\eta_2 = \frac{1}{2}\tan^{-1}\left(\frac{|E_+| - |E_-|}{|E_+| + |E_-|}\right)$$

where E^t_+ and E^t_- are the transmission fields of right-handed (RCP) and left-handed (LCP) circularly polarized waves, respectively; E^i_x and E^i_y are the x- and y-oriented fields of the linearly polarized incident waves, respectively; $T_{+x}(T_{-x})$ and $T_{+y}(T_{-y})$ are the transmission coefficients for the RCP (LCP) waves corresponding to x- and y-oriented linearly polarized incident waves, respectively; and η_2 is the ellipticity of transmitted light expressed in CP basis. It is noted that $\eta_2 = 45^\circ$ corresponds to pure circularly polarized light.

Apart from TE to TM (vice versa) and LP to CP conversion, two new optical phenomena of MTM FSS, namely, the asymmetric transmission (AT) for (i) LP and (ii) CP waves can be defined as follows: Suppose TE to TM conversion (and vice versa) of an MTM FSS is not the same for forward/backward incident excitations, in which $\Delta^{AT}_{\text{linear}} = |T_{xy}|^2 - |T_{yx}|^2 \neq 0$, then the MTM FSS exhibits AT for LP waves. Similarly if RCP to LCP conversion (and vice versa) is not same for forward/backward incident excitations, in which $\Delta^{AT}_{\text{circular}} = |T_{-+}|^2 - |T_{+-}|^2 \neq 0$, then the MTM FSS exhibits AT for CP waves.

While it is noted that the above formulation is used to characterize the polarization functionality of MTM FSS in transmission mode, one can straightforwardly use a similar approach to characterize the polarization functionality of MTM FSS in reflection mode, as reflective MS is also employed to control and tune the polarization stated of the e-m radiation.

Table 1 lists the polarization phenomena in MTM FSSs.

Dipole Interactions in MTM FSS Structures

Figure 3 shows co- and cross-polarization coefficient plots for four different MTM FSS bilayer structure (Rajkumar et al. 2013) of which Fig. 3c alone shows strong cross-polarization resonance. This can be explained based on the electric and magnetic dipole interactions in the MTM for the incident e-m field.

Table 1 Polarization Phenomena in MTM FSSs

Polarization phenomena in MTM FSS	Required conditions
Ideal cross-polarizer (TE to TM) (vice versa)	$T_{xy(yx)} = 1$ and $T_{yy(xx)} = 0$, $\phi = 90^\circ$ and $\eta = 0$ Pure optical rotation $\phi \neq 90^\circ$but $\eta = 0$
Circular Polarizer (LP to RCP/LCP)	$\|T_{xy}\| = \|T_{yy}\|, \theta_{xy} - \theta_{yy} = (2n-1)\frac{\pi}{2}$ and $\eta = 45^\circ$
Asymmetric transmission for LP waves	$\|T_{xy}\|_{\text{forward}} \neq \|T_{xy}\|_{\text{forward}}$ $\Delta^{AT}_{\text{linear}} = \|T_{xy}\|^2_{\text{forward}} - \|T_{yx}\|^2_{\text{forward}} \neq 0$
Asymmetric transmission for CP waves	$\|T_{+-}\|_{\text{forward}} \neq \|T_{-+}\|_{\text{forward}}$ $\Delta^{AT}_{\text{linear}} = \|T_{+-}\|^2_{\text{forward}} - \|T_{-+}\|^2_{\text{forward}} \neq 0$

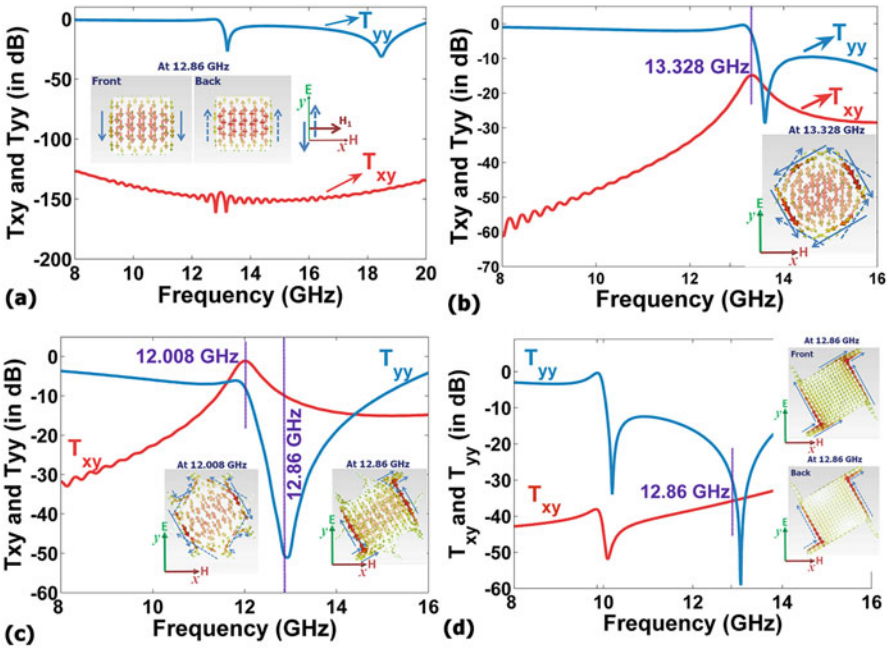

Fig. 3 T_{xy} and T_{yy} plots for MTM FSSs with (**a**) parallel square, (**b**) rotated square, (**c**) rotated-arm-square, and (**d**) mirror-symmetry rotated-arm-square unit cells. In the inset of (**a**), surface current plot for the front and back sides of parallel square unit cell is given at 12.86 GHz. Coupling scheme is also depicted in the inset of (**a**). In the inset of (**b**), surface current plot at 13.328 GHz is given for the rotated square unit cell. In the inset of (**c**), surface current plots at 12.008 GHz and 12.86 GHz are given for rotated-arm-square unit cell. In the inset of (**d**), surface current plot at 12.86 GHz is given at front and back sides of the mirror symmetry rotated-arm-square unit cell. In all the insets, solid and dotted arrows represent the surface currents at front and back sides of the unit cells, respectively. (©American Institute of Physics, Rajkumar et al. (2013))

Figure 3a corresponds to parallel-square MTM in which no cross-polarization is observed in the entire frequency range of 8 –20 GHz. In the inset of Fig. 3a, surface current plot is given at 12.86 GHz for front and back layers of the MTM FSS for normal incident TE polarized wave. Under TE excitation, electric field is oscillating along the vertical sides of the square layer. Therefore electric dipole moments are created along the vertical sides of the front square MTM pattern. Similarly vertical currents are also excited along the vertical sides of the back MTM layer. However, currents in the back metal oscillate antisymmetrically with respect to front layer. It is known that two oppositely oriented electric dipole constitutes a magnetic dipole in which a magnetic dipole moment is created in the direction perpendicular to the Amperian loop formed by the oppositely oriented electric dipoles. This scenario is depicted in the inset of Fig. 3a. Since the induced magnetic dipole moment is perpendicular to the E-field of the TE wave, cross-coupling interaction $U_{\text{interact}} = -\vec{\mu}_{\text{induced}} \cdot \vec{E}_{\text{external}} - \vec{p}_{\text{induced}} \cdot \vec{H}_{\text{external}}$ is absent, where U_{interact} is the interaction potential energy, $\vec{\mu}_{\text{induced}}$ and $\vec{p}_{\text{induced}}$ are the induced magnetic dipole and electric

dipole moment in MTM FSS, respectively, and $\vec{E}_{\text{external}}$ and $\vec{H}_{\text{external}}$ are the externally applied electric and magnetic fields (e-m wave), respectively. Hence there is no cross-polarization for a parallel square MTM FSS.

In case of a rotated square MTM FSS, a small cross-polarization peak occurs at 13.328 GHz in Fig. 3b due to the fact that oppositely rotated squares result in nonzero U_{interact}. However, dipole interaction is not sufficient enough to produce cross-polarization.

In Fig. 3c, co- and cross-polarization transmittances of a rotated-arm-square MTM FSS are plotted. In this case, strong cross-pol resonance is observed at 12.008 GHz and pure cross-polarization conversion, where zero co-polarization is observed at 12.86 GHz. Both of these frequencies support strong magnetoelectric coupling (surface current is plotted in the inset of Fig. 3c) and therefore cross-polarization conversion is dominant.

To emphasize the role of magnetoelectric coupling in cross-polarization conversion, co- and cross-polarization transmittances of mirror-symmetric rotated-arm-square MTM FSS are plotted in Fig. 3d. Mirror symmetric MTM FSS is the one where front and back layers have mirror symmetry with each other. Such patterns will not show any cross-polarization resonance as there is no cross-coupling interaction between induced magnetic dipole moment and applied electric field. This is clear from the surface current plot given in the inset of Fig. 3d at 12.86 GHz, where surface currents at the front and back sides of the metal layers are symmetric in nature. It indicates that the unit cell as a whole acts as an electric dipole and there is no magnetic response and hence no cross-polarization is observed.

Design Methodologies of MTM FSS Polarizers

Up to the last section various polarization phenomena and dipole interaction pictures in MTM FSS structures are presented. In this section, the realization method of MTM FSS polarizer is presented at microwave frequencies. For this purpose, a bilayered Fermat's spiral pattern deposited on a dielectric substrate is taken as shown in Fig. 4.

Optical Activity and Chirality Measurement of MTM FSS

Figure 4 shows the unit cell of the double-layer twisted Fermat's spiral chiral metamaterial (FSCMM). The chiral design given in Fig. 4a shows the front and back layers of FSCMM, where the back layer is a mirror of the front layer. This arrangement breaks the mirror symmetry along the direction of propagation of light and such design can induce a chiral coupling in planar structures. For the MS design, a quadfilar Fermat's spiral (FS) is designed in each layer, where four FS rings are arranged, with an angular separation of 90° with each other. It is known that the eigenstates of C_4 symmetry structures are purely circularly polarized light so that mixed polarizations are avoided. The FS ring proposed has a functional dependence of $r = A\sqrt{\theta}$, where A is the spiral constant, representing the degree of winding for

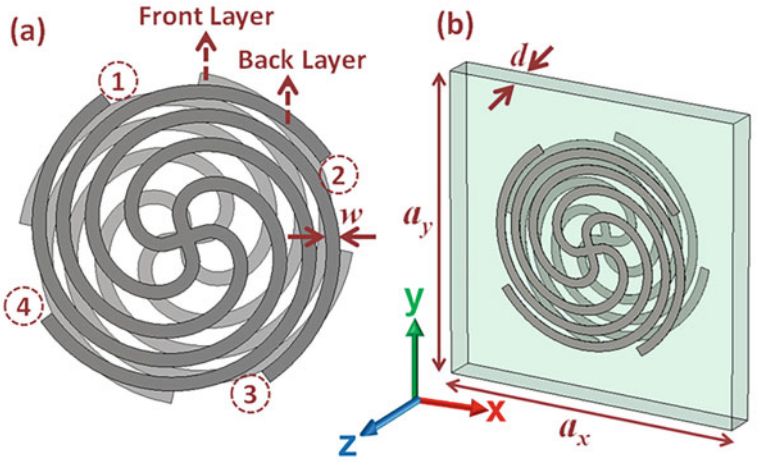

Fig. 4 (**a**) Chiral design of FSCMM. (**b**) Unit cell of the FSCMM

the given number of turns. Since FS rings are more compact than normal Archimedean spiral ($r = A\theta$), the CMM is adopted with FS rings of 2π turns.

Figure 4b shows the single unit cell of the FSCMM. The geometrical parameters $A = 1.4\ \text{mm}/\sqrt{\text{rad}}$, $w = 0.3$ mm, $t = 0.012$ mm, $d = 1$ mm, and a_x=$a_y = 10$ mm are the spiral constant of the FS ring, the width of spiral, the thickness of metal, the thickness of dielectric substrate, and lattice constants, respectively.

An MTM FSS polarizer can be generally designed by solving Maxwell's equations numerically using finite-difference time-domain method (FDTD) (Commercial e-m solvers CST Microwave Studio, Comsol Multiphysics, Lumerical, Rsoft, Empire Xcel), and frequency domain methods (FEM) (CST Microwave Studio, Comsol Multiphysics and HFSS).

To compute the optical activity of the proposed FSCMM, full-wave e-m simulations are carried out using commercial e-m solver CST Microwave Studio using both FEM and FDTD methods. In FEM method, periodic boundary conditions are applied along x and y directions in the unit cell, and Floquet ports are employed at the input/output interfaces. In FDTD method, an incident planewave is swept for the given frequency range and time-domain signals are recorded using probes placed at the input/output ports. After sufficient decay of input pulse, transmission and reflection coefficients are computed by taking Fast Fourier Transform of the time-domain signals. In FDTD calculations, periodic boundaries are specified along x- and y-directions of the FSS unit cell.

The FSCMM shows cross-polarized resonance at 14.372 GHz (Fig. 5a) with the transmission coefficients of $T_{xy} = 0.962$ and $T_{yy} = 0.002$. It is verified that an E-polarized beam is converted into an H-polarized beam and the computed polarization-rotation angle and ellipticity (Fig. 5b) are found to be 89.87° and 3.281×10^{-4}, respectively. In Fig. 5c chirality parameter $\kappa = (n_{rcp} - n_{lcp})/2$ is computed using standard retrieval procedure for a reciprocal bi-isotropic medium at normal incidence, as described in Plum et al. (2009). It is clear that real(κ),

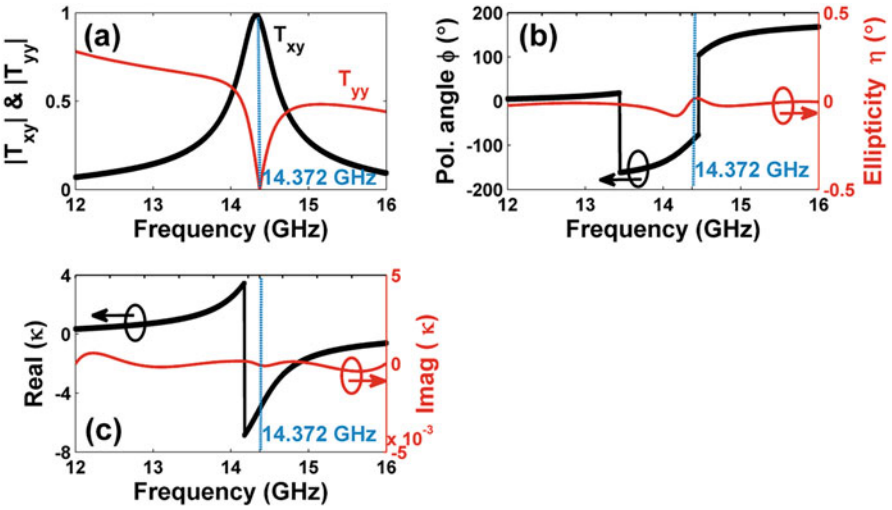

Fig. 5 Optical activity of FSCMM under ideal lossless case (Metal is treated as PEC and the dielectric substrate has $\varepsilon_r = 4.5$). (**a**) Plot of cross- (T_{xy}) and co-polarization (T_{yy}) transmission. (**b**) Polarization rotation angle (left y axis) and Ellipticity (right y axis). (**c**) Chirality plot

representing the optical activity, shows resonant behavior whereas imag(κ), representing the circular dichroism, is zero. Since this result is for ideal case, one has to affirm how the proposed CMM can be useful for microwave and THz regime as a cross-polarization converter and what the minimum values of ellipticity and imag(κ) under practical conditions would be.

Figure 6 accounts the effect of dielectric dispersion and metal losses in degrading the pure optical activity of FSS-MTM polarizer for differential metal-dielectric combinations. For instance, the polarization angle at cross-polarization resonance is nearly 90° for perfect electric conductor (PEC) and thermoset microwave material laminate 4 (TMM4), Silver (Ag)-TMM4, and copper (Cu)-glass reinforced epoxy laminate material (FR4). The polarization angle and change in the real part of chirality (Fig. 6c), representing optical activity, can be correlated with each other. However, the ellipticity (Fig. 6b) changes from 2° (ideal lossless case) to 5.17° (even for finite metal conductivity), and in case of strong dielectric losses, the ellipticity value is quite high (18.91°, for lossy FR4 substrate with Cu patterns). The change in ellipticity is in accordance with the increase in the imaginary part of chirality (measure of circular dichroism) of FSS-MTM polarizer as given in Fig. 6d.

Microwave Experimental Setup for Characterization of FSS-MTM Polarizers

To measure the optical activity of FSS-MTM polarizers, experimental setup shown in Fig. 7 is commonly used at microwave frequencies. For this purpose, an anechoic chamber for eliminating the unwanted reflection and surrounding noise is used. Standard

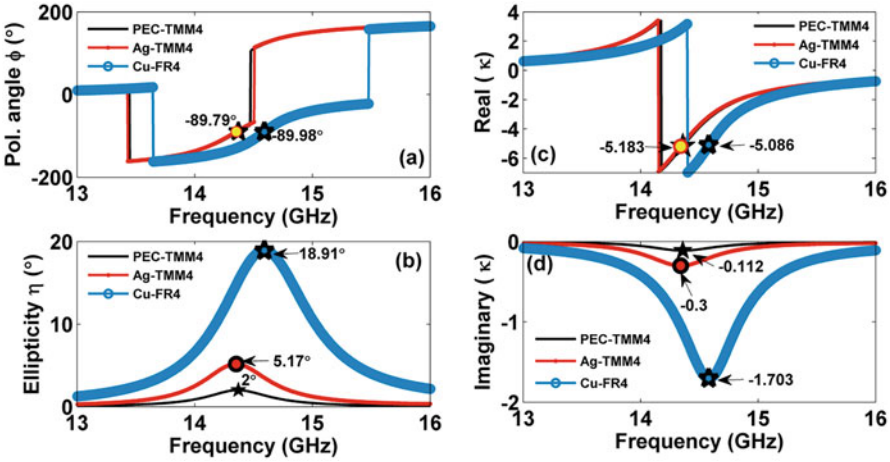

Fig. 6 Role of dielectric dispersion and metal losses in optical activity of FSS-MTM cross-polarizer. Plots of (**a**) polarization angle, (**b**) ellipticity, (**c**) real, and (**d**) imaginary parts of chirality for three different dielectric-metal combinations such as PEC-TMM4, Ag-TMM4, and Cu-TM4

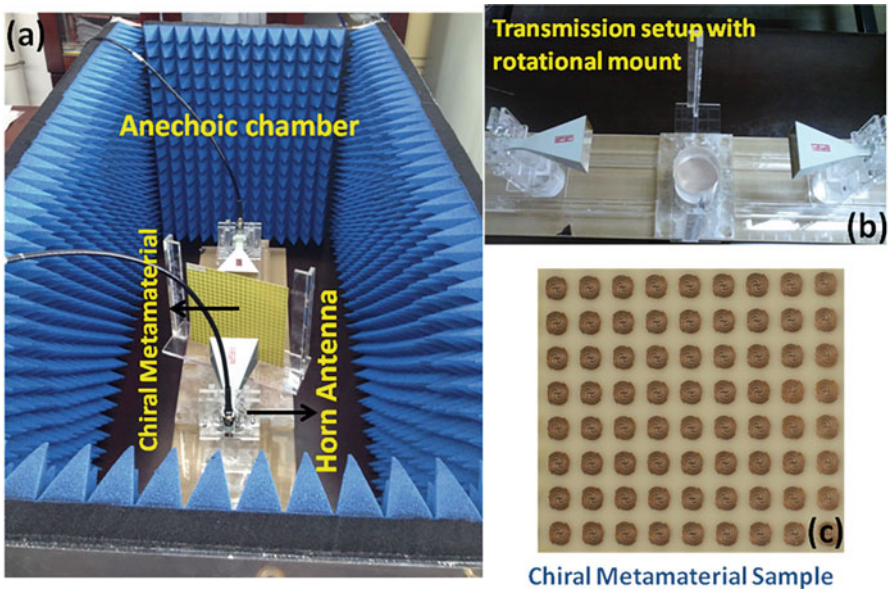

Fig. 7 (**a**) Experimental setup for the optical activity measurement of CMM at microwave frequencies. (**b**) Rotational mount. (**c**) Fabricated CMM sample using standard FR4 board

gain microwave horn antennas are used as the transmitter and receiver, and transmission coefficients are computed through microwave network analyzer (E8362B – 10 MHz to 20 GHz). Metamaterial sample at microwave frequencies are fabricated using UV

lithography commercially. The rotational mount in the measurement setup facilitates the ubiquity measurement, in which the sample can be rotated to any degree.

To verify the reliability of computational and experimental results, benching marking is presented in Fig. 8. The cross- and co-polarization transmissions of an FSS-CMM sample are compared with respect to transient, frequency domain and experimental methods. The FR4 board has a dielectric function of $\varepsilon_r = 4.3 - j0.1075$. Each of the front and back copper layers has a thickness of 12 μm. In FDTD method, CMM structure is meshed with hexahedral meshes with the rate of 50 mesh lines per

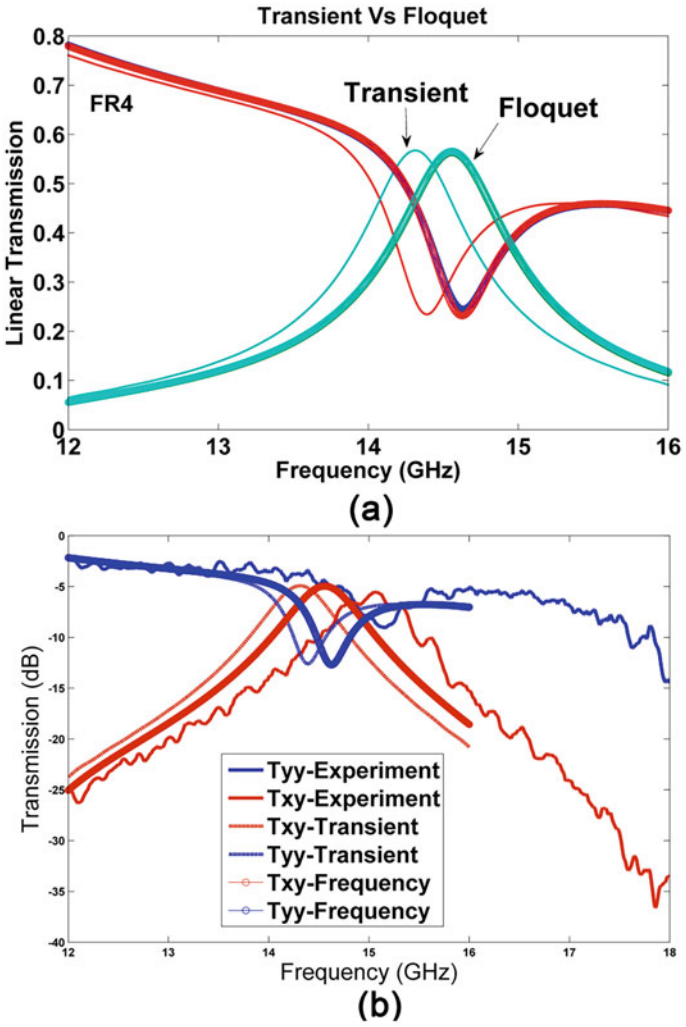

Fig. 8 (**a**) Comparison of transmissions obtained by FDTD and FEM. Co- and cross-polarized transmission coefficients of FSS-MTM polarizer. (**b**) Comparison of FDTD, FEM and experimental results

wavelength. In FEM method, the same CMM structure is meshed with tetrahedral meshes with the rate of 16 steps per wavelength. Figure 8a presents the comparison between FDTD and FEM methods, where slight shift in cross-polarization resonance is witnessed in FDTD results with respect to FEM results. However, the resonant trend has good agreement with each other. In Fig. 8b, cross-polarization transmissions are compared for FDTD, FEM, and experiments.

The FSCMM is fabricated using double-side copper cladding FR4 substrate by UV lithography. The testing sample consists of array of FSCMM unit cells. Optical activity is measured in an anechoic chamber using a microwave network analyzer (E8362B) and stand gain horn antennas as shown in Fig. 9. Cross- and co-polarizations are recorded under ubiquity measurement setup, in which FSCMM sample is rotated for various angles ranging from 0° to 30°.

It is found that FEM results agree closely with experimental ones. This gives a picture of relying on simulation results with respect to experimental one. Even though it is common to witness the frequency shift, it should be noted that FDTD results are employed in field analysis in MTMs that can solve extremely complex structures. It is found from Fig. 9c that the polarization angle at the cross polarization resonance is

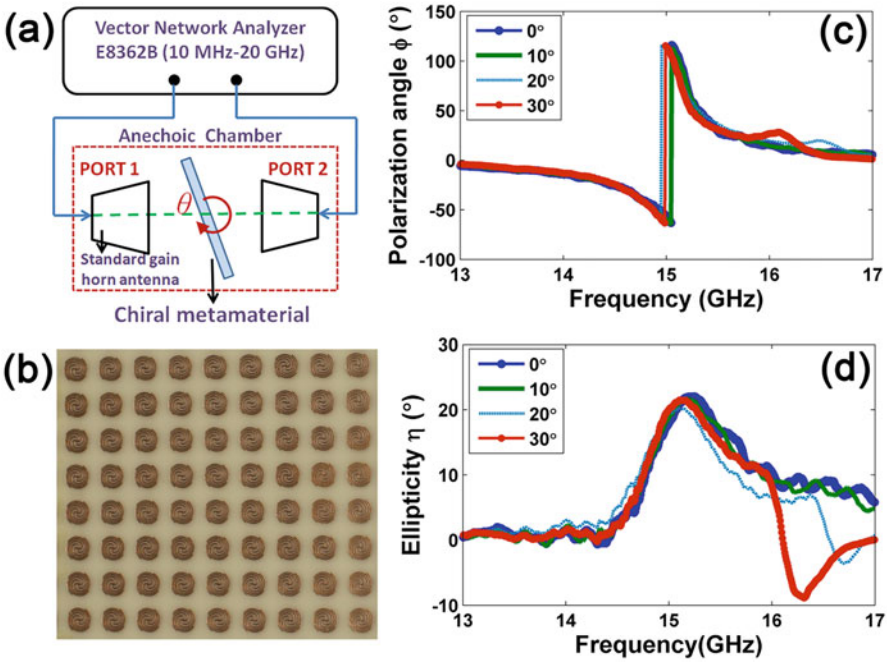

Fig. 9 Experimental verification of optical activity of FSCMM. (**a**) Schematic experimental setup. (**b**) Fabricated array of FSCMM unit cells. (**c** and **d**) Polarization and ellipticity responses for the angular variation of 0° to 30°, respectively

found as 89.47° (at 15.16 GHz), 90.10° (at 15.16 GHz), 90.38° (at 15.10 GHz), and 90.59° (at 15.10 GHz) for the angular variation of 0° to 30°, respectively. For the FSCMM thickness of 1.024 mm, this straight 90° rotation corresponds to the optical activity around 1740°/λ at cross-polarization resonance. On the other hand, the strong dielectric loss of FR4 boards witnesses drastic ellipticity value (Fig. 9d) as 21.88° (at 15.16 GHz), 21.67° (at 15.16 GHz), 20.21° (at 15.10 GHz), and 21.42° (at 15.10 GHz) for an angular variation of 0° to 30°, respectively. Thus, this experimental demonstration verifies the optical activity of the proposed FSCMM under lossy conditions.

Cross-Polarization Rotator MTM FSS

In section "Design Methodologies of MTM FSS Polarizers," the numerical and experimental methods to characterize a typical MTM FSS polarization structure are presented. Over the decade, many groups have explored ultrathin [λ/30] (Ye and He 2010; Rajkumar et al. 2013), unit transmittance (Mutlu and Ozbay 2012), dualband (Shi et al. 2013; Huang et al. 2013), broadband (Wei et al. 2011), and angle-insensitive (Xi 2013) cross-polarization transformers. This section covers major developments reported in the field of cross-polarization converter designs.

Niemi et al. (2013) had explored polarizability model to design 90° polarization rotator using FSS array formed by both reciprocal and nonreciprocal inclusions at microwave frequencies. The condition to realize 90° twist polarizer under symmetric magnetoelectric coupling in reciprocal uniaxial chiral inclusions reads as

$$\eta_0 \alpha_{ee}^{co} = \frac{1}{\eta_0} \alpha_{mm}^{co} = \frac{S}{j\omega}$$
$$\alpha_{em}^{co} = -\alpha_{me}^{co} = \frac{S}{j\omega} A$$

where η_0 is the free space impedance, S is the unit cell area, ω is the angular frequency of the incident e-m field, α_{ee}^{co} and α_{mm}^{co} are the co-polarized electric–electric and magnetic–magnetic polarizability interaction terms, respectively, and α_{em}^{co}, α_{me}^{co} are the co-polarized electric–magnetic and magnetic–electric polarizability interaction terms, respectively. The geometric design, which obeys these conditions is a pair of orthogonal canonical helices. The purpose of orthogonal helices is to nullify anisotropy in polarizabilities. The numerical results reported in Niemi et al. (2013) show that the analytically designed structure has near unity 90° polarization conversion at 1.5 GHz. It is interesting to note that Niemi et al. (2013) have also modified the chiral geometry suitable for printed circuit board and successfully demonstrated 90° polarization rotator at microwave frequencies.

Grady et al. (2013) experimentally realized broadband 90° polarization converter for terahertz frequencies. Their structure comprises of oriented gold strip

sandwiched between gold gratings, and polyimide is used as a dielectric spacer. The separation between the grating and cut-wire is 33 μm. The proposed structure exhibits 90° polarization conversion with 50% transmission efficiency in the frequency range of 0.52–1.82 THz. The peak transmission efficiency reported for the proposed design is 80% at 1.04 THz.

The main challenge in the cross-polarization conversion is the purity of the polarized state, which is measured through ellipticity (η). For a pure linearly polarized wave $\eta = 0$. Low-ellipticity MTM FSS configurations can be realized with supramolecule structure (where a front layer of FSS pattern itself a chiral) as shown in Fig. 10a. The front layer consists of 90° rotated spiral configurations and the back layer is mirror inverted version of the front layer. The T_{yy} and T_{xy} spectra for the designed pattern shown in Fig. 10b reveal that at 7.765 THz, the computed transmissions are $T_{yy} = 0.003$ and $T_{xy} = 0.427$ respectively. Though it is noted that cross-polarization transmission level is around 42.7%, the polarization angle and ellipticity value computed at 7.765 THz is 89.49° and 0° (due to cross-over point in Fig. 10c), respectively. This aspect indicates that using supramolecular designs one can realize low ellipticity cross-polarization converters.

Table 2 is a list of major developments in the designs of cross-polarization converters based on MTM FSS structures.

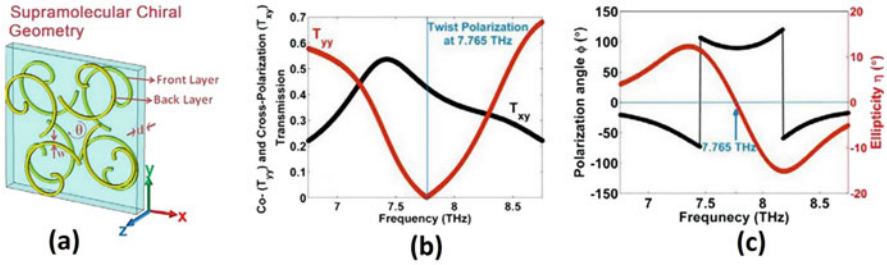

Fig. 10 (**a**) Low-ellipticity cross-polarizer formed by supramolecular Fermat's spiral chiral MTM geometry, (**b**) its co- and cross-polarization resonance spectra, and (**c**) its polarization angle and ellipticity response in the frequency range of 6.5–9 THz, respectively. (Yogesh et al. (2015a) reproduced courtesy of The Electromagnetics Academy)

Table 2 A Few of the notable cross-polarization designs reported in the literature

References	Design structure/method	Regime	Special feature
Khan and Tahir 2018	Anisotropic double slit SRR backed with metal layer/Reflective MS	Microwave/ Experiment	Broad bandwidth 100% (5 GHz to 15 GHz)
Liu et al. (2015)	Single layer complementary MS	Terahertz/ Numerical	Localized surface plasmon polariton resonances
Hu et al. (2017)	Symmetry breaking MS	Infrared/ Experiment	Single layer is sufficient

Circular Polarizer MTM FSS

Conventionally circular polarizer (CP) can be realized based on quarter waveplate (QWP) made of birefringent material, in which QWP produces $\pi/2$ phase difference between the parallel and perpendicular components of the electric field vector of the incident e-m field so that a linearly polarized wave is converted into a circularly polarized wave. One can directly adopt waveplate concept to design circular polarizers using MTM FSS structures.

However, it is worth to mention that there are many methods of CP designs available in the literature. Meanderline structures are one of the important categories of CP devices, in which inductive and capacitive effects of metal patterns provide necessary phase shift to produce CP wave (Young et al. 1973; Strikwerda et al. 2009). Helices are the foremost artificial pattern used for the production of CP wave, as Lindman (1920) demonstrated its CP functionality at microwave frequencies. Over the decade many groups explored helical MTM structures for the CP designs, which exhibits broadband CP conversion, strong circular dichroism (Decker et al. 2007; Silveirinha 2008; Gansel et al. 2009; Li et al. 2011a, b; Zhao et al. 2012). Chiral Photonic crystals in the form of chiral nets (Saba et al. 2011), rotational stacks (Thiel et al. 2007; Takahashi et al. 2014), and bi-chiral systems (Theil et al. 2009) were employed for the realization of CP devices.

There are various designs of MTM FSS structures such as U-shaped split-ring resonators (Li et al. 2010; Mutlu et al. 2011), wheel patterns (Ye et al. 2011), twisted double SRRs (Yan and Vandenbosch 2013), Hilbert tiles (Xu et al. 2013a), bisected arc (Martinez-Lopez et al. 2014), connected squares (Xu et al. 2013b, bilayered Fermat's spiral MTM (Yogesh et al. 2015b), of which three fundamental design methods are indicated.

Martinez-Lopez et al. (2014), designed LP to CP converters by stacking bi-sected SRR-based MS structures. In their design, when an incident wave with linear polarization of 45° is incident on MS stacking, it produces CP wave around K-band frequencies. According to their design, bi-sected SRR stacking acts as a kind of quarter waveplate, where the stacking gives necessary phase shift between parallel and perpendicular components of the incident E-field vector.

Naseri et al. (2020), designed a dual band LP to CP convertor using equivalent circuit model (ECM). ECM method allows one to design MTM FSS polarizer structure based on LCR circuit and derive expressions for mutual inductance and capacitances aroused from longitudinal and transverse interaction of MTM patterns. Several anisotropic effects in MTM can be successfully modeled out of ECM method. Though exact mathematical form for inductances and capacitances may not be available to model several interactions of MTMs unit cells, ECM method can serve as a semi-analytical model to design desired response and this method can serve as a footprint for artificial intelligence-based polarization elements design in the present and near future.

In our work, we had designed LP to CP conversion in bilayered Fermat's spiral chiral MTM structure (Yogesh et al. 2015b), as shown in Fig. 11a. The design procedure invokes chirality breaking where in the front and back layers of MTM

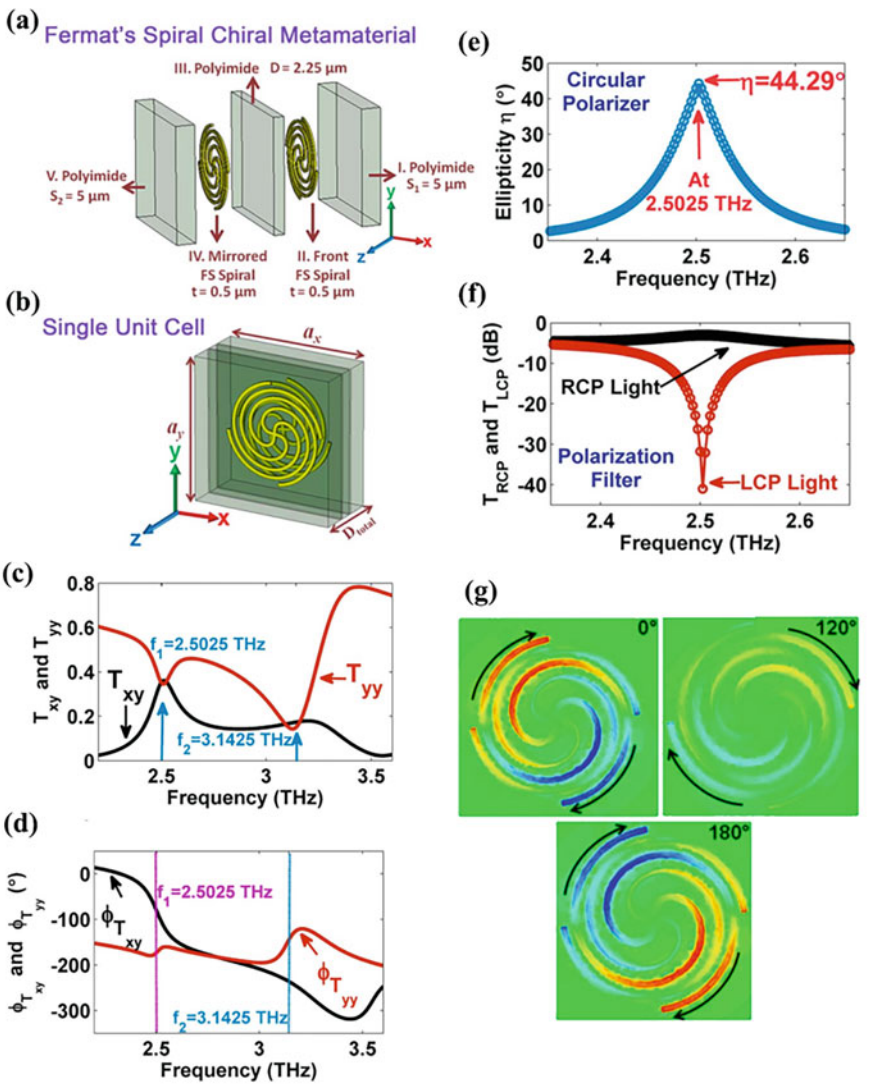

Fig. 11 Circular polarizer design at THz frequencies (Yogesh et al. 2015b). (**a** and **b**) realized bilayered Fermat's spiral chiral metamaterial and its unit cell, respectively. (**d** and **e**) its co- and cross-polarization transmissions and phase spectra, respectively. (**e** and **f**) ellipticity and CP transmission coefficients of Fermat's spiral CMM, respectively, and (**g**) electric dipole orientation under phases for LP incident wave at 2.505 THz. (Yogesh et al. 2015b © J. IEEE Photonics 2015)

structure (Fig. 11a, b), mirror symmetry is broken along the direction of propagation of e-m field. In the proposed structure, at two different frequencies (2.5025 THz and 3.1425 THz), the condition for CP conversion is met. Especially at 2.5025 THz, $T_{xy} = T_{yy} \simeq 0.35$ (Fig. 11c) and the phase difference between T_{xy} and T_{yy} is 90.32°

(Fig. 11d). The ellipticity computed at 2.5025 THz is 44.29° (Fig. 11e). Hence the designed CMM converts LP to RCP wave at 2.5025 THz. Moreover, the designed CMM acts as a circular polarization filter (Fig. 11f), in which it allows RCP to pass through whereas LCP wave is completely filtered out of CP resonance with the isolation efficiency of over 40 dB. The mirror symmetry breaking CMM exhibits strong anisotropic effects in which magnetic dipole interaction (Fig. 11g) with external e-m field enables CP conversion functionality in the realized design.

Asymmetric Transmission for Linearly Polarized Waves in MTM FSS

Asymmetric Transmission (AT) is the property of a material in which the wave propagation is allowed in one direction but restricted in the opposite direction. This property was well established in nonreciprocal medium such as Ferrites and magneto-optical structures. In MTM FSS, AT effect is easily achieved without any requirement of nonreciprocal medium and external magnetic field as in the case of ferrites. Fedotov et al. (2006) first demonstrated AT in MTM with planar chiral structures. Since then AT in MTM structures has been widely explored from microwave to visible frequencies (Huang et al. 2012; Menzel et al. 2010; Li et al. 2014b; Mutlu et al. 2011; Kang et al. 2011; Liu et al. 2015; Dincer et al. 2013; Chen et al. 2017; Dong et al. 2016; Pavithra et al. 2022).

The AT for LP waves in MTM FSS refers to that the conversion of TE to TM (vice versa) is not the same for wave incidence from forward/backward direction. As mentioned in section "Origin and Mechanism of Optical Activity in Artificial MTM Structures," the magnitude of linear AT is expressed as $\Delta^{AT}_{\text{linear}} = \left|T_{xy}\right|^2_{\text{forward}} - \left|T_{yx}\right|^2_{\text{forward}} \neq 0$ where T_{xy} (T_{yx}) is the TE to TM (TM to TE) cross-polarization transmission coefficient when TE (TM) wave is excited from forward direction of the MTM FSS, respectively.

In general, induced electric and magnetic dipole moments in a single MTM unit cell FSS are written as (Asadchy et al. 2018)

$$\vec{p} = \overline{\overline{\alpha_{ee}}} \cdot \vec{E}_{\text{local}} + \overline{\overline{\alpha_{em}}} \cdot \vec{H}_{\text{local}},$$
$$\vec{\mu} = \overline{\overline{\alpha_{me}}} \cdot \vec{E}_{\text{local}} + \overline{\overline{\alpha_{mm}}} \cdot \vec{H}_{\text{local}},$$

where $\overline{\overline{\alpha_{ee}}}, \overline{\overline{\alpha_{em}}}, \overline{\overline{\alpha_{me}}}$ and $\overline{\overline{\alpha_{mm}}}$ are the electric, magnetic, electromagnetic, and magnetoelectric susceptibility of an MTM FSS unit cell, respectively. Here $\vec{E}_{\text{local}}$ and $\vec{H}_{\text{local}}$ are local electric and magnetic fields due to the sources and currents generated from other unit cells to excite the particular unit cell in MTM FSS. Anisotropic interaction, in which magnetodielectric (or electromagnetic) susceptibility tensor of an MTM is nonzero, is requirement for AT for LP waves in MTM FSS.

For instance, Fig. 12 accounts AT for LP waves in MTM FSS at three different frequency regimes. In Fig. 12a, MTM FSS corresponds to telecommunication

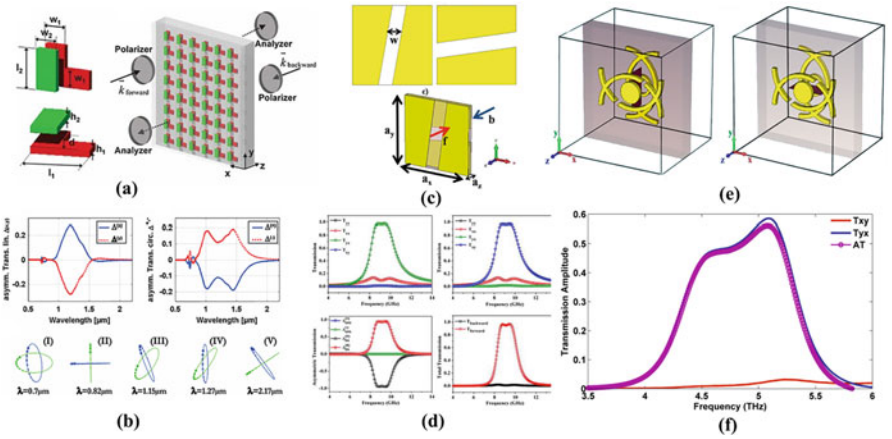

Fig. 12 AT for LP waves in MTM structures. (**a**) Designed geometry and (**b**) its AT spectra for LP waves of Menzel et al. (2010) © American Physical Society, operating at telecommunication frequencies. (**c**) MTM unit cell geometry and (**d**) its AT spectra for LP waves of Stephen et al. (2018) © American Institute of Physics, operating at microwave frequencies. (**e** and **f**) MTM unit cell geometry and (**d**) its AT spectra for LP waves of Pavithra et al. (2019) (reproduced courtesy of The Electromagnetics Academy), operating at THz frequencies

frequency (Menzel et al. 2010). It comprises of L and single cut wire pattern so that when light is excited from FW/BW directions, the unit cell results in different current distributions due to anisotropy. However the entire unit cell is reciprocal, where the magnitude of AT is the same for both FW/BW excitations. The MTM FSS shown in Fig. 12c exhibits broadband AT for LP waves with the magnitude of over 0.9 at microwave frequency range (from 8.67 GHz to 9.82 GHz) (Stephen et al. 2018). The MTM FSS design shown in Fig. 12d shows AT for LP waves in the terahertz frequency range (Pavithra et al. 2019).

Asymmetric Transmission for Circularly Polarized Waves in MTM FSS

Similar to AT for LP waves, AT for CP waves can also be achieved in an MTM FSS, which possesses anisotropy (Wu et al. 2013; Pfeiffer et al. 2014; Amaljith et al. 2020). In the work of Wu et al. (2013) a bilayered MTM unit cell formed by asymmetric splitting ring resonator (Fig. 13a) was used to achieve AT for CP waves with the magnitude of 0.6 at microwave frequency (Fig. 13b, c). Pfeiffer et al. (2014), explored the AT for CP waves at visible frequency. According to their structure design (Fig. 13d), an RCP wave is converted into an LCP wave when the RCP wave is incident from left direction whereas when the RCP wave incident from right direction is completely reflected. Their structure design consists of cascaded tiles (Fig. 13d), where the metal pattern in each tile is designed for specific sheet admittance using equivalent circuit model. Numerical and experimental AT spectra

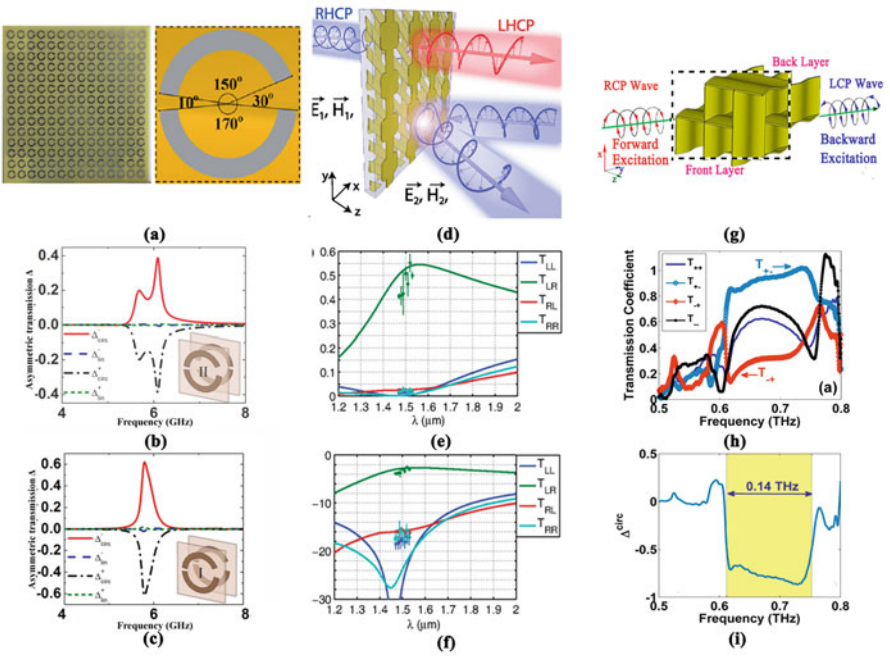

Fig. 13 AT for CP waves in MTM structures. (**a**) MTM formed by bilayered asymmetric SRRs, and its single unit cell, and (**b** and **c**) AT spectra for CP waves of inverted bilayered asymmetric SRR and mirror-symmetric bilayered asymmetric SRR unit cells at microwave frequencies (Wu et al. (2013), ©American Institute of Physics), respectively. (**d**) Realized optical metasurface geometry, and the numerical and experimental AT spectra (**c** and **d**) for CP waves, respectively (Pfeiffer et al. (2014) ©American Physical Society). (**e**) Unit cell geometry 3-D spline assembly realized at THz frequencies, (**f**) its CP transmission coefficients, and (**i**) its AT spectra for CP waves (Amaljith et al. (2020), ©Optica Publishing)

of cascaded tiles shown in Fig. 13e, f verified AT for CP waves at visible frequencies. Modeling sheet admittance for specific polarization functionality is one of the novel things of MTM FSS polarizers. In the work of Amaljith et al. (2020), a 3-D spline assembly (Fig. 13g) breaks mirror symmetry along *x*, *y*, and *z* axes and exhibits strong anisotropic AT for CP waves (Fig. 13h, i) along all three principal axes at THz frequencies.

Applications of MTM FSS Polarizers

The various polarization phenomena observed in MTM FSSs can be employed for controlling and manipulating the radiation pattern of antenna systems in communication, stealth, and RADAR applications. Figure 14 demonstrates two important applications of MTM FSS polarizers. According to the work of Zhang et al. (2020), a reconfigurable polarization conversion metasurface (PCM), which converts LP to

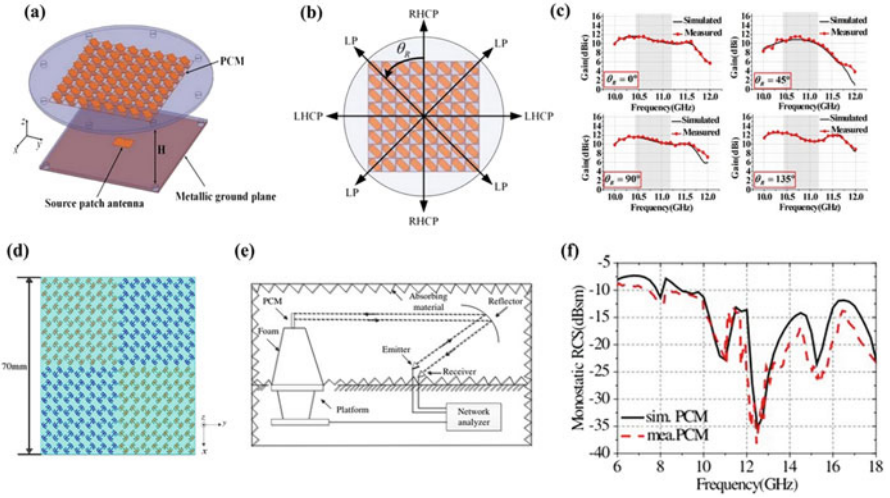

Fig. 14 (**a**) Configuration used for improving the gain of a microwave patch antenna, (**b**) geometry of polarization conversion metasurface with different orientation angles, and (**c**) its corresponding gain spectra (both numerical and experimental) (Zhang et al. 2020), reproduced courtesy of The Electromagnetics Academy). (**d**) MTM FSS proposed for Radar cross-section reduction, and (**e** and **f**) experimental setup and obtained results, respectively (Jiang et al. (2016), reproduced courtesy of The Electromagnetics Academy)

RCP, LP to LCP, and LP to orthogonal LP state based on the mechanical rotation of a PCM, can not only be able to give different polarization state for a patch antenna, but also improve the gain of a patch antenna when PCM is placed above it as shown in Fig. 14a. Figure 14b depicts how the rotation angle (θ_R) of PCM yields different polarization states where $\theta_R = 0^\circ$and 90° corresponds to RCP and LCP states, respectively, and $\theta_R = 45^\circ$and 135° corresponds to two different LP polarizations, respectively. The gain of patch antenna and PCM for four different $\theta_R s$ is given in Fig. 14c. It is observed that the gain of a PCM employed patch antenna is above 10 dBi at polarization conversion bandwidth whereas the patch antenna without PCM exhibits the gain of around 6 dBi to 8 dBi at the polarization conversion frequencies. The mechanism for gain improvement by PCM is attributed by the fact that PCM also acts as a partial Fabry-Perot resonator, in which multiple reflection of e-m radiation between patch antenna and PCM leads to gain enhancement.

Another straightforward application of MTM FSS polarizers is their circular polarization functionality. A CP antenna has good reception characteristics so that an MTM FSS structure finds as an inevitable component in antenna designs for microwave and space applications.

A second significant application of MTM FSS polarizers is their ability to reduce RADAR cross-section, also known as RADAR cross-section reduction (RCSR). Minimizing RADAR cross-section of any scattering object reduces the chance of detecting the object by antennas. This aspect of RCSR is useful for stealth technologies, in which an object under concealment will be coated or covered with an

absorbing material conventionally. Using MTM FSS, Jiang's et al. (2016), demonstrated broadband RCSR with the PCM geometry shown in Fig. 14d. According to their design, A PCM, which converts LP to orthogonal LP state (TE to TM or vice versa), is arranged with mirrored PCM in checker board fashion (Fig. 14d), so that 180° out of phase reflected LP waves from PCM, and its mirrored PCM will cancel each other, and the geometry (Fig. 14d) results in zero or minimum reflection at polarization conversion frequencies. The experimental setup shown in Fig. 14e can be useful to characterize MTM FSS for RCSR application. The simulated and experimental results of Jiang's et al. (2016), given in Fig. 14f, show that MTM FSS has at least 10 dB RCS reduction over a frequency bandwidth of 9.4 GHz–19.2 GHZ with respect to metal ground plane. From Fig. 14f one can note that PCM arrangement can exhibit maximum RCSR of 32 dB around 12.5 GHz.

Coded Metasurface (CM) is a new avenue to control radiation of pattern of antenna systems (Cui 2017). In CM literature, one can explore polarization feature of MS patterns. In the work of Varsha et al. 2019, an MS unit cell exhibiting AT for LP waves (Fig. 15a, b) is arranged with different coded patterns as shown in Fig. 15c, d. The coded patterns exhibit different polarization state as the computation of polarization angle given in Fig. 15e, f are significantly different for the geometries shown in Fig. 15c, d, respectively.

Generating e-m beams with different orbital angular momentum states such as radial and azimuthal polarization states using MTM FSS polarizing structure (Bijun et al. 2016) are also quite significant in vortex optics (Zhao et al. 2013).

Moreover, extension of MTM FSS polarizer designs to optical frequencies empowers the measurement of polarization techniques available for optical systems. It is noted that a separate literature review (Intravanne and Chen 2020) is available on how optical MTM FSS polarizers developed over this decade for the measuring and controlling polarization states.

Conclusions

In this chapter, how the metamaterial (MTM) frequency selective surface (FSS) structures serve as polarizers is presented. The polarization functionalities of MTM FSS including linear polarization, circular polarization, and asymmetric transmission for linear and circularly polarized waves are discussed. The origin and mechanism of polarization functionality of MTM FSS structure are presented with respect to dipole interaction picture and anisotropic effects in MTMs. Numerical methods in realizing MTM FSS polarizers such as FDTD and FEM are elucidated. A detailed characterization of MTM FSS cross-polarizer is presented at microwave frequencies including the necessary experimental setup. Based on the MTM FSS polarizers, applications such as improving the gain of microwave patch antenna, reducing the radar cross-section, and creating e-m radiation with different polarization states using coded metasurfaces are discussed.

The following aspects are the avenues available for the exploration of MTM FSS polarizers further: In designing tuneable and reconfigurable MTM FSS polarizers,

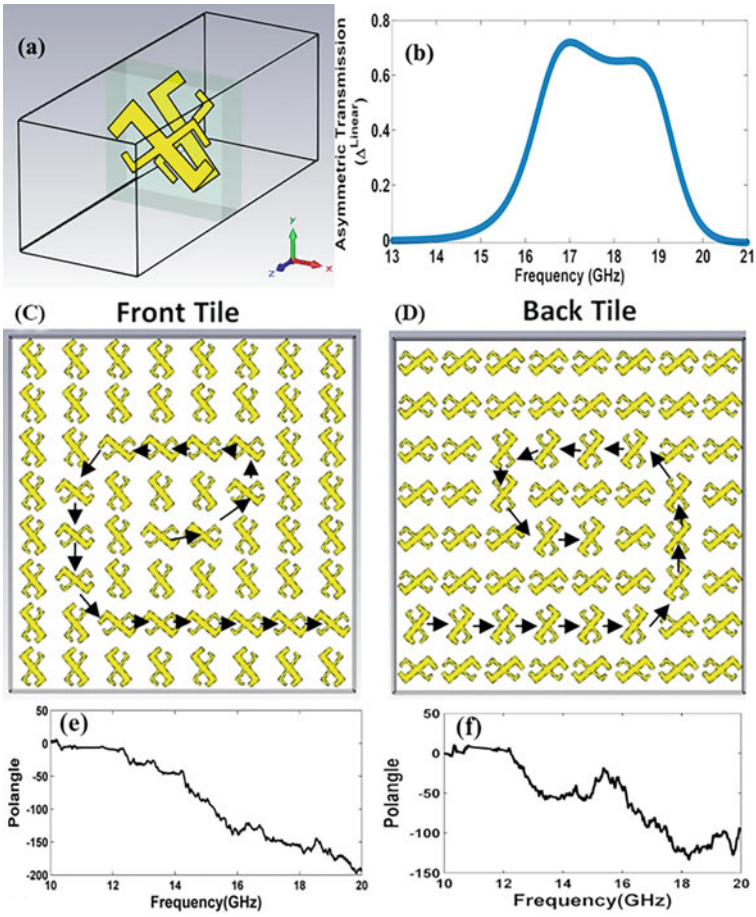

Fig. 15 Employing AT effect of MTM for designing coded metasurface. (**a**) Proposed bilayer unit cell and (**b**) its AT spectra for LP waves, respectively. (**c** and **d**) selective designs of coded metasurface. (**e** and **f**) its polarization angle spectra, respectively. (Varsha et al. (2019), reproduced courtesy of The Electromagnetics Academy)

researchers are exploring novel materials across e-m spectra. Ferroelectrics, ferrites, and piezo electric materials serve as possible candidates for microwave tuneable devices. Similarly, bulk-Dirac semimetal (Blessan and Yogesh 2021) and graphene (Xiong et al. 2020) serve as potential candidates for the design of tuneable devices at THz and visible frequencies, respectively.

The advent of 3-D printing enables one to realize 3-D MTM FSS polarizing structures at microwave/THz frequencies. Involving machine learning algorithm in designing MTM FSS polarizers is another thrust research area. Exploring adaptive optics devices for wavefront control using MTM FSS polarizer is equally an interesting avenue for microwave and optical researchers. Most of the MTM FSS polarizers are designed using semiclassical approximation where e-m radiation with

very high photons is considered. Single photon interacting with MTM FSS polarizer is another interesting theoretical and experimental area for quantum optics researchers for the applications in quantum computing and quantum communication.

Since the polarization state of e-m radiation also comprises of time dynamics, it is anticipated that MTM FSS polarizer will serve as probing tool for the study of polarization time-dynamics in condensed matter, proteins, and biological molecules.

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Metamaterial Based Microwave Coupler Design

9

Arun Kumar Gande and Gopi Ram

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Abstract

The design of microwave coupler in microstrip technology is complicated due to fabrication limitations. Many technologies are employed to overcome these limitations and enhance the performance of the couplers. The analysis of the coupler using equivalent lumped Lange coupler (LC) circuit configuration for conventional coupler and metamaterial loaded coupler is presented. The extraction process for the lumped element values of the coupler is also explained in detail. The enhancement in the fabrication limits is elaborated using the loaded metamaterial LC circuit. The basic microstrip Lange coupler design is explained and the equivalent circuits are substantiated with circuit simulation, High Frequency Simulation Software (HFSS) simulation, and measurement results.

Keywords

Coupler · Defected ground · Directivity · Even mode · Metamaterial · Odd mode · Left-handed transmission line

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Introduction

Couplers form an integral part of many communications systems. They are four-port components that sample a part of the power in the mainline to the adjacent line. The port on which the power is incident is called the input port; the mainline output, which contains most of the power, is the through port. The sampled power port is the one wherein a small fraction of the input power is transferred to the coupled port, and the remaining port which does not receive any power from the input port is the isolated port. The generalized network for the coupler is shown in Fig. 1. If the power coupled is smaller, i.e., 1/10th or 1/100th part of the power is coupled, then they are called directional couplers. If the power coupled is half, then they are known as 3 dB hybrids. Depending on the configuration of these couplers, the output ports, i.e., the through port and the coupled ports, have either 90° or 180° phase differences.

These couplers are used in the design of balanced amplifiers, mixers, modulators, and for power splitting and combining (Pozar 2005; Bahl and Bhartia 2003; Lai et al. 2008; Kim et al. 2009; Jin and Hsu 2008). Though there are couplers that provide 3 dB power division, such as branch-line coupler, rat-race coupler, etc., they are not compact, do not provide any DC isolation, and offer narrow bandwidths. Techniques for bandwidth and performance improvement of these couplers are being investigated by researchers (Paco et al. 2008; Eccleston and Ong 2003). Coupled line couplers are broadband, but the distance between the coupled lines is prohibitively small for PCB fabrication for increasing coupling levels. To enhance the coupling level of these couplers, Electronic Band Gap (EBG) structures are proposed (Chudzik et al. 2011); though this is a simple design methodology, the design is not compact. Tight coupling techniques for the design of components such as broadside coupled lines (Ma and Cheng 2009), re-entrant couplers (Chiang and Ma 2008), and tandem couplers (Cho et al. 2005; Abe et al. 2010) suffer from narrow coupling gaps, long coupling lengths, etc. It makes their usage in compact design absurd due to the tolerance limit of in-house fabrication processes. Miniaturization of broadside coupler using artificial slow-wave lines has been proposed, but these involve multilayer structures, increasing the complexity of the designs (Ma and Cheng 2009). Lange coupler is widely used for broadband applications and has the advantages of smaller size and large bandwidth (Lai et al. 2008; Kim et al. 2009; Lange 1969; Osmani 1981). This coupler also suffers from narrow line widths and coupling gaps. The amount of coupling can be increased by increasing the number of fingers, but this decreases the line width of the fingers. This problem is aggravated at microwave and millimeter-wave frequencies where the substrate thickness is lower, thus decreasing the widths of the line and the gap between the lines.

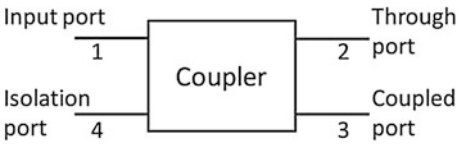


Fig. 1 4-port network of a generalized coupler

In view of the aforementioned discussion and PCB fabrication limitations, this chapter will design a patterned ground Lange coupler. Equivalent circuit analysis of the patterned ground coupler based on superposition principle, even and odd symmetry are carried out to investigate the effects of loading the coupler. All Electromagnetic (EM) simulations are performed using Commercial Finite Element Method (FEM) solver Ansys HFSS^{TM}. It is observed that the patterned ground structures increase the coupling level by increasing the self and mutual inductance of the Lange coupler.

Odd and Even Modes in Couplers and Parameters

A symmetrical directional coupler can be described by the S-parameter matrix as shown in (1), and the important parameters of the coupler are return loss, coupling coefficient, isolation, and directivity. Return loss (RL) characterizes the amount of power that is reflected at the port, coupling (C) represents the fraction of power coupled to the output port, and directivity (D) is a qualitative benchmark for couplers and provides the ability of the coupler to isolate forward and backward waves. These parameters are defined as

$$[S] = \begin{bmatrix} S_{11} & S_{21} & S_{31} & S_{41} \\ S_{21} & S_{11} & S_{41} & S_{31} \\ S_{31} & S_{41} & S_{11} & S_{21} \\ S_{41} & S_{31} & S_{21} & S_{11} \end{bmatrix} \tag{1}$$

$$RL = -20\log|S_{11}|\text{dB} \tag{2}$$

$$T = 101\log\left(\frac{P_1}{P_2}\right) = -20\log|S_{21}|\text{dB} \tag{3}$$

$$C = 10\log\left(\frac{P_1}{P_3}\right) = -20\log|S_{31}|\text{dB} \tag{4}$$

$$I = 10\log\left(\frac{P_1}{P_4}\right) = -20\log|S_{41}|\text{dB} \tag{5}$$

$$D = 10\log\left(\frac{P_3}{P_4}\right) = I - C \tag{6}$$

Where P_1, P_2, P_3, P_4 are the power at the ports 1, 2, 3, and 4, respectively, as shown in Fig. 1. T is the transmission coefficient, and I is the isolation.

Couplers can be implemented in various transmission structures such as waveguide, strip-line, suspended strip-line, and microstrip line. However, microstrip line implementation has gained importance due to its planar form and ease of integration with other integrated circuits. Considering two closely placed parallel

microstrip lines, power is transferred from one line to the adjacent line due to the electromagnetic field coupling between them. These coupled lines support two modes of propagation with different characteristics impedances. The propagation velocity of the two modes is the same if the medium is homogenous and different if the medium is not homogenous. In the case of microstrip lines, the fields travel in the dielectric medium of the substrate and air; hence, the medium is inhomogeneous, and the velocity of propagation of the two modes is different. The two modes are defined as odd and even modes of propagation. These are shown along with the field configurations in Fig. 2. It is observed from Fig. 2a that same voltage resides over the two microstrip lines, and hence, the electric field has even symmetry; this mode is known as even mode excitation. The electric fields lines are perpendicular to the microstrip lines, and this field configuration results in a magnetic wall. Odd mode excitation results in an odd symmetry and voltage applied to the two microstrip lines are 180° out of phase, resulting in an electric wall between the lines. The field configurations for the odd mode are shown in Fig. 2b. The equivalent circuits for the even and odd modes are also shown in Fig. 2. The even mode circuit consists of capacitances to the ground (C_{11} and C_{22}) from the microstrip line. The odd mode consists of capacitances to the ground (C_{11} and C_{22}) and also mutual capacitance between the microstrip lines ($C^{1}{}_{2}$). The even and odd mode impedances for the coupled microstrip lines are given as Z_{0e} and Z_{0o}. The characteristic impedance (Z_0) and coupling coefficient (C) of the coupled microstrip lines are defined as in Eq. 9 and 10.

$$Z_{0e} = \frac{1}{vC_e} \quad \text{where } C_e = C_{11}, \quad v \text{ is velocity of the wave in medium} \tag{7}$$

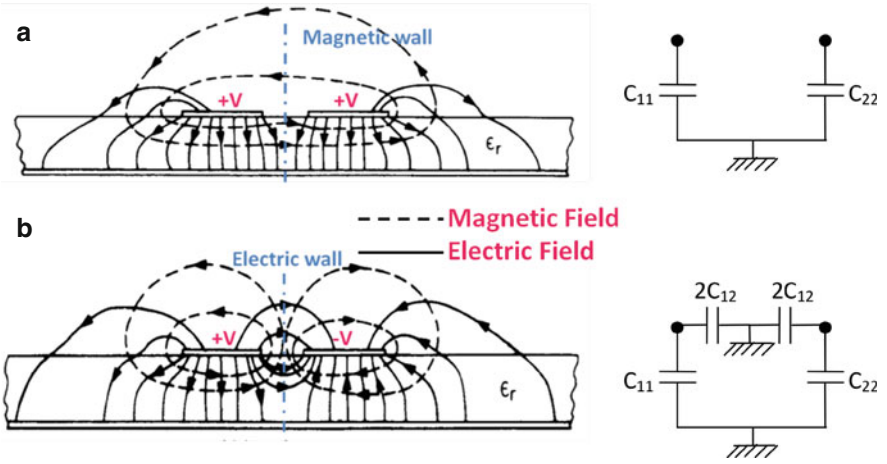


Fig. 2 (**a**) Electric and magnetic field distribution for even mode and its equivalent circuit (**b**) Electric and magnetic field distribution for odd mode and its equivalent circuit [1]

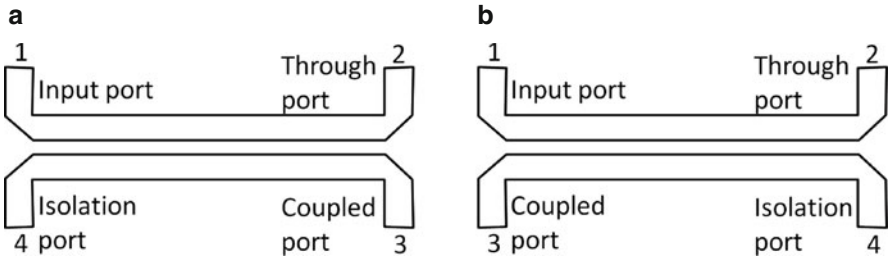


Fig. 3 (**a**) Forward-wave coupled line coupler port orientation (**b**) Backward-wave coupled line coupler port orientation

$$Z_{0o} = \frac{1}{vC_o} \quad \text{where } C_o = C_{11} + 2C_{12} \tag{8}$$

$$Z_0 = \sqrt{Z_{0e}Z_{0o}} \tag{9}$$

$$C = \frac{Z_{0e} - Z_{0o}}{Z_{0e} + Z_{0o}} \tag{10}$$

In microstrip line configuration, the couplers can be designed in two different coupling methods, i.e., forward coupling and backward coupling. In backward-wave coupling, the coupling is dependent on the difference between the even and odd mode impedances assuming equal phase velocities for the two modes. In forward coupling, it is due to accumulated phase difference between even and odd equal characteristic impedances for both the modes. The port orientations in forward-wave and backward-wave couplers are shown in Fig. 3.

Limitations in the Practical Implementation of Microwave Couplers

Microwave couplers are mostly implemented in microstrip technology. This results in certain practical limitations during implementation, especially in losses, directivity, and physical dimensions of the couplers. The losses in microstrip lines are due to ohmic losses, dielectric losses, and radiation losses. The ohmic losses arise due to the finite conductivity of the metals such as copper. Dielectric losses are due to the substrate and are dependent on the substrate material and its construction. Radiation losses occur due to fringing fields in microstrip lines; these fields travel through two different mediums, i.e., air and substrate. Dispersion is seen in microstrip lines, resulting in the variation of phase velocity of the even and odd modes. Any variation in the phase velocities of even and odd modes results in the degradation of directivity of the coupler. The limitation on the physical dimensions of the coupler occurs when the tight coupling is required. Coupling between the two parallel coupled lines is dependent on the gap between the lines. The smaller the gap greater is the field coupling between the lines. The gap can be relaxed by adding additional parallel

lines to the coupler, i.e., increasing the number of fingers. However, this will decrease the width of the lines. The width of the lines can be increased by increasing the height of the substrate. The substrate height determines the radiation losses and dielectric losses, which would increase with the height. These limitations can be overcome by using patterned or defected ground. The patterned/defected ground introduces a slow wave phenomenon, reducing the even mode and odd mode capacitance. The reduction in the even mode capacitance is higher than the odd mode capacitance. This will help maintain the even and odd mode impedances closer to relax the dimensional tolerances and achieve higher directivity. This method has a disadvantage of radiation losses due to patterned/defected ground. To reduce the radiation losses and also maintain the performance of the couplers, metamaterials are employed. Metamaterial structure being very small compared to the signal's wavelength, the radiation due to these structures when etched on the ground plane would be much lesser than patterned/defected ground structure. However, due to their exotic properties, they load the coupler to provide the effect of slow-wave structure in compact dimensions.

Equivalent Circuit Analysis and Simulation of Metamaterial Couplers

The physical dimensions of a Lange coupler can be relaxed by loading the Lange coupler with a metamaterial structure. The analysis and design process of the Lange coupler is discussed in this section. Lange coupler has been considered an inductively coupled coupler. The equivalent circuit analysis of the inductively coupled coupler presented by Frye et al. (2003) and Ozis (2009) is adopted for the analysis of the Lange coupler. The equivalent circuit of the Lange coupler is shown in Fig. 4. The conventional Lange coupler (LC) is initially analyzed by considering the even and odd mode impedances (Z_{0e} and Z_{0o}) about the reference planes P1 and P2, resulting in four conditions. As discussed earlier, the magnetic and electric wall conditions are applied at the two reference planes P1 and P2. The magnetic boundary wall condition results in an open circuit, and the electric boundary condition results in a short circuit. The conditions and their equivalent circuits are shown in Table 1. Cg and Cp are the mutual and the self-capacitances in the equivalent circuit, and L_e and L_o are the even and odd mode inductances. M is the mutual inductance. The input reflection coefficient for the four conditions is listed in Table 1. The

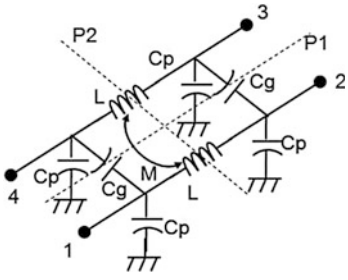
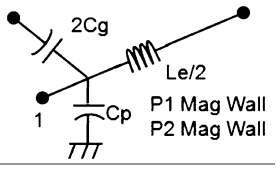
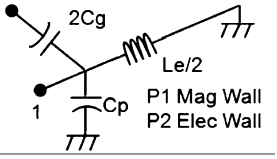
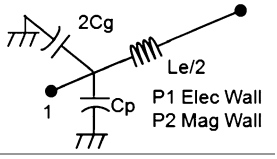
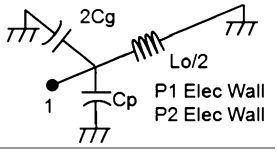


Fig. 4 Equivalent Circuit of Conventional Lange coupler

Table 1 Four cases resulting from two references Planes P1 and P2

Case 1	P1	P2	Input admittance	Reflection coefficient	Equivalent circuit
I	MW	MW	Y_{mm}	$\Gamma_{mm} = \frac{Y_0 - Y_{mm}}{Y_0 + Y_{mm}}$	
II	MW	EW	Y_{me}	$\Gamma_{me} = \frac{Y_0 - Y_{me}}{Y_0 + Y_{me}}$	
III	EW	MW	Y_{em}	$\Gamma_{em} = \frac{Y_0 - Y_{em}}{Y_0 + Y_{em}}$	
IV	EW	EW	Y_{ee}	$\Gamma_{ee} = \frac{Y_0 - Y_{ee}}{Y_0 + Y_{ee}}$	

MW magnetic wall, *EW* electric wall

S-parameters are calculated using the input reflection coefficients as shown in Eqs. 11, 12, 13 and 14.

$$S_{11} = \frac{\Gamma_{mm} + \Gamma_{me} + \Gamma_{em} + \Gamma_{ee}}{4} \tag{11}$$

$$S_{21} = \frac{\Gamma_{mm} - \Gamma_{me} + \Gamma_{em} - \Gamma_{ee}}{4} \tag{12}$$

$$S_{31} = \frac{\Gamma_{mm} + \Gamma_{me} - \Gamma_{em} - \Gamma_{ee}}{4} \tag{13}$$

$$S_{41} = \frac{\Gamma_{mm} - \Gamma_{me} - \Gamma_{em} + \Gamma_{ee}}{4} \tag{14}$$

The analysis of the loaded Lange coupler is based on the superposition principle. The equivalent circuits of the Lange coupler and the loading structures are evaluated separately, and then they are combined to get the response of the loaded Lange coupler. The equivalent structure of the dumbbell has been proposed by Kim et al. (2000) and Chen and Xu (2010), and it consists of a parallel resonant structure of L_r and C_r. The equivalent structure of the dumbbell-loaded Lange coupler (LCDB) is as shown in Fig. 5a, and two dumbbell structures are used for symmetry. The equivalent circuit for the split ring resonator has been proposed by Falcone et al. (2004) and Luo

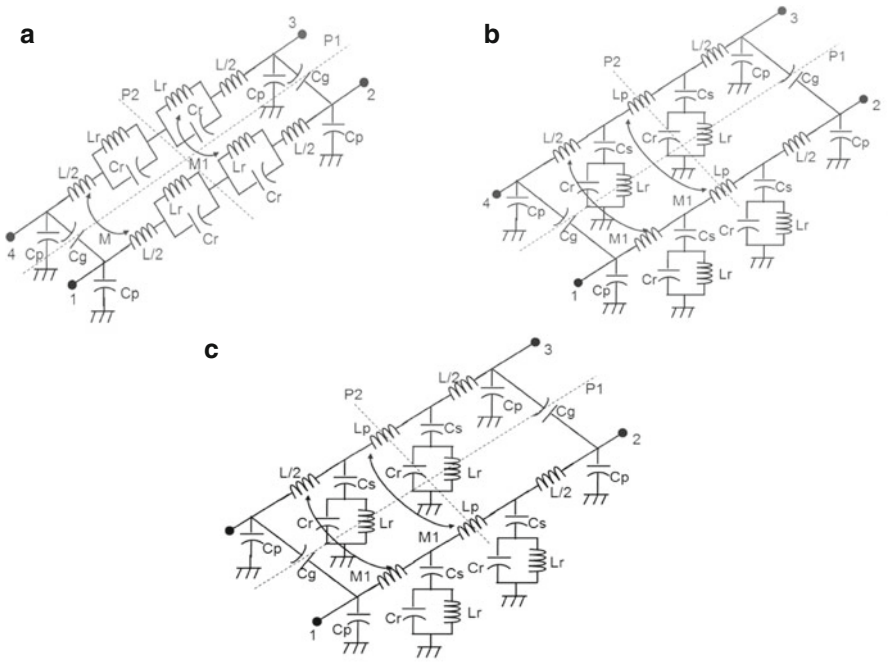


Fig. 5 Equivalent Circuit of (**a**) LCDB (**b**) LCSRR (**c**) LCDOS

et al. (2010). This equivalent circuit is used to analyze the Split Ring Resonator loaded Lange coupler (LCSRR). It consists of series Inductance (L_p) and shunt capacitance (C_s) representing the loaded line and the resonant circuit (C_r and L_r) representing the split ring resonator. The equivalent circuit of LCSRR is shown In Fig. 5b. The equivalent circuit of deformed omega structure has been proposed by R. K. Mishra et al. (2012). The equivalent circuit of the deformed omega-loaded Lange coupler (LCDOS) is shown in Fig. 5c. It is similar to the equivalent circuit of a split-ring resonator. The inductance in the series arm (L_p) and the shunt capacitance in the shunt arm (C_s) are due to the interaction of the microstrip line with the structure, and the parallel resonant structure (C_r and L_r) is due to the resonant behavior of the deformed omega structure.

The input admittance equations for the above three loaded Lange couplers are given in Table 2. L_{pe} and L_{po} are the even and odd mode inductances. The circuit's input admittances and reflection coefficients about the two reference planes are calculated, and the S-parameters are evaluated from the reflection coefficients.

Extraction of Equivalent Circuit Parameters

The above equivalent circuit analysis has been evaluated using a circuit simulator. The equivalent circuit element values have been obtained from the respective circuits loading a 50 Ohm microstrip line. For this purpose, the ground plane underneath the 50 Ohm

Table 2 Four cases resulting from two references Planes P1 and P2

Input admittance	LCDB	LCSRR/LCDOS
Y_{mm}	$j\ \omega\ C_p$	$j\omega C_p + \frac{1}{\frac{1}{j\omega C_S}+\frac{j\omega L_r}{1-\omega^2 C_r L_r}+\frac{j\omega L_e}{2}+\frac{j\omega L_{pe}}{2}}$
Y_{me}	$j\omega C_p + \frac{1}{\frac{j\omega L_r}{1-\omega^2 C_r L_r}+\frac{j\omega L_e}{2}}$	$j\omega C_p + \frac{1}{\frac{1}{Y_{eq}}+\frac{j\omega L_e}{2}+\frac{j\omega L_{pe}}{2}}$
Y_{em}	$j\omega C_p + j2\omega C_g$	$j\omega C_p + j2\omega C_g + \frac{1}{\frac{1}{j\omega C_S}+\frac{j\omega L_r}{1-\omega^2 C_r L_r}+\frac{j\omega L_o}{2}+\frac{j\omega L_{po}}{2}}$
Y_{ee}	$j\omega C_p + j2\omega C_g + \frac{1}{\frac{j\omega L_r}{1-\omega^2 C_r L_r}+\frac{j\omega L_o}{2}}$	$j\omega C_p + j2\omega C_g + \frac{1}{\frac{1}{Y_{eq1}}+\frac{j\omega L_o}{2}+\frac{j\omega L_{po}}{2}}$
	$Yeq = \frac{2}{j\omega L_{pe}} + \frac{1}{\frac{j\omega L_r}{1-\omega^2 C_r L_r}+\frac{1}{j\omega C_z}}$	$Yeq1 = \frac{2}{j\omega L_{po}} + \frac{1}{\frac{j\omega L_r}{1-\omega^2 C_r L_r}+\frac{1}{j\omega C_s}}$

microstrip line has been etched with the loading structures. The extraction of the circuit parameters is done by matching the EM simulated S-parameters with the equivalent circuit simulated S-parameters. Comparisons of the S-parameters have been made with respect to magnitude and phase. These values have been used for loading the LC. The circuit simulation values for the LC are obtained from Eqs. 15, 16, 17, and 18.

$$\boldsymbol{C_p = \frac{\tan\left(\frac{\theta}{2}\right)}{Z_{0e}\omega}} \tag{15}$$

$$\boldsymbol{C_g = \frac{\tan\left(\frac{\theta}{2}\right)}{2\omega}\left(\frac{1}{Z_{0o}} - \frac{1}{Z_{0e}}\right)} \tag{16}$$

$$\boldsymbol{L = \frac{\sin(\theta)}{2\omega}(Z_{0o}) + Z_{0e}} \tag{17}$$

$$\boldsymbol{M = \frac{\sin(\theta)}{2\omega}(Z_{0e}) + Z_{0o}} \tag{18}$$

From the odd and even mode impedances, and center frequency the values of the equivalent circuit can be evaluated. Considering a 3 dB coupler with 50 Ohms terminating impedance at center frequency of 10 GHz the equivalent circuit values obtained from the Eqs. 15, 16, 17 and 18 are $C_p = 0.1326$ pF; $C_g = 0.33$ pF; $L = 1.114$ nH; $M = 0.7958$ nH and $K_L = \mathrm{M/L} = 0.714$. The equivalent circuit parameters for the dumbbell structure, SRR and DOS used in EM simulations of LC have been obtained from parameter extraction of EM simulated structures as explained earlier. The dumbbell parameters are $L_r = 0.76$ nH; $C_r = 0.025$ pF and for SRR the parameters are $L_p = 0.4$ nH; $C_s = 0.09$ pF; $L_r = 0.12$ nH; $C_r = 0.17$ pF and DOS parameters are $L_p = 0.505$ nH; $Cs = 0.075$ pF; $L_r = 0.13$ nH; $C_r = 0.045$ pF.

The circuit simulation results for the conventional LC, LCDB, LCSRR, and LCDOS are shown in Fig. 6. LC parameters for 4 dB coupling are $C_p = 0.15$ pF; $C_g = 0.26$ pF; $L = 1.03$ nH; $M = 0.65$ nH and $K_L = 0.63$. These parameters have been used for the LC when loading the structures. It is seen that, with the loading of the Lange coupler, there is a shift in the frequency towards the lower side. This is due to the effect of the loading resonant circuit parameters, and the greater the value of the loading capacitance and inductance, the higher the effect on the frequency shift.

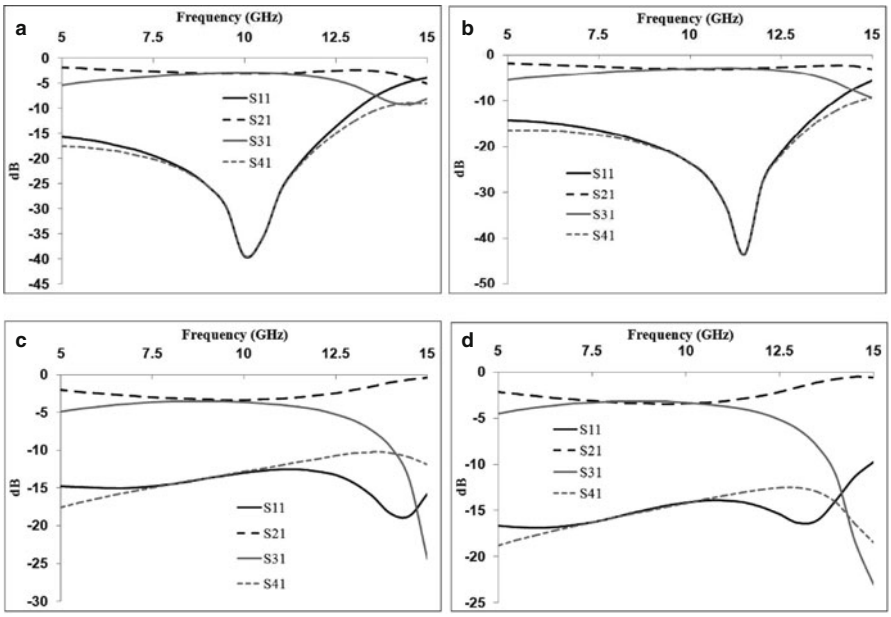


Fig. 6 Equivalent Circuit S-parameters of (**a**) Conventional Lange coupler. (**b**) LCDB (**c**) LCSRR (d) LCDOS

Conventional Lange coupler, LCDB, LCSRR, and LCDOS have been simulated and compared in a 3D EM simulator based on the finite element method. The substrate thickness is chosen to be 10mil, and the dielectric constant is 2.2. The dimensions of the coupler are selected based on the minimum dimensions that can be fabricated by the PCB vendor using the photolithography technique. ½ Oz copper thickness is used for simulations. The structures used for simulations are shown in Fig. 7, along with port numbers. The top copper layer is depicted in ash, and the bottom copper layer is in black. The white portions in the figure are etched-out portions in the ground plane. The dimensions of the LC are given in the caption of Fig. 7. The S-parameters of the EM simulated structures are shown in Fig. 8. From the extracted even and odd mode impedances, the coupling coefficient for the conventional LC is 4.08 dB, for LCDB is 2.8 dB, for LCSRR is 3 dB, and LCDOS is 2.8 dB.

Based on the equivalent circuit and EM simulated results, conventional LC, LCDB, LCSRR, and LCDOS have been fabricated on a ten mil RT Durlod substrate. The physical dimensions considered for fabrication are as mentioned above. The top part has been gold plated so that bonding between fingers could be performed. The fabricated structure is assembled in an aluminum box with grooves on the bottom side to accommodate DB, SRR, and DOS structures. The fingers of the Lange coupler are bonded using 0.7 mil gold wire. The assembled structure is shown in Fig. 9. The structures are tested using an E8363B vector network analyzer. The measured S-parameters for the structures are shown in Fig. 10. From Fig. 10c, d, loading of the LC provides better coupling than the conventional LC.

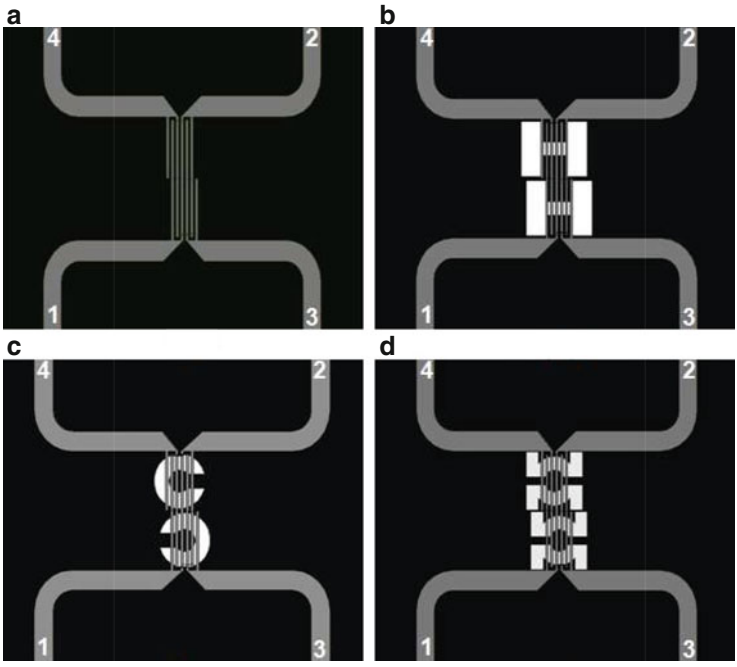


Fig. 7 EM simulated structure (**a**) Conventional Lange coupler. (**b**) LCDB (**c**) LCSRR (**d**) LCDOS (Length of the fingers = 4.2 mm, width of the fingers = 0.1 mm, gap between the finger = 0.1 mm and number of fingers = 6)

The return loss for all the structures is similar, and the isolation of the LCDOS is better than conventional and other loaded LC structures. The loss in the case of the loaded LC can be attributed to the radiation loss due to the loading structure. The coupling coefficient of the LCDOS is better than other structures, and it also provides better bandwidth with not much variation in the phase characteristics. However, there is a deviation in the measured results with respect to the EM simulated results due to the PCB fabrication tolerances. The fabricated width of the fingers is 80 microns, and the gap between the fingers is 120 microns. Due to the variation in the physical dimensions, the coupling coefficient of the Lange couplers is affected. This explains the sensitivity of the dimensions of the coupler for conventional Lange coupler and hence, the necessity of loading the structures with the dumbbell, SRR, and DOS. Considering that over-etching is uniform for all the finger widths and gaps, the three structures are simulated with a line width of 80 microns, a gap width of 120 microns, and a comparison of the measured results with EM simulated results in terms of amplitude imbalance is shown in Fig. 11. The increase in the insertion loss is due to the radiation loss occurring due to the patterned ground structures. The measured results show that LCDOS provides better performance than the other two loading structures in terms of amplitude and phase differences.

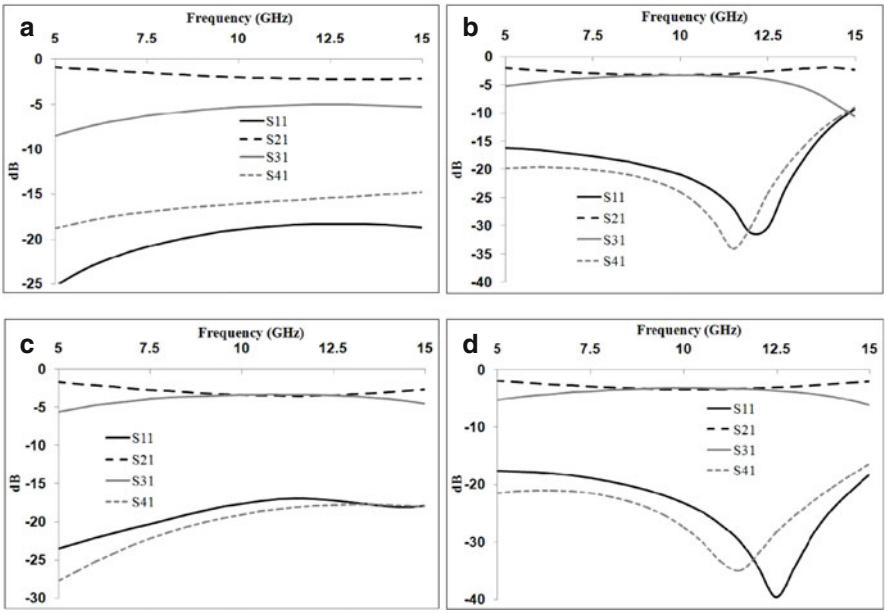


Fig. 8 EM simulated S-parameters of (**a**) Conventional Lange coupler. (**b**) LCDB (**c**) LCSRR (**d**) LCDOS

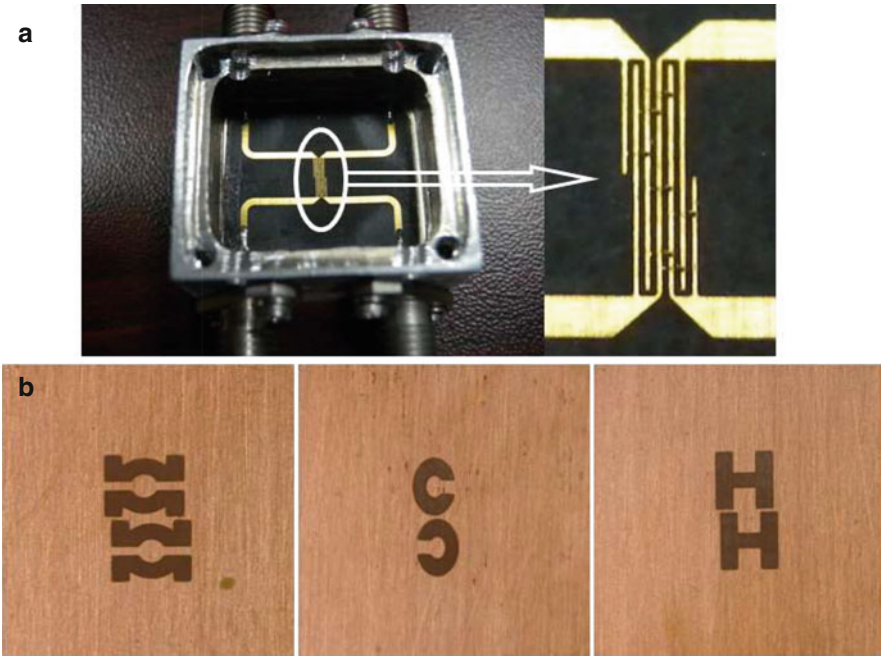


Fig. 9 (**a**) Fabricated Lange coupler in Aluminum box in inset the expanded view showing wire bonding between the fingers. (**b**) Bottom side of the Lange coupler (DOS, SRR, and DB)

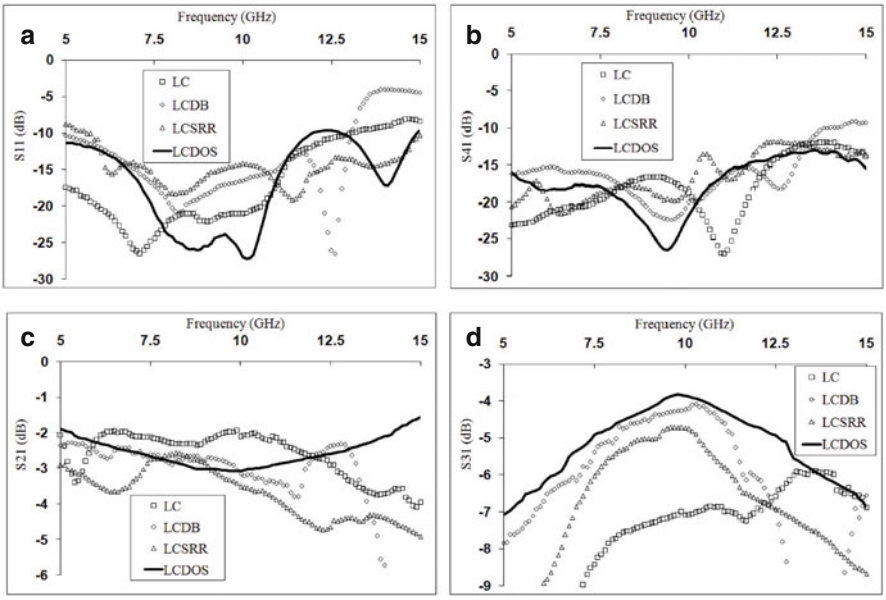


Fig. 10 Measured S-parameters (**a**) S11 curve (**b**) S41 curve (**c**) S21 curve (**d**) S31 curve

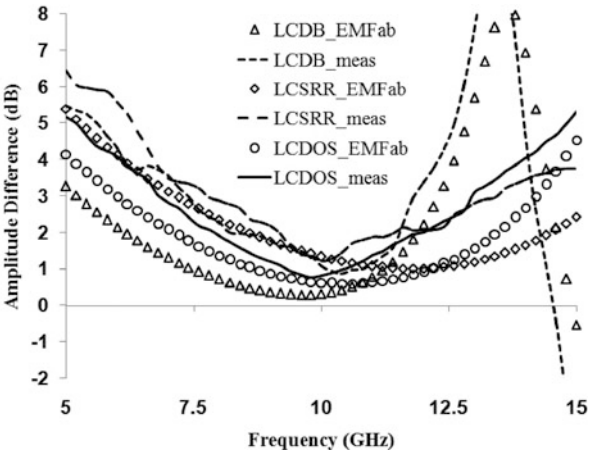


Fig. 11 Comparison of the amplitude difference of the measured and EM simulated results of the LCDB, LCSRR, and LCDOS with width 80 microns and gap between fingers as 120 microns

Conclusion

The basic design principles of couplers and equivalent circuit analysis of the Lange coupler with DB, SRR, and DOS loading have been presented. By using odd & even mode symmetry in conjunction with the superposition principle, four conditions have been derived for the loaded c. The proposed equivalent circuit analysis has been

verified using circuit analysis by substituting the loading structures' values obtained from DB, SRR, and DOS parameter extraction. The effectiveness of the proposed method has been demonstrated through the practical design of the loaded Lange coupler.

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Metamaterial Lens for Nondestructive Evaluation of Composites

10

Saptarshi Mukherjee, Srijan Datta, Xiaodong Shi, and Mahmoodul Haq

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Abstract

Structural fiber-reinforced polymer composites have been increasingly used to replace metallic parts in various industrial applications, due to their lightweight and superior mechanical properties. Highly tailorable characteristics of composites have made them an attractive solution in aerospace, automotive, civil, and other sectors alike. However, composite material structures are complex and vulnerable to various defects and damages during manufacturing or usage. Hence, nondestructive evaluation (NDE) of composites to detect such defective

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structures is crucial in avoiding structural failures and significant losses. Microwave NDE is the method of using microwaves to detect embedded defects in low-loss dielectric structures. Microwaves can travel large distances in free space and penetrate deep into nonconducting materials. This makes them suitable for rapid inspection of composites as the probe standoff distances can be large in far field mode. Conventional microwave-imaging systems of different types of defects in composites using microwaves from far field distances are diffraction limited, which means that defects or anomalies that are smaller than the operating wavelength are undetected. Metamaterials are specifically designed periodic engineered materials to offer properties that can overcome and enhance diffraction-limited imaging resolution. This chapter focuses on the recent developments of theoretical, numerical, and experimental aspects of metamaterial lens structures, for enhancing imaging resolution of far field microwave sensors, and thereby offer subwavelength detection capabilities from far field.

Keywords

Metamaterials · Lenses · Microwave imaging · Microwave theory and techniques · Image resolution · Numerical models · Nondestructive evaluation

Introduction

Over the past few decades, composite materials have become a significant area of research and development and are being increasingly used due to their application in aerospace and automotive industries, structural applications such as rehabilitation of buildings, bridges, vehicles, wind turbines, etc. Research work on using FRP Composites to strengthen structures was initiated in the late 1980s by Ku et al. (2011). Their lightweight and high mechanical strength can significantly increase their performance. Combined with the strategic tuning of mechanical properties and design flexibility, these key advantages contributed to their wide acceptance in aviation, space, marine, automotive, sporting, and construction industries. For example, nearly half of the airframe of the Boeing-787 Dreamliner is now created using carbon and glass FRP composites (Fig. 1). Metals are being replaced by composites, either partially or fully using adhesively bonded joints, thus providing lightweight and mechanically stronger designs (Fig. 2). For a good insight into the usage and functioning of these materials, significant research results on their quality and long-term durability are reported in Kline (2017).

However, there are several major challenges of using such advanced materials. Composites are prone to several types of flaws both during manufacture and in service use. These include delamination, disbonds, voids, cracks and fiber breakage, and foreign objects. An example is the American Airlines 587 disaster in 2001, which was caused by a failure in the carbon fiber-reinforced polymer (CFRP) vertical stabilizer (Weiss et al. 1981). Several factors such as material contrasts in interface properties and anisotropy are a prime reason for many flaws Lorraine and Kline (2006). Further, FRP composite laminates may be disbonded from the

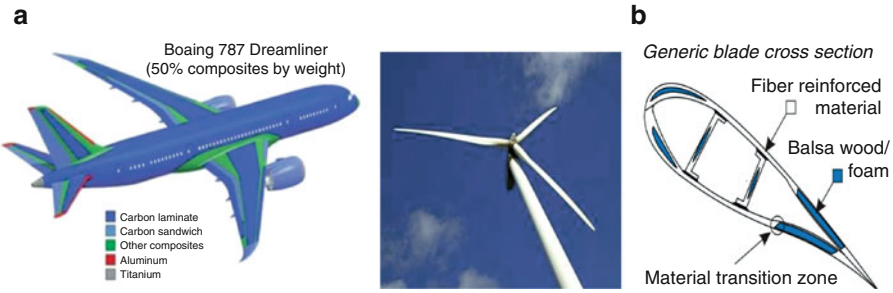

Fig. 1 Applications of composite materials: (**a**) assembly of Boeing-787 Dreamliner; (**b**) wind turbine blade

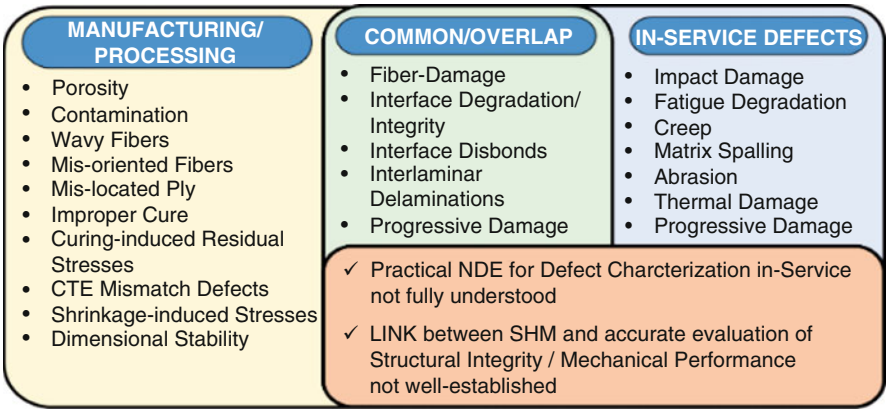

Fig. 2 Summary of common defects in composites. (Adapted from [4])

substrate metals due to various issues such as temperature gradients, high stress, and use of improper adhesives. Although the applications of these materials have received significant attention, methods to monitor their structural health are still a focus of considerable interest. Hence, accurate NDE techniques are essential for such critical components to maintain the overall structural integrity of the system (Banerjee et al. 2019); Bray and Stanley 1996).

Some of the widely used inspection techniques are based on visual testing, which is ineffective in case of buried anomalies that are not visible. Industrial practices employ a range of well-established NDE techniques such as ultrasonic inspection with water-coupled and air-coupled transducers, ultrasonic testing (UT)-phased arrays, pulsed-flash thermography, X-ray, and optical inspection (Karpenko et al. 2016; Martz et al. 2016). Electromagnetic NDE sensors are of particular interest because of their inherent advantages such as low cost, noncontact nature that allows high scan speed, single side access, and inspection at large liftoff distances (far field measurements) (Kharkovsky and Zoughi 2007). A range of electromagnetic NDE techniques spanning the electromagnetic spectrum is growing rapidly, including

eddy current testing, capacitive imaging, and microwave NDE systems (Yang et al. 2009); Yin et al. 2010); Mukherjee et al. 2018) due to recent advances in wireless electronics, manufacturing technologies, and computational capabilities. Microwave NDE is particularly tailored toward lossless and low-loss dielectric materials because of the capability of microwaves to penetrate through the material thickness, without suffering much attenuation (Arunachalam et al. 2006). Contrasts in dielectric scattering of microwaves by the materials can be utilized by tomographic algorithms to reconstruct the material's subsurface properties and hence provide information about its structural integrity.

Microwaves fundamentally work in two modes, namely near field (distances typically less than a wavelength from the source) and far field (distances generally more than two wavelengths). The advantages of noncontact measurement provided by Microwave NDE can be fully realized in the far field mode. However, the spatial resolution of far field microwave imaging is limited due to the imposition of diffraction limits. Defects or anomalies much smaller in size than the operating wavelength are generally not accurately imaged (Mukherjee et al. 2020). Near field microwave sensor systems can provide spatial resolution beyond the diffraction limit, but the probe standoff distance is very small leading to large scanning time, and hence are not suitable for building noncontact rapid scan systems. The advantages of both modes can be exploited by the use of metamaterial devices.

Metamaterials have been increasingly used in the past few decades in the design of novel microwave circuits and sensor systems (Veselago 1968). Metamaterials are artificially engineered periodic structures with properties not found in natural materials. The unique properties of metamaterials offer several benefits such as super-resolution and subwavelength feature size that is not found in conventional materials (Smith et al. 2004). Prior experiments have demonstrated the capability to design a metamaterial lens with a negative relative refractive index, to form images that overcome diffraction limits (Pendry 2000). A lens constituting of metamaterial, by virtue of its unique properties, can focus EM waves beyond the diffraction limit (Ozbay et al. (2008). Therefore, an object/defect, smaller in size than the operating wavelength, once placed at the focal plane of the MM lens, will interact with microwaves at that frequency. Subwavelength defects and features, which are undetectable in far field, can be imaged using the lens. This phenomenon can be exploited to build a microwave NDE system integrated with a metamaterial lens, and thereby provide rapid scanning at large standoff distances as well as enhanced spatial resolution.

Electromagnetics of Metamaterials

Maxwell's Equations

Metamaterials are designed to have both negative permittivity and permeability that leads to its negative index of refraction, given by (Veselago 1968)

$$n = -\sqrt{\mu\varepsilon}. \tag{1}$$

The most general dispersion equation connecting frequency ω and wave vector of a monochromatic wave k is given by

$$\left|\frac{\omega^2}{c^2}\varepsilon_{ij}\mu_{ij} - k^2\delta_{ij} + k_i k_j\right| = 0, \tag{2}$$

where ε_{ij}, μ_{ij}, and δ_{ij} are the permittivity, permeability tensors, and the Kronecker delta function. It might be seen that a simultaneous change of signs of e, g has no changes to the above relations. Hence substances with $\varepsilon < 0$ and $\mu < 0$ have properties different from substances having positive ε, μ. Looking at Maxwell's equations for time-harmonic waves in k-space,

$$k \times \mathbf{H} = -\frac{\omega}{c}\varepsilon\mathbf{E}; k \times \mathbf{E} = -\frac{\omega}{c}\mu\mathbf{H}. \tag{3}$$

From the above sets of equations, it is clearly seen that E, H, and k form a left-handed triplet of vectors for metamaterials, while they form a right-handed triplet for a normal material. This results in unique wave properties such as phase velocity being antiparallel to group velocity and reversed Doppler effects (Lee et al. 2010); Belonogaya et al. 2011).

High Resolution of a Metamaterial Lens

Due to the negative refractive index of metamaterials, electromagnetic waves radiating from a point source can be manipulated through a planar metamaterial slab/lens in order to focus at some distance away from the slab Pendry (Pendry 2000). Additionally, evanescent waves that are typically not transmitted in conventional materials are amplified in metamaterials due to their negative refractive index and thus can be recovered at distances away from the lens (Smith et al. 2004). Thus, while the imaging resolution in normal materials is diffraction limited with the maximum resolvable wave number of $k_{\max} = 2\pi/\lambda$, the recovery of evanescent waves with metamaterials leads to an enhanced resolution ($k_{\max} << 2\pi/\lambda$). The perfect focusing nature and recovery of evanescent fields are primarily responsible for providing a resolution that breaks diffraction limits, capable of resolving sub-wavelength defects for NDE.

It is also important to mention that while Veselago formulated the negative refraction capability of metamaterials mathematically by matching boundary conditions for E and H fields (Veselago 1968), the solution was purely mathematical, and all such solutions always need to be checked for physical validity and realization using numerical simulations involving periodic structures. Munk in his book "Metamaterials Critique and Alternatives" (Munk 2009) has discussed the difficulties and misconceptions associated with the synthesis of negative index of refraction materials using periodic structures to avoid misinterpretation among the scientific community.

Numerical Studies

Numerical methods for computing electromagnetic fields and their interaction with a metamaterial lens are critical due to several factors as follows:

- Analytical computations are nontrivial for such complex media.
- Number of prototypes that need to be fabricated is reduced.
- Numerical computations can be used as a tool for inverse design.

Homogenization of Metamaterials

Numerical simulations of a metamaterial lens consisting of periodic structures may be computationally expensive. In an attempt to reduce the computation time, the electromagnetic properties of an ideal metamaterial slab can be homogenized. It has been studied that the electromagnetic properties of a metamaterial can be described by plasma wave eqs. A physical design of metamaterials typically consists of thin wire structures that mimic an electric plasma, and conducting wire loops that behave like a magnetic plasma, with resonances in the microwave spectrum (Pendry et al. (1999, 1998), ω_{pe}. Electric and magnetic plasma equations are given by

$$\varepsilon = \varepsilon_0\left(1 - \frac{\omega_{pe}^2}{\omega^2}\right), \quad \mu = \mu_0\left(1 - \frac{\omega_{pm}^2}{\omega^2}\right), \tag{4}$$

where ω_{pe} and ω_{pm} are the peak electric and magnetic plasma frequency. ε is negative for $\omega < \omega_{pe}$, while μ is negative $\omega < \omega_{pm}$. The frequency-dispersive electric and magnetic properties of a lossy Drude model characterizing the DNG layer can be further expressed as (Goswami et al. 2014):

$$\varepsilon(\omega) = \varepsilon_0\left(1 + \frac{\omega_{pe}^2}{\omega_{oe}^2 - \omega^2 + j\tau_e\omega}\right), \ \mu(\omega) = \mu_0\left(1 + \frac{\omega_{pm}^2}{\omega_{om}^2 - \omega^2 + j\tau_m\omega}\right), \tag{5}$$

where e_0 and g_0 are the free space permittivity and permeability, m_{oe} and m_{om} are the electric and magnetic frequency cutoffs, and τ_e, τ_m are associated with electric and magnetic losses. It can be seen from (5) that $e < 0$ when $\omega_{oe} < \omega_{pe}$ and $\mu < 0$ when $\omega_{om} < \omega < \omega_{pm}$. Tuning these parameters, it is possible to attain a value of $\varepsilon = -1$ and $\mu = -1$ at a single frequency. As an example, the refractive index associated with specific metamaterial parameters is shown in Fig. 3.

Using the above frequency-dispersive relationship shown in (5), the displacement current term can be modified as follows:

$$D(\omega) = \varepsilon_0 E(\omega) + P(\omega), \ P(\omega) = \frac{\varepsilon_0\omega_{pe}^2}{\omega_{oe}^2 - \omega^2 + j\tau_e\omega}E(\omega) \tag{6}$$

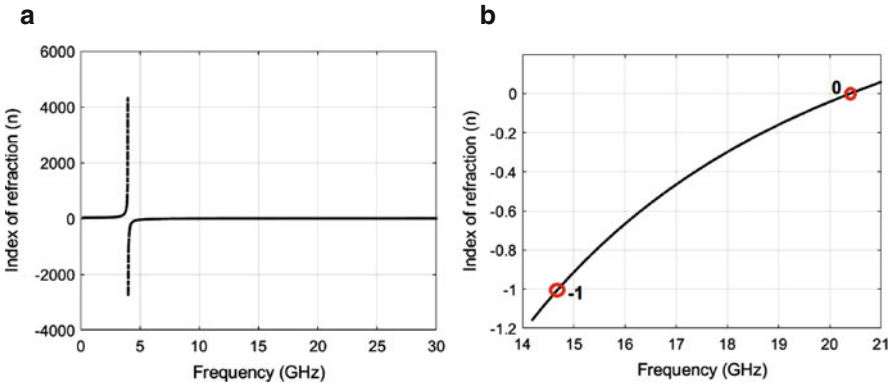

Fig. 3 Metamaterial parameters: $\omega_{pm} = \omega_{pe} = 20$ GHz, $\omega_{om} = \omega_{oe} = 4$ GHz, and $\tau_e = \tau_m = 0$ (**a**) Refractive index of a DNG medium, (**b**) zoomed in plot shows $n = 1$ and zero crossover point

where $E(\omega)$, $D(\omega)$, and $P(\omega)$ are the electric field, displacement current, and polarization density of the medium, respectively. In order to model the above equation with time domain simulations, the frequency dispersiveness term can be converted from frequency to time domain using different techniques such as auxiliary differential equation (ADE) method that converts frequency domain into time domain difference terms. For example, using Fourier transform, $j\omega P(\omega)$ can be expressed as $\frac{\left(P^n - P^{n-2}\right)}{2\Delta t}$, while $(j\omega)^2 P(\omega)$ as $\frac{\left(P^n - 2P^{n-1} + P^{n-2}\right)}{\Delta t^2}$, where P^n, P^{n-1}, and P^{n-2} are polarization fields at time n, $n-1$, and $n-2$, respectively. Using the above relations, the fields in 6 can be updated by time-marching algorithm as follows:

$$P^n = \left(\frac{\frac{2}{\Delta t^2} - \omega_{oe}^2}{\frac{\tau_e}{2\Delta t} + \frac{1}{\Delta t^2}}\right) P^{n-1} + \left(\frac{\frac{\tau_e}{2\Delta t} - \frac{1}{\Delta t^2}}{\frac{\tau_e}{2\Delta t} + \frac{1}{\Delta t^2}}\right) P^{n-2} + \left(\frac{\omega_{pe}^2 \varepsilon_0}{\frac{\tau_e}{2\Delta t} + \frac{1}{\Delta t^2}}\right) E^{n-1} \tag{7}$$

A similar approach can be used to model magnetic fields by updating the magnetic polarization density $M(\omega)$ inside the metamaterial layer.

Utilizing the above relationships in a numerical model, the wave propagation in a metamaterial slab can be visualized. As an example, a simulation involving wave propagation simulations in a DPS-DNG-DPS medium with the schematic and spatial dimensions shown in Fig. 4(a) using a finite difference time domain (FDTD) approach is shown in Fig. 4(b). Electromagnetic waves undergo a phase change of 180° at the first metamaterial interface, followed by a similar phase change of 180° at the second interface, due to a backward propagation in the DNG region and forward propagation in the DPS region (Mukherjee et al. 2019) (Fig. 5).

Simulations of a Metamaterial Lens

It can be shown in Fig. 7(a) that fields radiating out from an excitation source located at a distance (d_1) from the lens undergo two focusing phenomena, once inside the

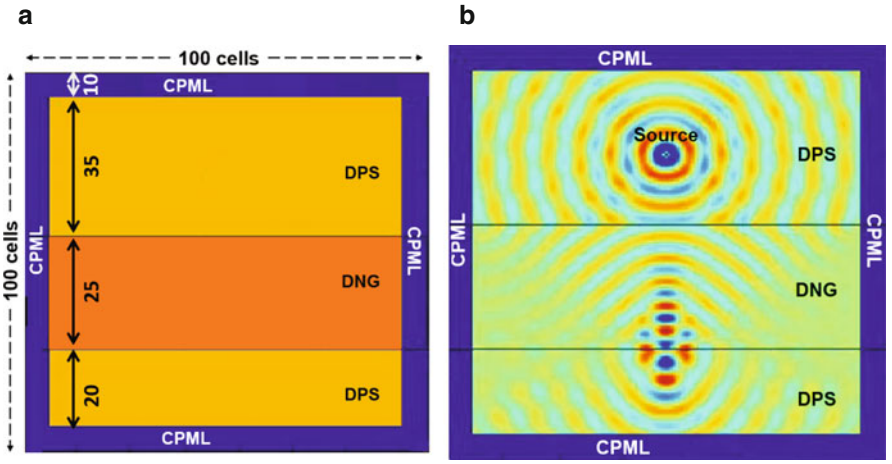

Fig. 4 DPS-DNG-DPS interface: (**a**) schematic for FDTD simulation model, (**b**) wave-propagation shows a phase reversal of 180° at the first DPS-DNG interface and another phase reversal of 180° at the second DPS-DNG interface

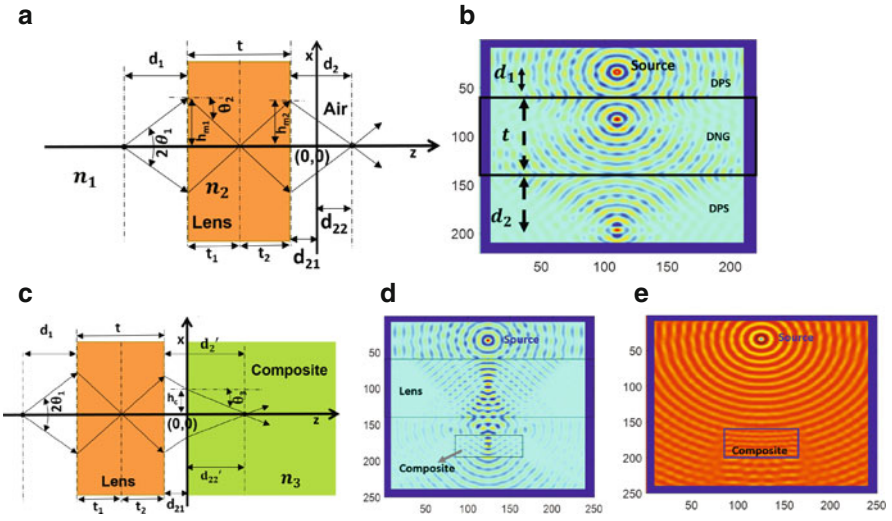

Fig. 5 Properties of a metamaterial lens: (**a**) Ray diagram of metamaterial lens in the presence of air; (**b**) focusing nature of the metamaterial lens in free space; (**c**) ray diagram of metamaterial lens in the presence of composite half space; (**d**) focusing nature of the metamaterial lens in free space; (**e**) near plane wave illumination of composite without the lens. Simulation parameters: d1 = 3.5 λ_0, t = 8 λ_0, and the thickness of the composite is $4\lambda_0 \times 8\lambda_0$, d_2^1 is obtained as 3.5 λ_0 (Mukherjee et al. 2019).

lens followed by outside the lens in free space at a distance (d_2) from the lens interface. The relationship between the parameters d_1, d_2, t, and incident angle (0_i) can be derived using Snell's principle of refraction and geometrical optics and is expressed as (Mukherjee et al. 2019),

$$d_1 + d_2 = t\frac{\cos\theta_1}{\sqrt{n^2 = \sin^2\theta_1}}. \tag{8}$$

For a perfect metamaterial lens (where $n = -1$), (8) is reduced to $d_1 + d_2 = t$. A numerical example was constructed with FDTD in order to demonstrate the focusing property of a perfect metamaterial lens (n = −1) with the chosen parameters $d_1 = 3.5$ λ_0 and t = 8 λ_0 (Mukherjee et al. 2019). The electric field snapshot shown in Fig. 7 (b) shows the two focal spots inside and outside the lens, with $d_2 = 4.4$ λ_0, which satisfies the perfect focusing condition. Similar to free space, the lens can also be used to focus electromagnetic waves inside a composite material, as seen in the corresponding ray diagram in Fig. 7 (c). As observed in the ray diagram, an additional refraction at the composite boundary, diverges the waves to focus at a different location inside the composite. Figures 7(d, e) show that while plane waves are incident on the composite in the absence of the lens, well-focused fields are incident on the composite in the presence of the lens.

Focusing Performance in Composite

The focusing performance in the composite is directly related to the imaging resolution of the lens and thus its defect detection capability. As seen in the ray diagram in Fig. 7(d), in accordance with Snell's refraction theory, the wave is focused at a larger distance $d\backslash$ from the lens instead of d_2 $\left(d_2^1 > d_2\right)$. The focusing point inside the composite sample can be determined from the ray diagram. Utilizing Snell's theory of refraction and geometrical optics, we can obtain an analytical expression of the focal spot $\left(d_{22}^1\right)$ in terms of d_{22}, £$_{r,}$ and θ_1, as follows:

$$d_{22}^1 = d_{22}\left(\frac{\sqrt{\varepsilon_r - \sin^2\theta_1}}{\cos\theta_1}\right). \tag{9}$$

It is observed from (9) that $d_{22}^1 = d_{22}$, when $\varepsilon_r = 1$, leading to a focal point at d_2 in free space. A comparison between the above expression and wave-based simulations for a dielectric sample showed good correlation and provided a simple way to calculate the focusing region in the composite material, with the knowledge of its dielectric properties (Fig. 6) (Mukherjee et al. 2019).

The degree of focusing by a metamaterial lens is key toward resolving finer defects in materials. Simulations conducted in Mukherjee et al. (2019) show that while an ideal focal spot with symmetric cross range and downrange resolution is obtained in free space, in the absence of the composite sample, a nonideal focal spot with poor localization along the down range is obtained in the presence of the composite due to wave scattering at the composite layer, as shown in Fig. 7. The relatively poor downrange resolution would make it harder to resolve defects in the downrange than the cross range directions. Inverse computational techniques such as a deep learning technique for a topologically modified metamaterial lens are currently an active area of research in order to improve the resolution (Yao et al. 2019).

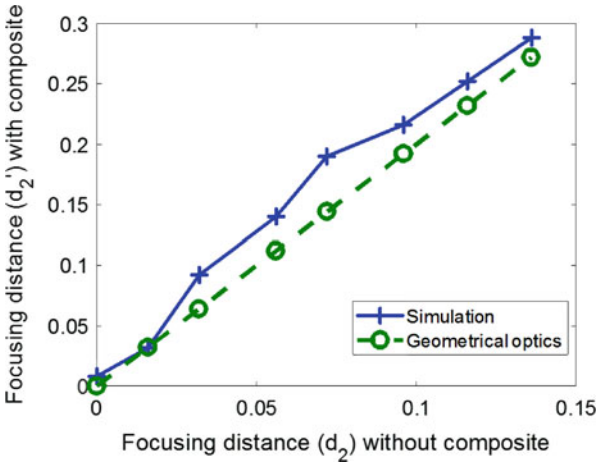

Fig. 6 Comparison between geometrical optics theory and simulation for determining focusing distance: Composite $\varepsilon_r = 4\left(d_{22}^1 \tilde{2} d_{22}\right)$. (Mukherjee et al. (2019))

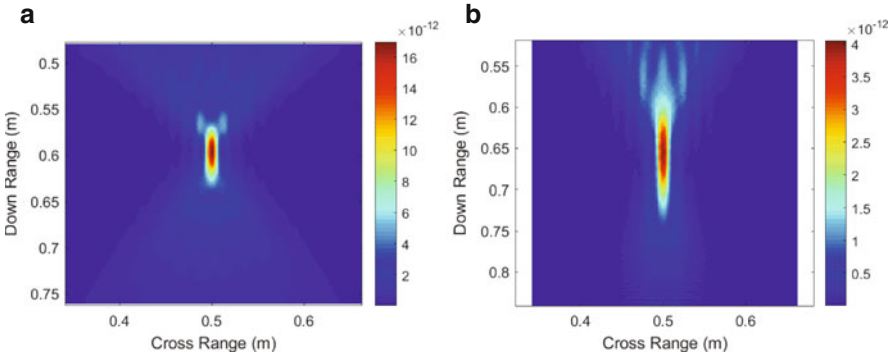

Fig. 7 Focusing resolution of the lens in (**a**) free space, (**b**) composite. Focal spot is calculated numerically using the time-integrated energy of the electromagnetic wave. (Mukherjee et al. (2019))

Nondestructive Imaging with the Lens

Numerical simulations were performed to study the capability of a metamaterial lens for nondestructive imaging applications (Mukherjee et al. 2019). A far field time-reversal-imaging method was developed to numerically compare imaging performance of a pulsed time-domain microwave NDE system with and without the presence of a lens. The method was utilized to understand the benefits and limits of the metamaterial lens for resolving closely spaced subwavelength defects in the subsurface of a composite slab. Details on the time reversal-based imaging technique can be found in Mukherjee et al. (2018). Transmission mode measurements with a metamaterial slab placed between the receiving antenna array and the composite sample were performed as shown in Fig. 8(a). Preliminary numerical results demonstrate the superiority in imaging the excitation source in both cross and downrange

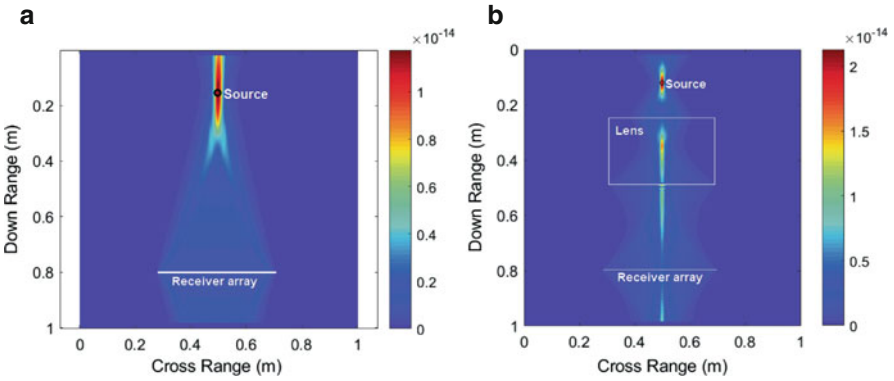

Fig. 8 TR with metamaterial lens shows enhancement of microwave imaging for source imaging, (**a**) TR energy without metamaterial lens, (**b**) TR energy with lens. A modulated gaussian pulse of pulse width 73 ps corresponding to a 12 GHz bandwidth and center frequency 14.7 GHz (frequency at which metamaterial lens has n = − 1) is the excitation source. The metamaterial lens is 40 cm × 40 cm in dimensions. (Mukherjee (2018))

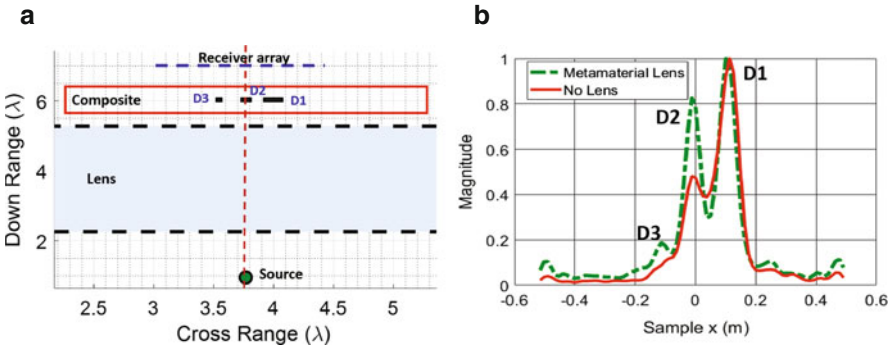

Fig. 9 TR with metamaterial lens shows enhancement of microwave imaging for a subwavelength defect in a composite slab with three defects (D1, D2, and D3) of dimensions 1 cm × 4 mm (λ/2 × λ/5), 4 mm × 4 mm (λ/5 × λ/5), and 2 mm × 4 mm (λ/10 × λ/5), separated in cross-range by a distance of 4 mm (λ/5) where X is the wavelength corresponding to the pulse center frequency, (**a**) schematic of test geometry, (**b**) TR energy comparison

directions in the presence of the lens. Although the authors pointed out other artifacts in the time-reversed images due to wave scattering at different material interfaces, the artifacts did not degrade the quality of source reconstruction and were not eliminated as part of the research. Additional simulations were also conducted to test subwavelength defect detection sensitivity of the lens. Studies involved different configurations of closely spaced subwavelength defects in the cross range and downrange directions. The excitation source location and thickness of the metamaterial lens were chosen such that it focused fields inside the composite sample around the defect region. The signal line-outs were derived from the microwave-imaging results to compare the superiority of the lens, as shown in Fig. 9(b). Results showed higher

sensitivity and imaging efficiency toward closely spaced defects with the presence of the metamaterial lens than free space and strongly indicated that a physically designed metamaterial lens can significantly enhance the resolution of far field microwave imaging. The lens can also be used to vary the focusing spot inside the composite sample by translating the source location in order to tune the sensitivity toward different sample locations. Numerical results showed best detection when the source is horizontally aligned to the defect. The implication of this result is that one can focus the energy to any desired location in the composite to obtain a high resolution image of the composite sample. Further simulations showed that the overall microwave-imaging system has a weaker sensitivity and focusing along the downrange direction ($\wedge X_0/3$), but to a lesser extent than the cross range direction ($\wedge X_0/10$) due to the finite thickness of the lens and limited antenna array aperture (Mukherjee et al. 2019).

Modeling of the Physical Structure

Theory

Although homogenized model of a metamaterial lens permits rigorous numerical study on the focusing behavior of a bulk LHM material, negative refractive index of an actual physical metamaterial occurs due to the electromagnetic (EM) response of its periodic building blocks, known as unit cells. Therefore, engineering a metamaterial lens requires one to start from the design of its unit cells. To that end, the theory governing the working principles of metamaterial unit cells and how they effectively act as a bulk metamaterial medium is briefly discussed. Following that, a designing guide to the first and most common left-handed metamaterial unit cell design is presented using a commercial FEM full wave simulator.

A periodic array of conducting elements or unit cells can behave as an effective scattering medium in the presence of incident EM waves when the cell dimension and lattice spacing are much shorter than the operating wavelength. Pendry et *al.* in their seminal papers introduced periodic structures, which behaved as effective media having negative permeability and negative permittivity in the microwave regime (Pendry et al. 1999); Pendry et al. 1998). This led to the pioneering paper by Smith et al. in 2000, where a negative refractive index material was realized for the first time, by combining both the negative permittivity and permeability structures, which we will be discussing below (Smith et al. 2000).

It was shown in Pendry et al. (1999) that an array of resonating rings can exhibit an effective permeability behavior, similar to that of magnetic plasmas, in response to microwave radiations. Splits and gaps were introduced in the rings to act as capacitive elements to considerably increase the magnetic effects of the array. The splits also decreased the resonating frequency, consequently driving the plasma behavior toward microwave regime. Figure 10 shows the design schematic of split ring resonators (SRR). Under the excitation by an external time-varying magnetic field parallel to the axis of the rings, the array behaves as a bulk medium having an effective permeability given by,

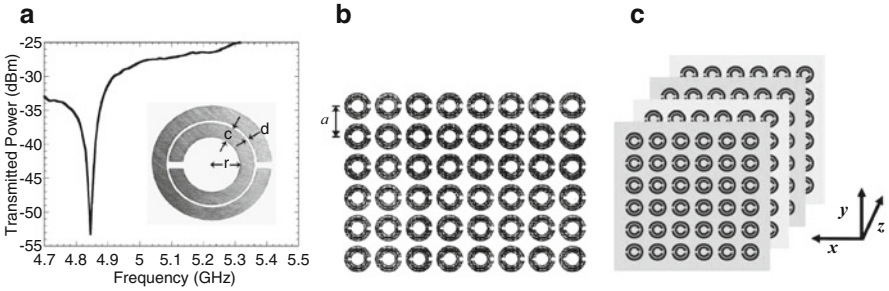

Fig. 10 (**a**) Resonance curve of an actual copper split ring resonator (SRR). c = 0.8 mm, d = 0.2 mm, and r = 1.5 mm. The SRR has its resonance at about 4.845 GHz0 (Smith et al. (2000)); (**b**) plain view of a split ring structure in a square array (lattice spacing a) (Pendry et al. (1999)); and (**c**) a periodic array of SRRs on a dielectric board [8]

$$\mu_{eff}(\omega) = 1 - \frac{\frac{\pi r^2}{a^2}}{1 - \frac{3lc_0^2}{\pi\omega^2 \ln\frac{2c}{d} r^3 + i\frac{2l\rho}{\omega r \mu_0}}}, \tag{10}$$

where a is the lattice constant, l is the distance between layers, r, c, and d are defined in Fig. 10(a), p is the resistance/unit length of the rings measured around the circumference, c_0 is the speed of light in vacuum, μ_0 is the magnetic permeability of vacuum, and ω is the angular frequency of incident radiation. A more generic form of the effective permeability of an SRR array is given by,

$$\mu_{eff}(\omega) = 1 - \frac{F\omega^2}{\omega^2 - \omega_0^2 + i\omega\Gamma}, \tag{11}$$

where F is the fractional volume of the unit cell occupied by the rings and r is the dissipation factor. Expressions for F and r can be found out by comparing terms in Eqs. 10 and 11. ffl$_0$ is the resonant frequency given by,

$$\omega_0 = \sqrt{\frac{3lc_0^2}{\pi \ln\frac{2c}{d} r^3}} \tag{12}$$

Equation 11 shows that the real part of μ_{eff} is negative in between the resonant frequency ω_0 and magnetic plasma frequency ω_{mp} given by $\omega_{mp} = \frac{\omega_0}{\sqrt{1-F}}$. Propagating wave modes are inhibited in this frequency band due to of the effective negative permeability of the SRR medium. The upper solid curve in Fig. 12(c) shows the transmission characteristics of a periodic array of SRR. We observe a dip in the transmission power near the resonant frequency of the single SRR element, which was studied numerically and attributed to be a consequence of negative permeability of the material (Smith et al. 2000).

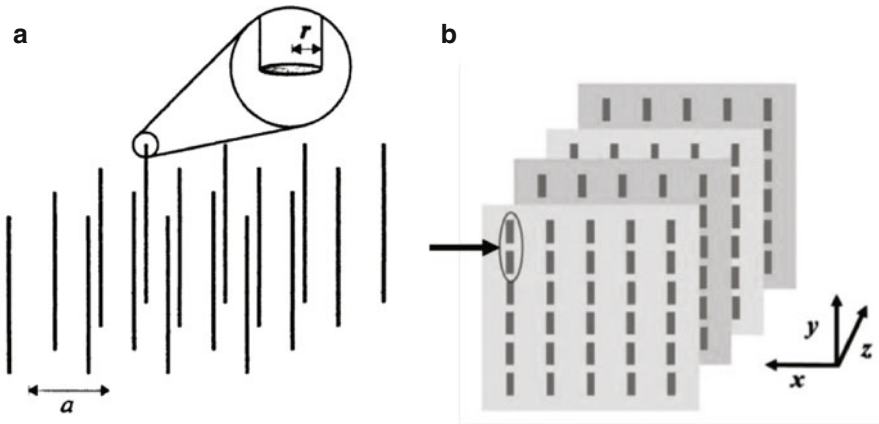

Fig. 11 (**a**) An array of wires with radius r and lattice spacing *a* (Pendry et al. 1998), (**b**) A periodic array of thin wires on a dielectric board [8]

A periodic array of thin metallic wire structures, under the influence of a time-varying electric field, can mimic an electric plasma at microwave frequencies (Pendry et al. 1998). A periodic arrangement of such wires is shown in Fig. 11.

Effective permittivity of such an array of thin metallic wires, in presence of an external electric field parallel to the wires, is given by

$$\varepsilon_{eff} = 1 - \frac{\omega_{ep}^2}{\omega^2}, \tag{13}$$

where ω_{ep} is the electric plasma frequency given by

$$\omega_{ep} = \frac{2\pi c_0^2}{a^2 \ln\left(\frac{a}{r}\right)} \tag{14}$$

where c_0 is the speed of light in free space, a is the lattice parameter, and r is the radius of the wires. (13) shows that the real part of effective permittivity has negative values for frequencies $\omega < \omega_{ep}$. This causes the wire medium to impede propagating modes in that frequency regime.

Combining both the SRR and wire arrays in a periodic manner gives rise to a metamaterial medium having simultaneous negative permeability and permittivity. A typical unit cell model of a metamaterial consisting of a single SRR and wire is shown Fig. 12(a). Due to the simultaneous negative properties of the combined structure, a passband with negative phase velocity occurs within the previously forbidden band of split rings (Fig. 12c). Left-handed behavior of the metamaterial is observed in this transmission band, and the transmission peak is known as a true left-handed peak.

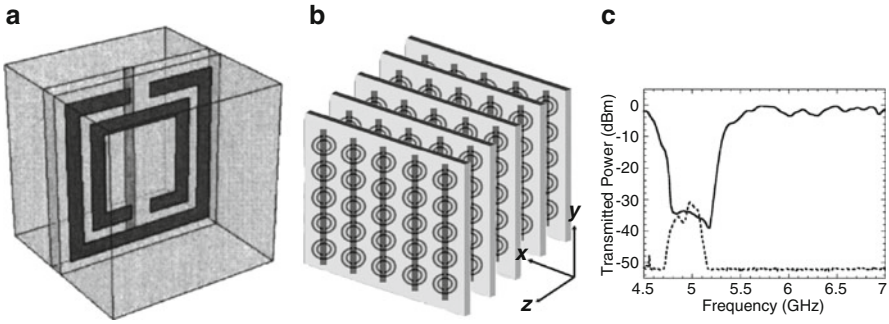

Fig. 12 (**a**) A single unit cell of a negative index metamaterial. The rings can have circular or square structure as shown. The split ring resonator responds magnetically, and the wire responds electrically to electromagnetic fields [6]; (**b**) PCB implementation of a periodic array of 5 × 5 × 5 unit cells [8]; (c) S21 comparison for SRR only and combined medium. The upper curve (solid line) is that of the SRR array with lattice parameter $a = 8.0$ mm. By adding wires uniformly between split rings, a passband occurs where μ and ε are both negative (dashed curve). (Smith et al. (2000))

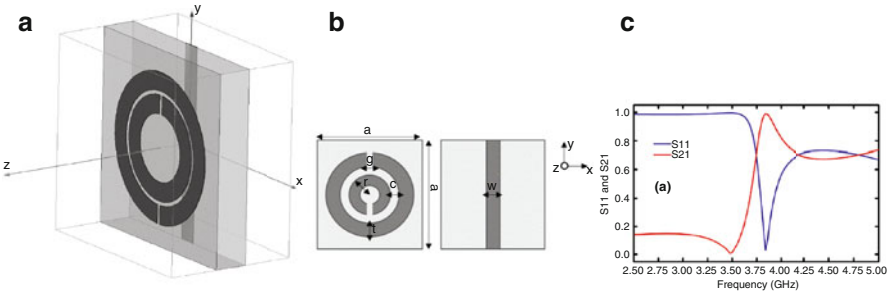

Fig. 13 (**a**) HFSS unit cell model. Perfect H and perfect E boundary conditions are applied to the x-y and z-x boundaries to simulate an infinite array of the unit cells and ensure necessary wave polarization. Wave ports are assigned to the y-z boundaries. The metal rings and wires can be modeled as PEC's to allow faster implementation; (**b**) schematics for unit cell model showing front and back of PCB; (**c**) S parameters results obtained from unit cell simulation. Left-handed transmission peak occurs at 3.75 GHz. The dimensional parameters used for modeling: d = g = 0.2 mm, w = 0.9 m, t = 0.9 mm, r = 3.6 mm, and a = 8.8 mm. A lossy FR4 PCB of thickness 1.6 mm is used as the dielectric substrate. The distance between consecutive PCB layers is 6.5 mm. (Ozbay et al. (2008))

Unit Cell Model

A finite metamaterial array can be numerically modeled in a Finite Element Method EM simulator by implementing an infinite array of the unit cells in the direction of the lattice vectors. This numerical approximation is achieved by imposing periodic boundary conditions on the unit cell. A typical unit cell model using commercial EM simulator HFSS is shown in Fig. 13a. The Scattering (S) parameters obtained from a

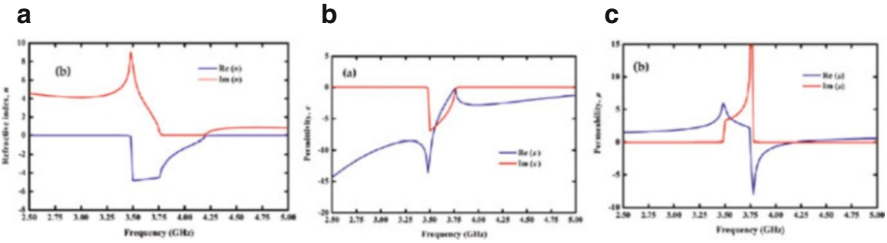

Fig. 14 Extracted real and imaginary part of: (**a**) refractive index; (**b**) permittivity; and (**c**) permeability from simulation model. (Ozbay et al. (2008))

full wave analysis of the model are shown in Fig. 13c. As expected, a left-handed pass band around the resonating frequency of the SRRs is observed.

Extraction of the constituent properties from S parameter data of a metamaterial unit cell will further verify the left-handed nature of the medium and facilitate its design and optimization process. Details of the material extraction are presented in Appendix. The extracted refractive index, permittivity, and permeability are shown in Fig. 14. It can be seen that the real part of the properties is negative around the left-handed peak, a prerequisite for metamaterial behavior.

Experimental Work

After the first physical LHM metamaterial design was realized by Smith et al., extensive research on experimental verification of left-handed behavior followed. In this section, key research works for characterization and demonstration of metamaterial behavior are briefly discussed. Some of the important experimental works are summarized in a chronological order in Table 1 for quick reference.

The transmission and reflection measurements of metamaterials reported by Smith et al. were performed in a waveguide chamber (Smith et al. 2000). Scattering experiments of unit cells performed in a TEM chamber provide requisite boundary conditions to mimic an infinite array. Although good results can be obtained this way, waveguide implementation limited one of the dimensions of the LHM structures on top of having practicality issues. Researchers in Li et al. (2003) for the first time described a focused beam method for free space measurements of LHM structures. A typical setup for a free space S parameter measurement of an LHM structure is shown in Fig. 15(a). Figure 15(b) shows the measured S parameters of an LHM material designed to operate at 14.2 GHz.

Ozbay et al. investigated the transmission and reflection properties of SRR medium, thin wire medium, and combined metamaterial medium separately (Shelby et al. 2001). A left-handed passband was demonstrated within the stop bands of the SRR and thin wires medium. This is consistent with the simulation results discussed in the previous section. The measured transmission and reflection characteristics for the three cases are shown in Fig. 16.

Negative refraction in an LHM was first demonstrated in Shelby et al. (2001) using a 2D metamaterial structure. The metamaterial structures shown till now can

Table 1 Brief summary of work done in the field of design of metamaterial lens

Research work	Motivation	Key features	Frequency
D. R. Smith et al. Phys. Rev. Lett. 84, 4184 – Published 1 May 2000	Demonstrate LHM behavior for the first time	Studied two cases of incident H polarization on SRR medium in waveguide chamber	4.6 GHz
R. A. Shelby, Science, Vol. 292, Issue 5514, 06 Apr 2001	Negative refraction demonstration	Demonstrated negative refraction using an LHM wedge and Teflon wedge. Evaluated negative refractive index using Snell's Law	10.5 GHz
G.V. Eleftheriades et al., IEEE Transactions on Microwave Theory and Techniques, vol. 50, no. 12, Dec. 2002	Alternative approach to designing of LHM using L-C loaded transmission lines	Demonstrated negative refraction and subwavelength focusing in a planar LHM medium. Periodic L-C loaded transmission lines are used instead of SRRs and wires	1.5 GHz
Kin Li et al. Appl. Phys. Lett. 82, 2535, Apr 2003	Free space characterization of LHM	Showed a focused beam method for experimental characterization of LHM. Agreed well with HFSS infinite unit cell model	14.2 GHz
E. Ozbay et al., IEEE Transactions on Antennas and Propagation, vol. 51, no. 10, Oct 2003	Transmission and Reflection measurements of LHM	Measured S_{21} and S_{11} for three separate mediums: SRRs only, wires only, and combined metamaterial structure	9.5–14.5 GHz
Koray Aydin et al., Optics Letters Vol. 29, Issue 22, Nov 2004	Experimental demonstration of true left-handed peaks	Compared closed and split ring resonators (SRR) to demonstrate true left-handed nature of the passband due to a combined SRR and wire medium	3.9 GHz
Koray Aydin et al., Optics Express, Vol. 13, No. 22, Oct 2005	First subwavelength focusing demonstration using a 2D LHM lens	Demonstrated subwavelength resolution beyond the diffraction limit is achievable using a 2D LHM lens	3.9 GHz
M. Gokkavas et al., Physical Review B, 73, May 2006	First millimeter wave LHM design	Designed an LHM material at 100 GHz using the conventional SRR and wire arrays	100 GHz
E. Ozbay et al., Journal of Physics Condensed Matter, 20, Jul 2008	Subwavelength focusing demonstration using a 1D LHM lens	Achieved a very high subwavelength focusing resolution of 0.2λ	3.8 GHz
D. Shreiber, Sensors and Actuators A: Physical, Volume 144, Issue 1, Jan 2008	Metamaterial lens for NDE of composites	Detect a defect of 0.037l in a composite	3.5 GHz

(continued)

Table 1 (continued)

Research work	Motivation	Key features	Frequency
Gang Wang et al., International Journal of Hyperthermia, Oct 2017	Metamaterial lens for biomedical applications	Demonstrated that microwave hyperthermia with the LHM lens can reach deeper seated tissues	2.45 GHz
S. Datta. et al., Journal of Nondestructive Evaluation, Diagnostics and Prognostics of Engineering Systems, Vol. 3, May 2020	Metamaterial lens for NDE of composites	Enhancement of microwave detection capability in the far field for subwavelength defects in composites	3.4 GHz

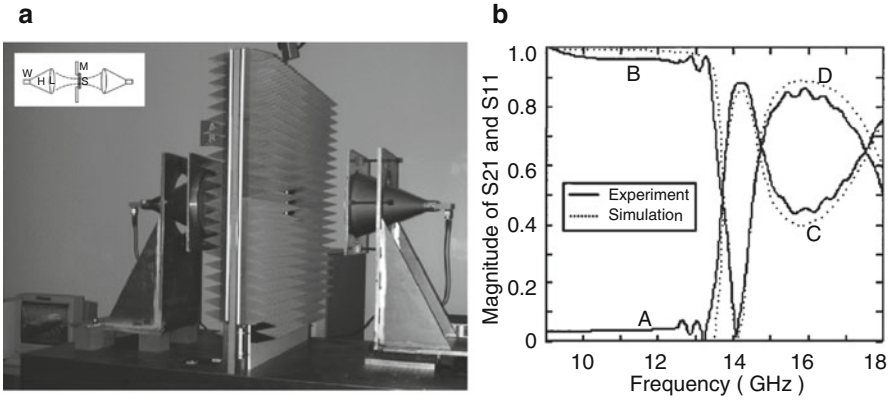

Fig. 15 (**a**) Typical experimental setup used for free space S-parameter measurement of metamaterial. This particular system focuses the beam in the sample region to increase the signal to noise ratio. The setup schematic is shown in inset of figure [8]. (**b**) Measured and simulated S parameters for a left-handed material slab. Line A: measured transmission; Line B: measured reflection; Line C: simulated transmission; and Line D: simulated reflection. (Li et al. (2003))

respond to magnetic field only in a particular directional plane, determined by the axis of the SRRs. Thus, from the perspective of the degree of freedom of the magnetic interactions, this unit cell structure can be categorized as a one-dimensional (1D) metamaterial element. To increase the degree of freedom of metamaterial functions so that they respond to different angles of incident wave, additional differently orientated structures can be designed as shown in Fig. 17. To determine the refractive index, the deflection of a beam of microwave radiation is measured as the beam passes through a wedge-shaped sample. Experiments were performed with a wedge-shaped LHM sample, as well as with a similarly shaped Teflon sample as a reference. The deflected angle measurements were used to determine the refractive index using Snell's law. Negative refraction can also be demonstrated via a gaussian beam shift experiment as described by Aydin et al. (Aydin et al. 2005).

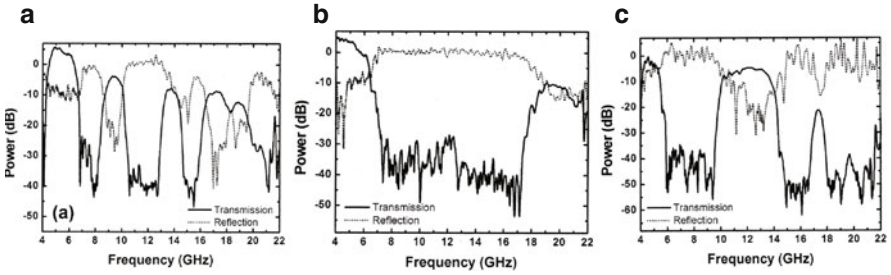

Fig. 16 (**a**) Measured broad-range transmission and reflection spectra of SRR medium in free space. The transmission spectrum exhibits series stopbands and passbands. The negative permeability regions do not allow the propagation of electromagnetic waves through the SRR structure (Shelby et al. 2001); (**b**) measured transmission and reflection characteristics of the thin wire medium. The transmission spectrum exhibits a wide stopband extending from 7 to 18 GHz. The lower passband is observed due to the discontinuous nature of the wires. The reflection data also indicates the strong rejection of electromagnetic waves from the structure for the negative permittivity region (Shelby et al. 2001); and (**c**) measured transmission and reflection spectra of the DNG composite

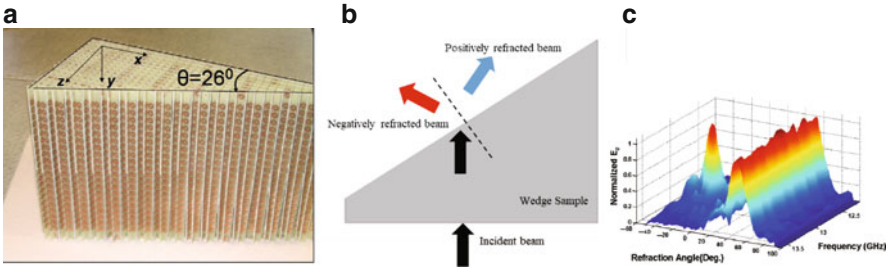

Fig. 17 (**a**) A typical 2D wedge-shaped LHM sample. (Aydin et al. (2006)); (**b**) negative and positive refraction in a wedge-shaped sample; (**c**) surface plot of measured normalized peak amplitude of electric field component for a Teflon and an LHM wedge. Note that the electric field refracted by the Teflon wedge peaks at a positive refractive angle of 48.2 (corresponding to an index of refraction of 1.4) and is independent of frequency. The electric field refracted by the LHM wedge, however, peaks at negative angles that are a function of the frequency (since the LHM has negative refractive index only in a certain frequency band). (Parazzoli et al. (2007))

After Pendry, in his seminal paper, discussed the theoretical possibility of obtaining subwavelength imaging by virtue of metamaterial lens, experimental work on demonstration of subwavelength focusing followed (Aydin et al. 2005, 2006); Ozbay et al. 2008); Aydin et al. 2007). EM waves emerging from an omnidirectional source located near such an LHM will first come into focus inside the material and then outside the lens as discussed earlier. High angular frequency wave components are amplified in a metamaterial lens, thus allowing subwavelength focusing beyond the diffraction limit. A typical experiment schematic for subwavelength-focusing demonstration is shown in Fig. 18(a). Measured power scans exhibiting subwavelength focusing are shown in Fig. 18(b) and (c). Extensive research on validating negative refractive index behavior of metamaterials paved the

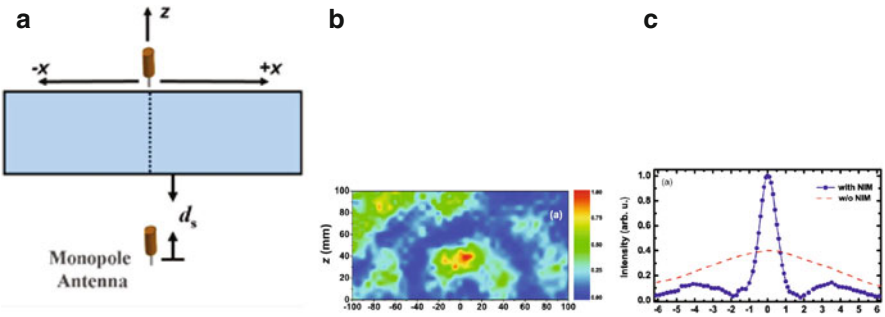

Fig. 18 (**a**) Schematic drawings of the top view of the experimental setup for verifying LHM lens focusing (Aydin et al. (2005)); (**b**) measured transmission spectra along the x-z plane for a point source located at $d_s = 39$ mm (Aydin et al. (2005)); and (**c**) the measured power line scan at the focal image plane with (blue) and without (dashed red) negative refractive index (NIM) superlens. This shows that the fields are more focused by using an LHM material than that of free space. (Aydin et al. (2007))

way for its huge potential in various novel applications. Schreiber et al. reported for the first microwave NDE application of a metamaterial lens (Shreiber et al. 2008). A fiberglass sample with a hole of subwavelength diameter was evaluated by keeping the sample at the focal point of 1 2D LHM lens. The experimental setup and results are described in Fig. 19. The authors also did a comparative study of 1D and 2D metamaterial lens for NDE of composites (Shreiber et al. 2011). 1D metamaterial lenses were shown to exhibit higher transmission and allow greater probe standoff distances. 2D lens however showed improved subwavelength focusing over a 1D lens.

Datta et al. recently reported an enhanced microwave detection capability using a metamaterial lens (Datta et al. 2020). The authors demonstrated increased sensitivity to subwavelength defects in composites in the left-handed pass band in comparison to free space measurements. One healthy sample and two defect samples with varying defect dimensions were tested. The experimental setup along with the results is described in Fig. 20. The results reported in the work are preliminary, and further extensive applications of Metamaterial lens for NDE of composites are being studied by the authors.

In conclusion, even though negative refractive index and subwavelength focusing are demonstrated in literature, one needs to keep in mind that they are valid only for a strict regime of parameters. Inherent losses and periodicity associated with a fabricated metamaterial force it to deviate from Pendry's perfect lens condition. The limitations for subwavelength imaging have been studied by Smith et al. in detail (Smith et al. 2003).

Microwave devices inspired from a typical metamaterial lens described above have also been performed for NDE applications. Some specific examples have been briefly described below.

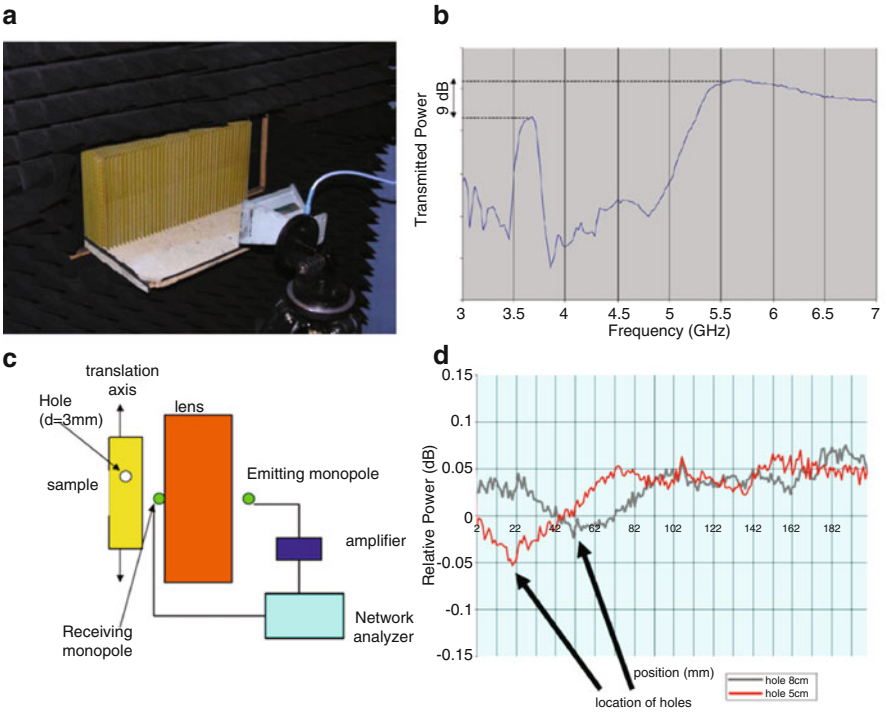

Fig. 19 (**a**) Experimental setup for lens characterization (Shreiber et al. (2008)); (**b**) measured transmission characteristics of 2D LHM lens. Left-handed pass band is observed around 3.65 GHz. The pass band after 5.5 GHz is due to simultaneously positive e and ^ at these frequencies, where traditional transmission occurs (Shreiber et al. (2011)); (**c**) sketch of the NDE sensor system using an LHM lens (Shreiber et al. 2008); and (d) relative power scan of a sample when two holes were drilled alternately is shown. The LHM lens was able to successfully determine the position of the two holes (Shreiber et al. 2008)

Conical Swiss-Roll Lens as an NDE Sensor

A conical Swiss-roll metamaterial lens was utilized as an EM NDE sensor operating at MHz frequencies in Savin et al. (2015) for detecting defects in composite materials. The lens is constructed of a number of spiral-wound-insulated conductor on a conical mandrel, as shown in Fig. 21. To decrease the losses at high frequencies, the insulated conductor layers and the polyamide layers are usually laminated without adhesive. Grimberg et al. (2011) propose a nondestructive method which employs the lens as a radio-frequency flux concentrator. The method is used to examine carbon fiber-reinforced plastic (CFRP) plates with potential delaminations caused by low-energy impacts. As shown in Fig. 22, the woven material's structure can be recognized and the delaminated area can be observed in the central zone.

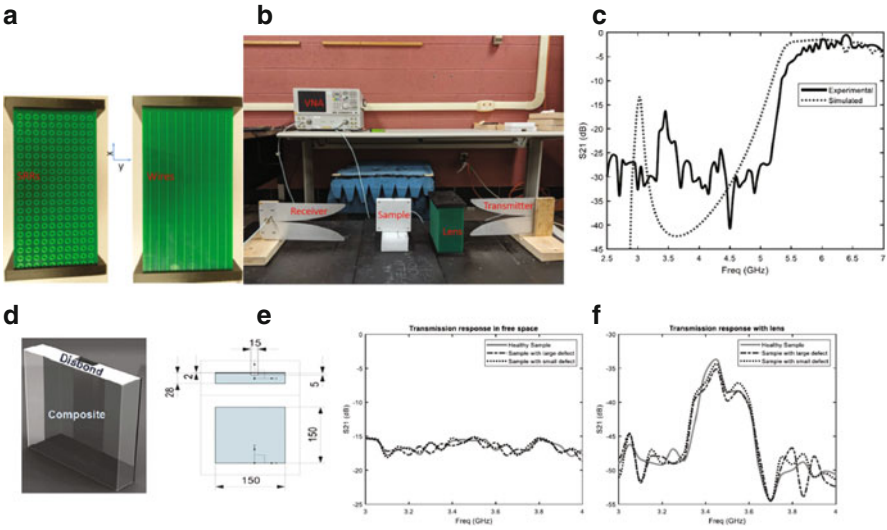

Fig. 20 (**a**) Fabricated 1D LHM lens; (**b**) microwave NDE setup using a metamaterial lens; (**c**) experimental and simulated transmission characteristic of the fabricated lens. The shift in frequency in the design can be attributed to error tolerances in fabrication. A left-handed pass band is obtained around 3.4 GHz; (**d**) composite sample under test. Teflon samples with a machined groove along the length to mimic disbonds in metal-composite joints were used. Three samples (one healthy and two with varying defect dimensions) were tested with two sets of reading from each sample. The first set of readings was in free space without the lens while the second set of measurements was taken with the lens; and (**e**) transmission response measured using in free space and using metamaterial lens. In the case of measurement with the lens, we observe a clear trend at the left-handed transmission peak at 3.44 GHz. The sample with a larger defect shows a greater change relative to the sample with a small defect. The free space measurements do not show any such trend (Datta et al. 2020)

Frequency Selective Surfaces for Strain and NDE Measurements

Frequency Selective Surfaces (FSSs) have also been inspired from the field of metamaterials for NDE applications. FSS is any repetitive surface with periodic arrays of elements that are designed to reflect, transmit, or absorb a particular resonant response when illuminated with high-frequency electromagnetic energy. The element geometry, inter-element spacing, and substrate properties of the FSS will determine the resonant response (Munk 2005). FSSs have been widely used in all kinds of applications such as high-frequency filters (Li et al. 2012); Sarabandi and Behdad 2007), reflectors, lenses, radomes, and absorbers (Munk 2005); Edalati and Sarabandi 2013). FSSs have also shown potential as sensors for NDE applications (Kinzel 2014). For example, in Jang et al. (2012), a high-precision crack detection is achieved by monitoring the shift in the resonant frequency of the FSS-based sensor. An FSS-sensing system has been utilized for normal and shear strain sensing in Kinzel (2014). An FSS is used to evaluate the layered structures with potential

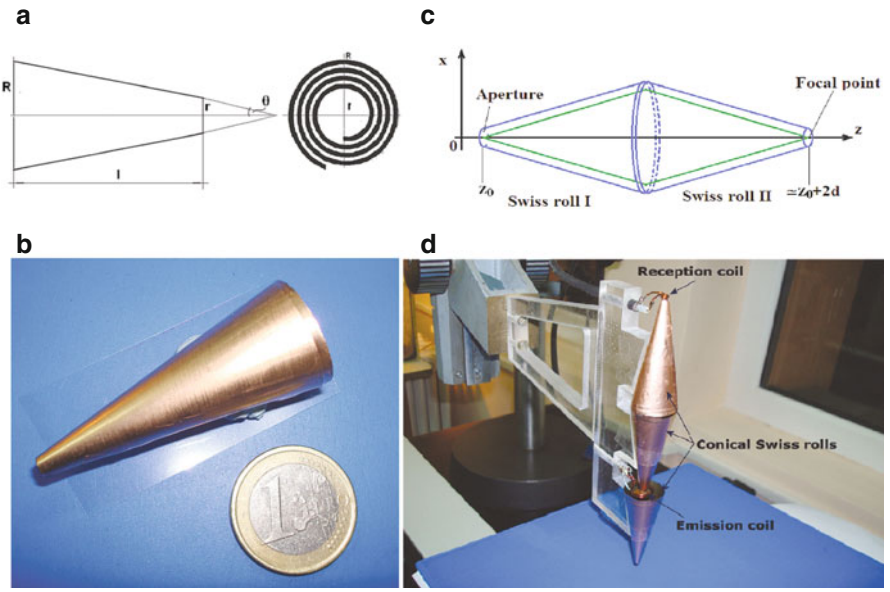

Fig. 21 Conical Swiss roll: (**a**) schematic of a conical Swiss roll; (**b**) photograph of a conical Swiss roll; and (**c**) and (**d**) the electromagnetic transducer with metamaterial lens. (Adapted from Grimberg et al. (2011))

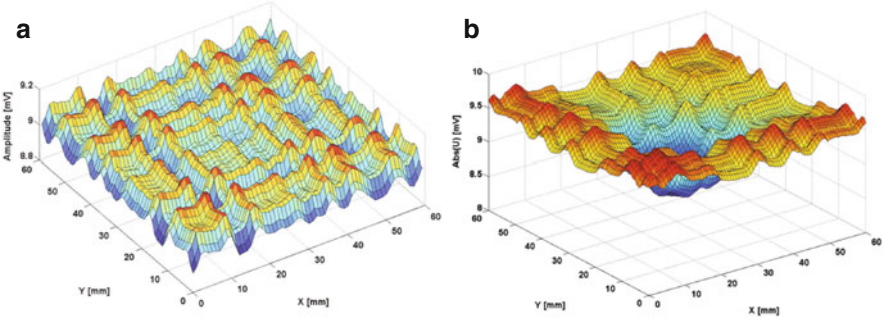

Fig. 22 The measured amplitude of the signal delivered by the electromagnetic transducer with metamaterials from the scanning of a CFRP zone: (**a**) without delamination, and (**b**) with delamination due to a 2.5 J energy impact. (Adapted from Grimberg et al. (2011))

delaminations or disbonds. In Mahmoodi et al. (2020), Mahboobeh et al. show the optimization of FSS-based sensor for the strain sensing. The sensor is shown in Fig. 23(a, b), and the sensor response is shown in Fig. 23(c). The measurements were performed when the antenna is in line with the strained area of the sensor and offset (in the negative x-direction).

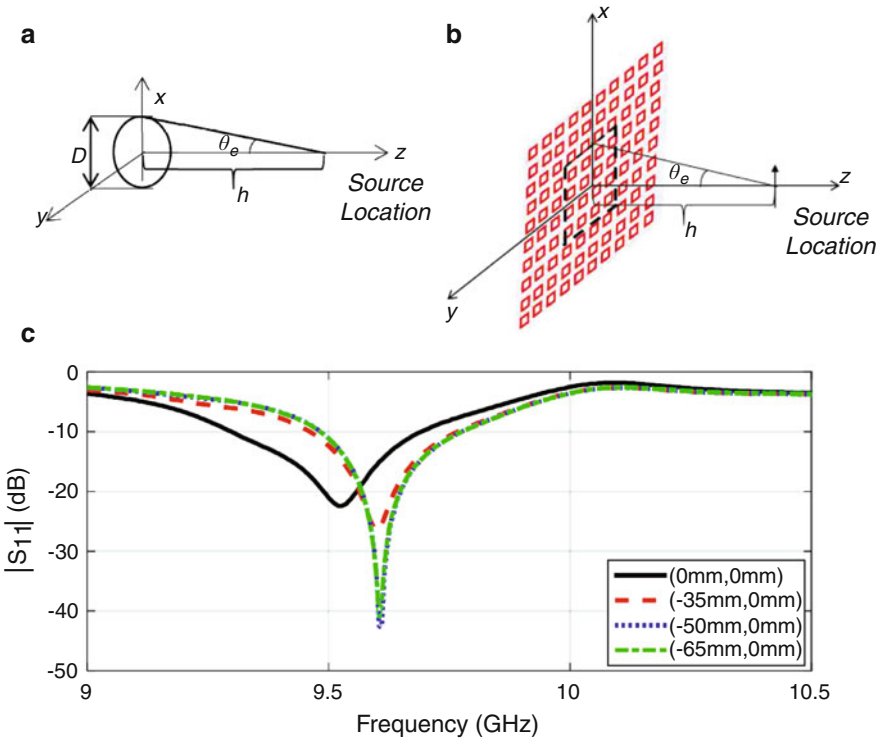

Fig. 23 (**a**) The schematic of aperture and illuminating source geometry for a reflector antenna; (**b**) schematic of a large FSS sensor with 3 × 3 strained elements indicated in the black square. The dimensions of the unit cell within this area have been increased by 5% along the y-direction in order to model the effect of unidirectional localized strain; and (c) measured —S11— of FSS sensor at different location. Adapted from Mahmoodi et al. (2020)

Conclusion

The motivation for developing a robust NDE technology for rapid inspection of composites is critical for a wide variety of applications. This chapter summarizes the breadth of research done in the field of metamaterial lens tailored toward microwave NDE of composites. The discussion is done from theoretical, computational as well as experimental aspects. Due to the long wavelengths associated with microwave fields, a metamaterial lens may become significant in enhancing the resolution of microwave imaging due to its unique subwavelength properties. The physics of the problem is understood from a holistic point of view, by discussing a computational framework for simulating the lens and discussing the geometrical optics of the lens. Numerical modeling of the unit cell in commercial EM simulators is followed by methods to extract the electromagnetic properties of the lens. Finally, state-of-the-art

experimental methods and measurement results are discussed to understand the current progress and also highlight the key challenges in this field.

This discussion sets a platform for determining future research needs in this field in order to serve as a robust NDE technique. Although extensive research has been done in characterization and demonstration of metamaterial behavior, their applications on NDE has been relatively less, especially for far field microwave NDE. The perfect lens envisioned by Pendry to break diffraction limit can only be realized when real part of $n = -1$ and imaginary part is 0. However, inherent losses associated with fabricated LHM lenses make its physical implementation for far field still an issue. Hence, research is developing low loss-flexible LHM structures. Finally, inverse imaging algorithms, which are shown to work well with numerical implementations of homogenized metamaterial lenses, need to be studied for their potential applications in subwavelength defects imaging using fabricated metamaterial lens. Moreover, finite element models of whole lens structure instead of homogenized approximation will allow one to study inverse imaging algorithms numerically. Due to the computational complexity associated with modeling the whole lens structure, numerical techniques to reduce the computational burden of such models should be investigated. Finally, a material parameter extraction technique based on microwave measurements of fabricated finite LHM slabs assisted by finite element models should be developed to study ambiguities associated with material properties extraction and determine the electromagnetic properties of the lens with higher accuracy.

Appendix

Inversion of S parameters is an established method for the experimental characterization of unknown materials Nicolson (1968). Using this approach, an inhomogeneous metamaterial structure, assumed to be a homogenous medium under the effective medium theory, can be characterized by an effective refractive index n and effective impedance z (Smith et al. 2005; Chen et al. 2004). Relatively simple analytic expressions relate n and z of the metamaterial to the S parameters. The procedure to extract the constituent properties of metamaterials is summarized in the following steps:

Step 1: Calculation of z from S parameters

The effective impedance z of a metamaterial unit cell is related to its scattering parameters by the equation,

$$z = \pm\sqrt{\frac{(1+S_{11})^2 - S_{21}^2}{(1+S_{11})^2 + S_{21}^2}} \tag{15}$$

The ambiguity in sign can be resolved by using prior information about the metamaterial. Since the metamaterial is a passive device, the reflected power in a surface cannot exceed the incident power. Therefore, real part of $z > 0$, which in turn resolves the sign ambiguity in equation.

Step 2: Calculation of refractive index. The refractive index of a metamaterial unit cell is related to the S parameters by the following:

$$e^{ink_0d} = \frac{S_{21}}{1 - S_{11}\frac{z-1}{z+1}} \tag{16}$$

The value of refractive index (n) can be evaluated as,

$$n = \frac{1}{k_0 d}\left\{\left[\ln\left(e^{ink_0d}\right)'' + 2m\pi\right)\right] - i\left(e^{ink_0d}\right)]'\right\} \tag{17}$$

where (.)' and (.)" are the real and imaginary operators, respectively. k_0 is wavenumber, d is the maximum dimension of the unit cell, and m is an integer. The ambiguity in branch selection of the logarithmic function can be resolved by choosing correct integer value m. Discussion regarding the correct choice of branches for n of a metamaterial is beyond the scope of this book and can be found in Parazzoli et al. (2007). In case of the design presented in the modeling section, the principal branch (m = 0) can be chosen since it is an electrically thin material. Therefore, the above equations can be used to directly determine the n and z of the unit cell model in HFSS.

Step 3) Calculation of permittivity and permeability

The effective permittivity and permeability of a metamaterial can be computed from z and n using these relations:

$$\varepsilon = n/z \quad \mu = nz \tag{18}$$

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Metamaterial Filters

11

D. Vakula and A. Sowjanaya

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Abstract

Filters are one of basic blocks of any communication systems. Filters allow a certain band of frequencies and reject rest of the frequencies. The filters are classified based on the frequency selection. Planar filters are realized using microstrip lines. Microstrip filters are designed using a combination of transmission lines by modifying the alignment of the lines to either couple or reject the selected

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frequencies. The appropriate equations for design of microstrip line are included. But, these microstrip filters have some disadvantages in getting compact, multiple band and sharp rolls. Metamaterials have properties of negative permittivity and permeability, which can control the electromagnetic wave propagation in the medium. Instead of controlling the wave propagation only through microstrip line coupling, metamaterials are incorporated along the transmission line to control the electromagnetic wave propagation. As the bandpass filter is the widely used filter in the transmitter and receivers in communication systems, the design of metamaterial-based microwave bandpass filters is elaborated. Design of bandpass filter with metamaterial structures with coplanar waveguide transmission line is elaborated. The modifications to be done to the transmission line for obtaining bandpass filter are given. Design of bandpass filter with split ring resonators and complementary split resonators is explained. A dual-bandpass filter design with complementary split ring resonators is also illustrated. Novel metamaterial resonators are represented. The role of the gaps and vias along the microstrip line is also discussed. The simulation results of the designed filters are presented.

Keywords

Metamaterials · Frequency response · Passband · Stopband · Split ring resonators

Introduction

Filters are used to allow certain range of frequencies and block other frequencies. Filters have many applications in wireless communication systems, Direct current (DC) power supplies, in audio electronics, analog-to-digital converters, etc. In addition, any Radio Frequency (RF) and microwave communication system, radar, test and measurement system require filters at various stages.

To represent as an electrical network, filter consists of two ports, transmits frequencies in passband and will not pass frequencies in stopband. To achieve the desired characteristics of filters, the network elements are designed based on the impedance matching of network, known as image parameter method. In this, the impedance of network is matched to source impedance in passband, where the input impedance of network is real. But, the stopbands attenuation is not large at all frequencies.

The desired frequency response can be achieved by designing the network based on network synthesis techniques, where the network is determined for the given frequency response. This method is called as insertion loss method. The frequency response is selected based on the polynomial. The network elements are determined for the required transfer function. The network is initially designed for low-pass filter for prototype design with normalized impedance and cutoff frequency and later converted to the desired frequency and impedance level.

Filters with microstrip technology have become popular due to low profile, compact structure, and easy integration with RF circuitry. Microstrip filters are

designed using transmission lines of quarter wavelength, which act as resonators to allow the passband of frequencies (Zhu and Menzel 2003; Deng et al. 2006, 2013; Wu et al. 2005; Cheng et al. 2013). U-shaped hairpin resonators are used on two layers with aperture coupling to design a bandpass filter (Djaiz and Denidni 2006). Resonators are also designed with multimode stub loaded on the microstrip line (Chen et al. 2013) and open loop with fragment loading on the microstrip line (Zhao et al. 2014) to allow a band of frequencies.

The filters designed with resonators have compact size, well-defined pass and stopbands with multiple poles in transmission band and also transmission zeros in stopbands. But the filters have complex structures and achieving compact size with sharp passband characteristics is difficult to realize. Artificial materials with controllable electromagnetic properties are developed using miniature resonators of sub wavelength dimensions called as metamaterials. These metamaterials resonators are split ring resonators (SRR) and complementary split ring resonators (CSRR). The SRR and CSRR change the electrical properties like permittivity and permeability of the materials to control the electromagnetic wave propagation along the materials. A major breakthrough has come in the design of filters using metamaterials, which are known as metamaterial filters. A bandpass filter is designed for a frequency range of 2.8–3.1 GHz using split ring resonators and metallic vias between feed line and ground plane (Gil et al. 2005). A double negative medium is generated with negative permeability and negative permittivity. The bandpass filter has a fractional bandwidth (FBW) of 4.1% and insertion loss in passband is 3.54 dB, which is a large value. A compact filter is designed by varying gaps of the split ring resonators. The passband is between 1.83 and 2.13 GHz with insertion loss of 1.3 dB. The fractional bandwidth is 15.2% (Li et al. 2015). Open-loop resonator is coupled to the microstrip line to get a passband from 2.38 to 2.58 GHz. The fractional band width of the filter is 8%. Another bandpass filter is determined with SRRs loaded along the microstrip line from 2.4 to 2.6 GHz and has a fractional bandwidth of 8% (Jayakrishnan et al. 2016).A wide stop is achieved with double-sided open split ring resonators and the number of transmission zeros in stopband are also increased, which can attenuate frequencies in spurious band (de Dios Ruiz and Hinojosa 2012). Circular-shaped resonators are placed on the top and back sides of the substrate, and U-shaped slots are placed along the transmission line. The filter has passband bandwidth of 1.7–2.6 GHz with FBW of 42.8% with an insertion loss less than 1 dB. Coplanar waveguide transmission line is loaded with split ring resonators and metallic lines to design a bandpass filter with passband from 7.8 to 8.1 GHz (Martin et al. 2003). The proposed bandpass filter has fractional bandwidth of 3.7% at center frequency 7.95 GHz with an insertion loss of 4 dB. Second-order and third-order split ring resonator-based coupled bandpass filters are proposed in Ahmed et al. (2018). The proposed second-order bandpass filter operating in the frequency range from 2.1 to 2.55 GHz with FBW of 19% at center frequency 2.31 GHz with insertion loss of 0.7 dB. The proposed third-order bandpass filter frequency range is from 2.15 to 2.6 GHz with FBW of 19% at center frequency 2.36 GHz and insertion loss is 1.5 dB. Complementary split ring resonators are used for design of bandpass filter. The resonators consist of series gaps and shunt stubs. The bandpass filter is designed from 2.38 to 2.65 GHz with fractional bandwidth of 9%.

The insertion loss in the passband for the filter is 1.9 dB (Bonache et al. 2006). A wideband, bandpass filter is developed with split ring resonators with series gaps and the model is further improved with grounded stubs (Gil et al. 2007). The proposed hybrid approach results in wideband range from 3.5 to 10 GHz with FBW of 109% with center frequency of 5.9 GHz, insertion loss of 3 dB. Complementary S-shaped split ring resonators are used for filter design using differential microstrip technology. A compact design is proposed and common mode noise is also suppressed (Horestani et al. 2014a). The passband ranging from 0.9 to 1.1 GHz with FBW of 20% at center frequency of 1 GHz, insertion loss of about 2.8 dB. A design of high temperature superconducting structure based on split ring resonators is presented in Tan et al. (2018), which results in passband from 2.48 to 2.53 GHz with FBW of 2.14% at center frequency of 2.515 GHz, insertion loss of 0.1 dB. Shunt capacitive and series inductive loading is implemented to the basic edge coupled bandpass filter and for folded edge coupled bandpass filter in Lim et al. (2014), which results in passband from 0.88 to 0.92 GHz with FBW of 4.4% at center frequency 0.9 GHz, with insertion loss of nearly 3 dB for basic design and passband from 1.85 to 2 GHz with FBW of 7.7% with 1.5 dB insertion loss.

A compact bandpass filter is determined with four CRLH TL resonators, which resulted in passband from 4.93 to 4.97 GHz with 2.7 dB insertion loss (Abdalla et al. 2015). Coplanar waveguide transmission lone is loaded with S-shaped split ring resonators and series gaps are presented in Horestani et al. (2014b) to design compact bandpass filter, which results in passband range from 1.15 to 1.3 GHz with FBW of 12% at center frequency 1.25 GHz with 2.4 dB insertion loss. A split ring resonator with step impedance is presented in Zhou et al. (2011) to design a compact, stopband expanded bandpass filter with four transmission zeros, high selectivity, which results in passband range from 2.15 to 2.3 GHz with FBW of 6.8% at center frequency of 2.2 GHz, insertion loss of about 1.7 dB. A second-order filter is designed with SRRs that improved selectivity of the filter in Fouda et al. (2018), which results in passband range from 2.3 to 2.5 GHz with FBW of 5.4% at center frequency 2.4 GHz, with insertion loss less than 0.5 dB. Folded stepped resonators coupled to coplanar waveguide are used to design differential mode bandpass filter and good selectivity and suppression of common mode is achieved (Fernández-Prieto et al. 2016). The proposed filter operates from 1.48 to 1.53 GHz with FBW of 5% at center frequency 1.5 GHz, 1.6 dB insertion loss. Coplanar waveguide transmission is introduced with gaps in between the line. The gaps are loaded with split ring resonators to design bandpass filter in Sanz et al. (2014), which results in passband ranging from 2.85 to 3.12 GHz with FBW of 8.9% at center frequency 3 GHz, 0.86 dB insertion loss.

In the chapter the use of metamaterials for design of microwave filters is presented. The application of metamaterials improves the characteristics of the filters and is also used to obtain compact structures. Metamaterials are frequency selective surfaces that have resonant structures. The resonant structures are made of rings with series gaps which are known spilt ring resonators and circular slots with series gaps which are called as complementary split ring resonators. Band Pass filter with SRRs or CSRR are demonstrated.

Classification of Filters on Frequency Response

The filters can be classified into four categories based on frequency response.

Low-Pass Filter

The lower frequencies will be allowed without attenuation, and high frequencies are attenuated. Ideally, at a particular frequency called cutoff frequency, the attenuations have to vary from low to high value. The frequencies less than the cutoff frequency are allowed in low-pass filter, which is passband. The frequencies higher than cutoff frequency are attenuated, which is stopband. The passband and stopband separation is given by cutoff frequency. In practical filters, at the cutoff frequency, the transition of attenuation is gradual. So, the cutoff frequency from passband to stop is defined as the frequency at which attenuation is 3 dB above the attenuation in passband. The filters consist of only reactive elements. The impedance of network varies with frequency and voltage drop in the network varies when compared with the voltage drop across load.

High-Pass Filter

High-pass filter has complementary characteristic of low-pass filter, where lower frequencies are attenuated and allows higher frequencies. In ideal filter, the transition from high attenuation to zero attenuation occurs at single frequency called cutoff frequency. But, a transition happens over a range of frequencies. The cutoff frequency is the frequency at which attenuation is above 3 dB of attenuation in passband. In high-pass filter, below cutoff frequency stopband exists and above cutoff frequency passband exists.

Bandpass Filter

In contrast to low-pass and high-pass filter, there is need to allow only certain range of frequencies, which is bandpass filter. There are two frequency limits for bandpass filter. The frequencies below and above the limits are not allowed. These frequencies are cutoff frequencies. A stopband is present below the first cutoff frequency; the first cutoff frequency is lower cutoff frequency. Another stopband is present above the second cutoff frequency, which is higher cutoff frequency. A passband is present between the lower and higher cutoff frequency. In practical filters, the lower cutoff frequency is the frequency at which attenuation is 3 dB above the attenuation in passband and higher cutoff frequency is the frequency that occurs after passband frequencies, where the attenuation is 3 dB above the passband attenuation. The bandpass filter can be assumed as cascades of high-pass and low-pass filter if the cutoff frequency of low pass filter is greater than the cutoff frequency of high-pass filter.

Then the frequencies between the two cutoff frequencies will be allowed without attenuation and above and below cutoff frequencies will be attenuated.

Band-Stop Filter

In some applications, there is requirement to not allow certain range of frequencies. In this requirement again, there are two cutoff frequencies. This type of filter attenuates the signal between the two cutoff frequencies, which is also known as **Band-Stop filter** or **Notch Filter**.

The band-stop filter has two passbands and one stopband. The frequency below the lower cutoff frequency will be allowed and frequencies above cutoff frequency will be allowed.

At lower cutoff frequency, the attenuation is above 3 dB of the attenuation in first passband. Above the lower cutoff frequency stopband starts. At second or higher cutoff frequency the attenuation is above 3 dB of the attenuation in second passband. Bandstop filter can be realized as parallel combination of low-pass and high-pass filters for which the cutoff frequency of low-pass filter must be lower than the cutoff frequency of high-pass filter.

The frequencies between the two cutoff frequencies will be attenuated and frequencies below and above will be allowed without attenuation.

Type of Response in Passband and Stopband

A two-port network is represented in Fig. 1, where $v_1(\omega)$ and$v_2(\omega)$ are voltages at the input and out ports. The transfer function of the network is represented as $H(\omega)$. The ideal transfer functions for low-pass, high-pass, bandpass, and bandstop filter are shown in Fig. 2. The amplitude of the response in the pass- and stopband is constant or it can vary within certain limits. The attenuation varies between pass- and stopband, the transition increases with the order of the filter. The variation of passband and stop attenuation for various orders of filters are shown in Fig. 3. The phase of transfer function varies in both pass- and stopbands with frequency. The two port network has to be matched to the source and load.

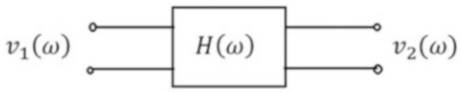

Fig. 1 Two port network representation

$$H(\omega) = \frac{v_2(\omega)}{v_1(\omega)}$$

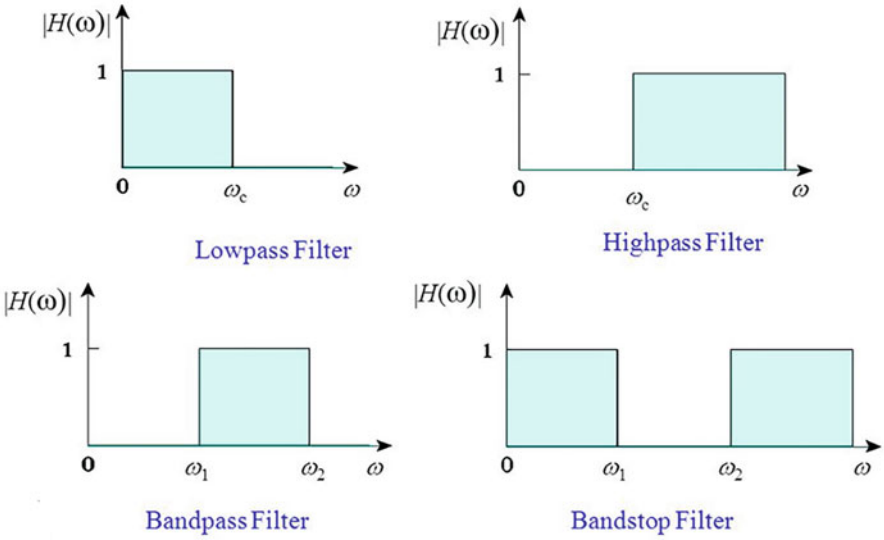

Fig. 2 Response of different types of filters

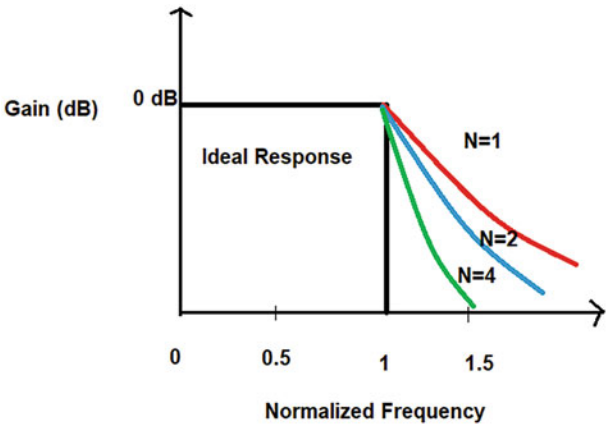

Fig. 3 Amplitude response of the filters with different orders

Microstrip Filter

Filters can be realized using lumped elements or distributed elements. Low-frequency filters are realized using lumped elements. Microwave filters are realized using distributed elements. Transmission lines or coaxial lines are used to realize distributed reactance values. Planar filters are realized using strip lines or microstrip lines. Filters designed with microstrip lines are discussed in the chapter.

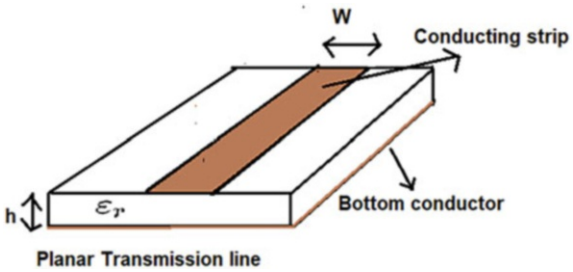

Fig. 4 Microstrip line

Design of Microstrip Line

Microstrip line is a planar transmission line, which consists of top conductor of width "w" and bottom conductor, which is referred as ground plane. The top and bottom conductor are separated by a dielectric of thickness "h." The dielectric constant of the substrate is "ε_r" as shown in Fig. 4. While the transmission of the signal takes place along the microstrip line, the fields due to current travel through the dielectric and also free space. So, the wave propagation is through quasi TEM mode instead of TEM mode. When the longitudinal components are much smaller than the transverse components, a quasi TEM mode is applicable to facilitate the design. And, also along the microstrip line, the TEM travels through dielectric medium and partially through air, hence the velocity is different from the velocity of the wave when the wave travels entirely though dielectric medium. The effective dielectric constant is determined from the dielectric constant of the substrate, height of the substrate, and width of the top conductor. The characteristic impedance of the line depends on the width of the top conductor, effective dielectric constant, and height of the substrate.

The design equations for design of microstrip line for a given characteristic impedance are given below (Garg et al. 2001).

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2\sqrt{1 + \frac{12h}{W}}} \tag{1}$$

$$Z_0 = \frac{60}{\sqrt{\varepsilon_e}} \ln\left(\frac{8h}{W} + \frac{W}{4h}\right) \quad \text{for} \quad \frac{W}{h} \leq 1 \tag{2}$$

$$Z_0 = \frac{120\pi}{\sqrt{\varepsilon_e\left[\frac{W}{h} + 1.393 + 0.667 \ln\left(\frac{W}{h} + 1.444\right)\right]}} \text{ for } \frac{W}{h} \geq 1 \tag{3}$$

The parameters of transmission line are given by the following:
Guided wavelength is

$$\lambda_g = \frac{\lambda_o}{\sqrt{\varepsilon_{re}}} \text{ or } \lambda_g = \frac{300}{f(GHz)\sqrt{\varepsilon_{re}}} mm \tag{4}$$

The propagation constant is given by

$$\beta = \frac{2\pi}{\lambda_g} \tag{5}$$

The phase velocity is given by

$$v_p = \frac{\omega}{\beta} = \frac{c}{\sqrt{\varepsilon_{re}}} \tag{6}$$

The electrical length is given by

$$\theta = \beta l \tag{7}$$

Vias and Coupled Transmission Lines

Metallic Vias

Metallic vias are connected between transmission line and ground plane with a diameter "d" and spacing "s." The vias offer a capacitance and resistance. These vias create a negative dielectric constant, which can manipulate the wave propagation along the transmission line (Fig. 5). The spacing and thickness are modified to get a passband characteristic. These metallic vias act like single negative index material (Reja et al. 2015).

Coupled Transmission Line

Coupled lines are designed with half wavelength at resonance frequency. The number of coupled lines is decided by the order of the filter. Coupling can be done in different ways: Parallel coupling, edge coupling, and end coupling. Bandpass characteristic is achieved by the coupling between the line. Coupled line filter is the best option to design bandpass filter (Fig. 6).

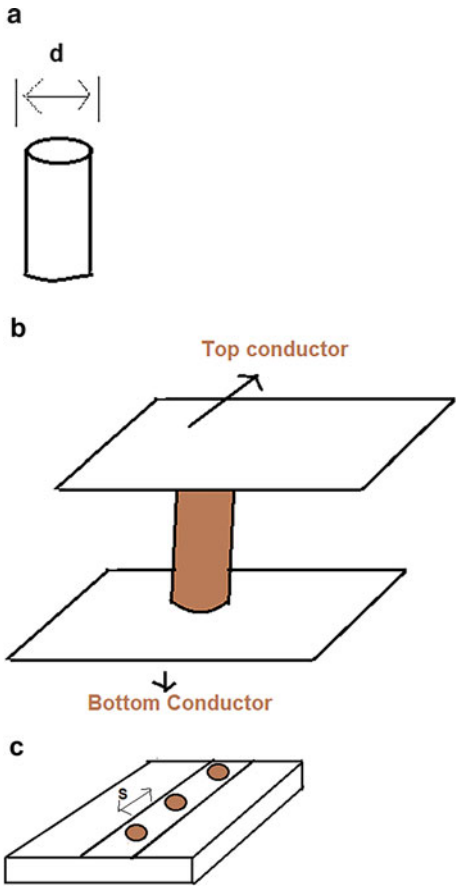

Fig. 5 (**a**) Via, (**b**) via connecting the top and bottom conductors, (**c**) Multiple vias with spacing "s"

The response of the filter is controlled by designing the cutoff frequencies between the pass- and stopbands. These cutoff frequencies are identified by attenuation poles or transmission zeros. The signal path between the input and out ports is controlled by the coupling along the transmission line with gaps and additional coupling structures at the Gaps. Open-loop resonators are used to enhance coupling at the gaps or along the transmission line. The open-loop resonators are designed at half wavelength at the desired frequency. The coupling between line and resonators can be done using electric coupling, magnetic coupling, and mixed coupling.

Metamaterial

Metamaterial structures control EM wave propagation through the dielectric medium by modification of the permeability and permittivity. The structures can

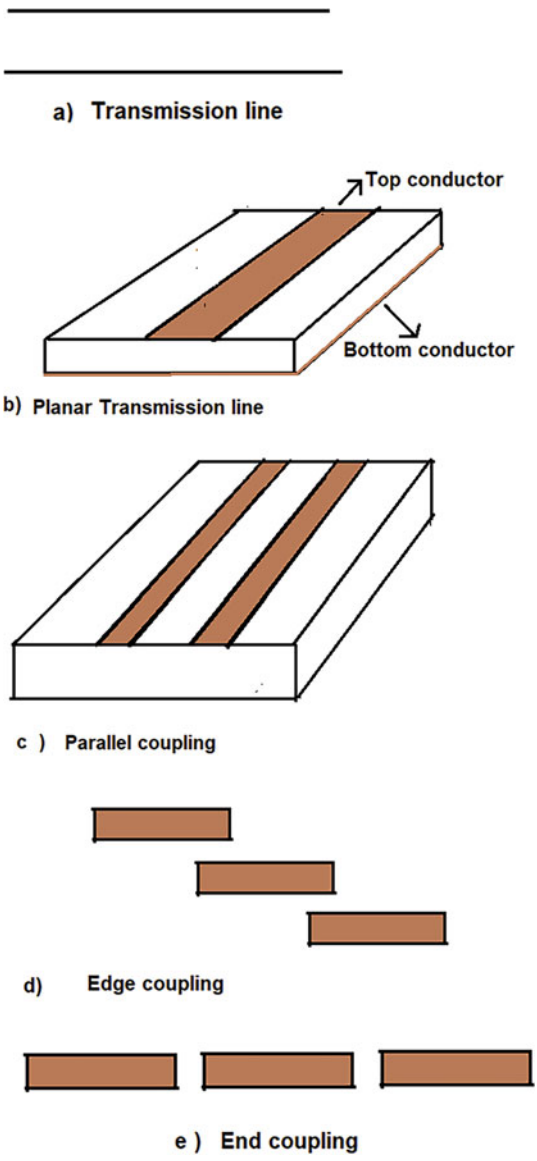

Fig. 6 (**a**) Transmission line, (**b**) planar transmission line, (**c**) parallel coupling between lines, (**d**) edge coupling, (**e**) end coupling

be designed to have either negative permittivity or permeability, which are called as single negative media (SNG) and both negative permittivity and permeability known as double negative media (DNG). Metamaterials are made using split ring resonators. Metamaterials are used to make compact filters. SSR have capacitive coupling between the spilt rings and series inductance along the rings (Fig. 7).

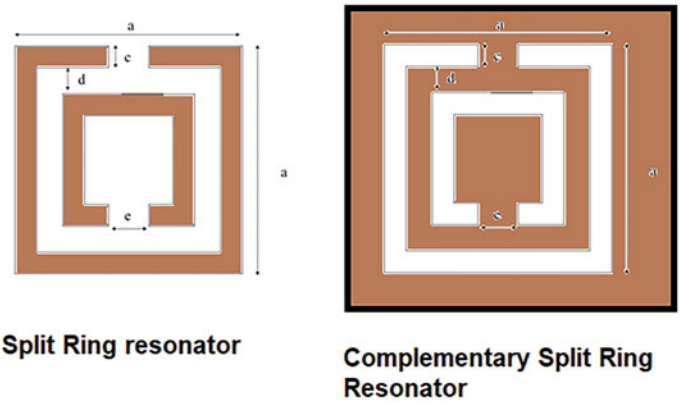

Fig. 7 Split ring resonators and complementary split ring resonators

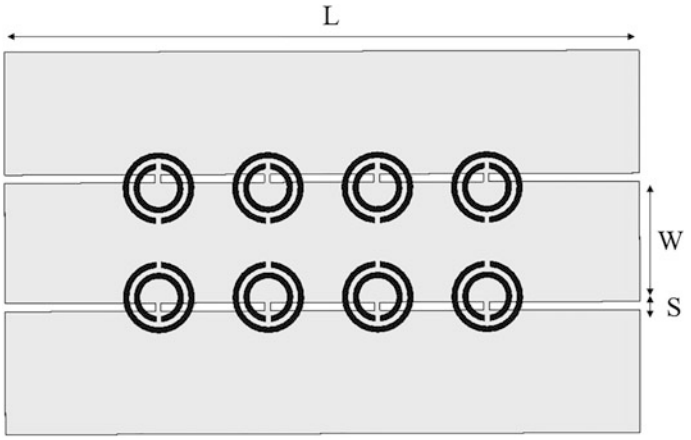

Fig. 8 Structure CPW loaded with SRR bandpass filter

Coplanar Waveguide Loaded with SRRS Bandpass Filter

A Coplanar waveguide (CPW) transmission line is considered for design of bandpass filter.

CPW transmission line allows all frequencies, and by introducing conducting strips between the feed line and ground plane, backward wave propagation occurs. Split ring resonators are added on the opposite side of the strips, which allows a band of frequencies (Figs. 8 and 9).

Negative effective permittivity is introduced along the CPW transmission line by introducing strip lines at appropriate positions. The magnetic field is generated surrounding the strips due to the current flowing along the line. The SRRs have to

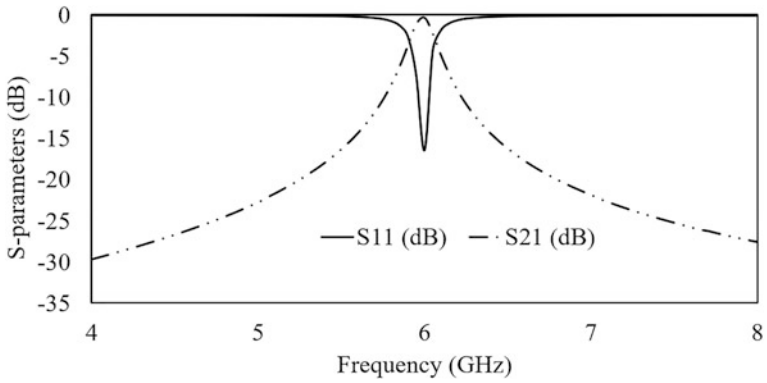

Fig. 9 S-Parameters of CPW loaded with SRR bandpass filter

be placed at such locations to introduce current in axial direction along the rings. A significant amount of axial current is to be introduced in the rings. The rings are excited by the induced axial currents and resonate at a particular frequency. At frequency above the resonance frequency an effective medium is created, which have negative permeability. SRRs can be etched as slots on the top conductor and they can also be placed at the bottom side of the substrate. The bottom side of the substrate has no metallization. The SRRs are placed at the bottom side of the substrate just under the strips. The strips behave like inductors which create a medium with negative permittivity and SRRs create medium with negative permeability. Hence, the combinations of negative permittivity and permeability act like double negative medium, which makes the wave propagate along the CPW transmission line. The strips generate the negative permittivity above the plasma frequency. If the plasma frequency is above the resonance frequency of the SRR, double negative medium exits at these frequencies. Signal is allowed through the line, which acts like Band Pass filter (Gil et al. 2008).

The Coplanar waveguide has a length of 23.2 mm, trace width of 4.2 mm, ground plane spacing of 0.3 mm for proper impedance matching to 50 Ω. The ring has a width of 0.2 mm, separation between rings is 0.2 mm, the inner ring radius is 1.1 mm. The rings are etched backside of RT Duroid 4003C material of dielectric constant of 3.55, loss tangent of 0.0027, height 1.524 mm, copper metallization of 0.035 mm. The proposed design has passband range from 5.95 to 6.025 GHz, with FBW of $<1\%$ at center frequency 5.98 GHz and with minimum insertion loss of 0.3 dB as shown in Fig. 9.

Microstrip Line Loaded with SRRs Bandpass Filter

Bandpass filter is realized with microstrip line vias and split ring resonators. Microstrip line is the planar transmission, which allows all frequencies. Current flows on the top conductor which has a magnetic field surrounding the conductor. Split ring

resonators are placed close to the top conductor. Due to the magnetic field, axial currents are induced in the split ring resonators. The effective permittivity of the medium changes and above resonant frequency, the medium has negative permittivity. In the proposed square split ring resonators are considered with spacing between the rings as "d." The size of the split ring resonator is considered as "a." The spacing of the splits in the resonator is "e" for the inner and outer ring. The thickness of the rings is "c." The dimensions of the ring and the spacing from the microstrip line decide the resonance frequency of the split ring resonator. The split rings are placed close to the microstrip line to improve the magnetic coupling. The placement of split ring resonators will inhibit the wave propagation above the resonance frequency due the negative permeability introduced in the medium. The wave can be allowed by introducing negative permittivity in the medium through vias between the top and bottom conductor.

Vias are introduced between the transmission lines, which introduce negative permittivity. Vias act as shunt inductance. When the negative permittivity and permeability exists at the same frequency range, then the medium behaves as double negative medium, which allows wave in the forward direction. A design acts like a band pass filter (Figs. 10 and 11).

The split ring resonators and the microstrip line are etched on the top of a dielectric substrate RT Duroid 4003C material having dielectric constant 3.55, dielectric loss tangent 0.0027, and copper metallization of thickness 0.035 mm on both sides. The microstrip line of width 1.4 mm is chosen for having proper impedance matching. The dimensions of split ring resonator unit cell are shown in Table 1. The size of the filter is 9.2 × 28.6 mm, which is approximately 0.9 × 0.3 λ_g, where λ_g is guided wavelength at the center frequency. Simulation results show that the proposed bandpass filter has passband range from 5.07 to 5.12 GHz. The square SRR has a fractional bandwidth of 1%. The center frequency of the filter is

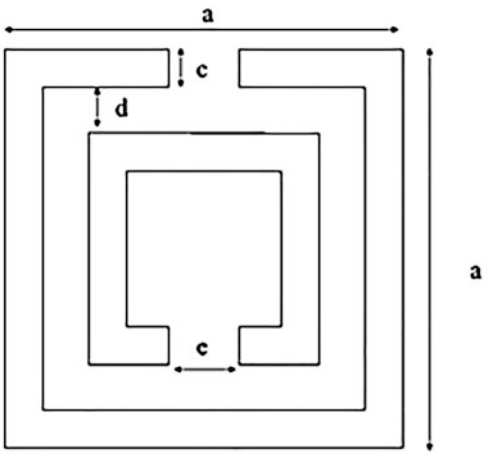

Fig. 10 Structure of split ring resonators

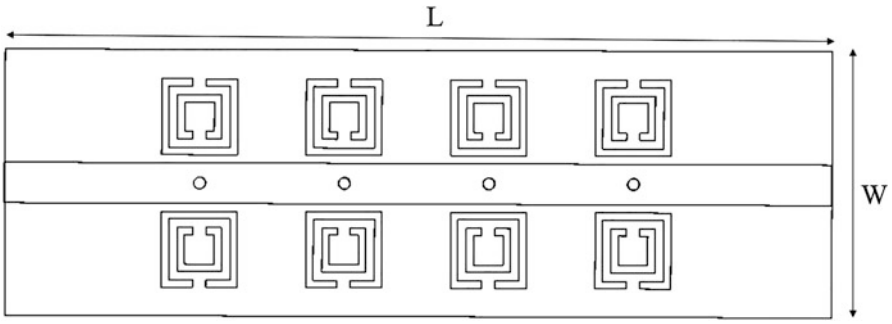

Fig. 11 Structure of microstrip line loaded with SRRs bandpass filter

Table 1 Dimensions of split ring resonator

Dimension	a	c	d	e
Value (mm)	2.62	0.25	0.3	0.46

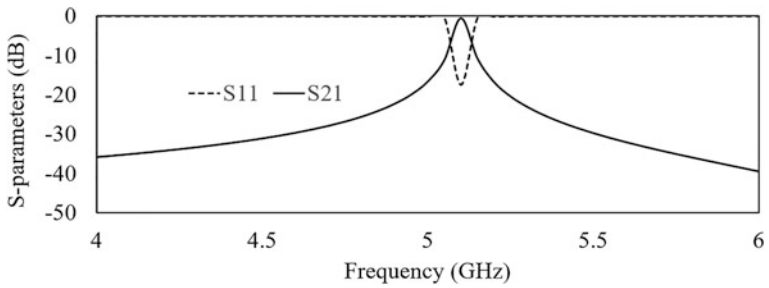

Fig. 12 S Parameters of microstrip line loaded with SRRs bandpass filter

5.09 GHz. The proposed filter has a minimum insertion loss of 0.6 dB. The return loss of the passband is 17.5 dB as given in Fig. 12.

Bandpass Filter Design Using EI-Shaped Resonator Structures

Transmission line is designed using the Eqs. (1), (2), and (3) shown above for proper impedance matching of the microstrip line with 50 Ω line. Designed transmission line acts as ALL PASS FILTER. Microstrip transmission line acts as the basic structure for signal transmission (Sowjanya and Vakula 2021). In the proposed bandpass filters, split ring resonators and vias are used for designing bandpass filter. Resonators are considered of EI shape, where an I shape is overlapped by an E shape structure. The structures introduce inductance, and the spacing between the shapes induces capacitance, which acts like a resonator. The EI-shaped resonator resonates at a particular resonant frequency and the coupling between the microstrip line and resonators make axial currents to flow. The size of the EI-shaped resonator is "a" and

width of the structure is considered "c" and spacing between the structures is "d." The dimensions of the EI-shaped unit cell are given in Table 2.

EI-shaped resonators create negative permeability in the medium where wave is propagating. Metal vias are connected between the top conductor and bottom conductor. These vias are placed at locations, which are lying between the positions of the EI resonators. Vias induce negative permittivity in the medium. Metamaterial-based EI-shaped resonator structures are incorporated along the transmission line for modifying the transmission charateristics. Resonator structures are placed side by side as shown in Fig. 13. The number resonators and vias placed along the transmission line decide the order of the filter. As the order of the filter is increased, the transition of attenuation from passband to stopband is sharp. The roll of the filter is sharp. But the insertion loss of the filter increases.

Table 2 Dimensions of proposed structure

Dimension	a	c	d
Value (mm)	2.6	0.2	0.2

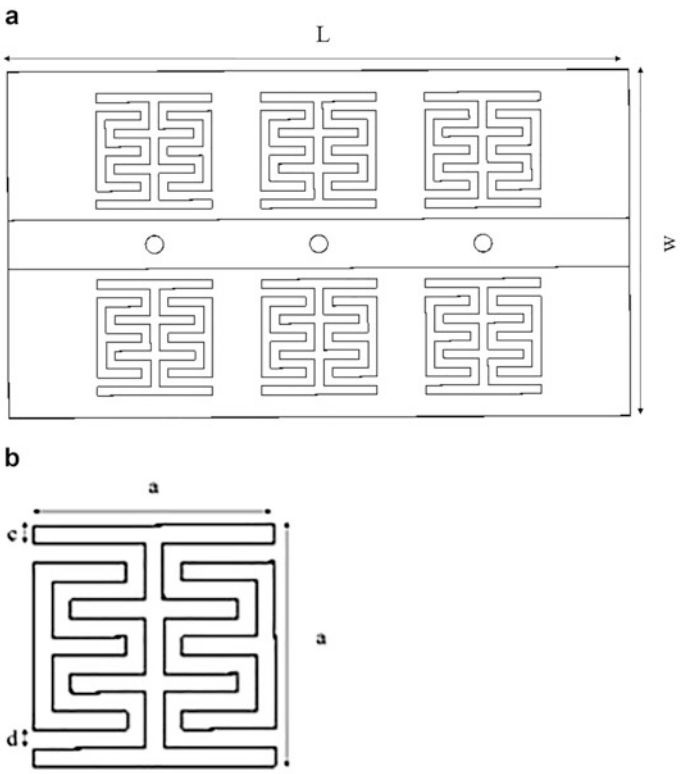

Fig. 13 (**a**) Top view of proposed layout of the bandpass filter, (**b**) unit cell

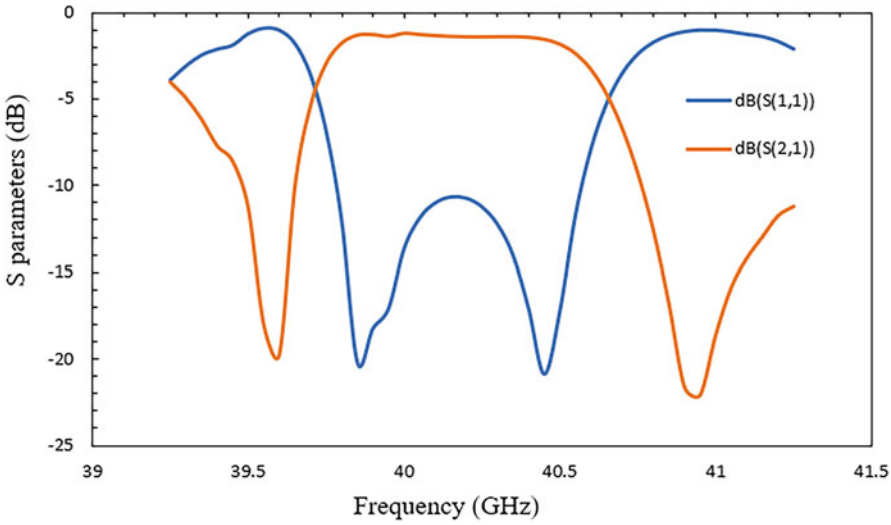

Fig. 14 Reflection coefficient (S_{11}) and transmission coefficient (S_{21}) of EI-shaped band-pass filter

The proposed filter is simulated by the finite element method (FEM) using software package ANSYS's HFSS. E bounded I-shaped metamaterial structure and microstrip line are printed on a high frequency laminate available from Rogers corporation, RT Duroid 4003C substrate material with 0.035 mm thickness copper patterns on both sides and substrate height of 0.813 mm. Dielectric constant of the material is 3.55 and dissipation factor is 0.0027. Microstrip line of width 1.1 mm is chosen to get better impedance matching with 50 Ω transmission line, and length of line is considered as 14 mm. Vias have a diameter of 0.4 mm and gaps between the line and resonators, in between resonators, are modified to enhance performance of the proposed filter design. S parameters of the simulated filter are shown in Fig. 10. The passband range of the proposed filter is 39.7–40.5 GHz having central frequency of 40.1 GHz, with fractional bandwidth (FBW) 2%. In-band insertion and return loss is better than 2.5 and 12 dB, respectively, as given Fig. 14.

Metamaterial Dual-Bandpass Filter

Initially a microstrip transmission line is designed with the following parameters. The material considered as substrate is Rogers RT Duriod 4003C with 0.813 mm as thickness. The width of the microstrip line is considered as 1.81 mm to get impedance matching for the BPF to 50 ohms. The size of the substrate is considered as 5.81 × 12.5 mm to accommodate the resonators. The microstrip line acts as all pass filter. The reflection coefficient (S_{11}) and transmission coefficient (S_{21}) are given in Fig. 15 for the transmission line.

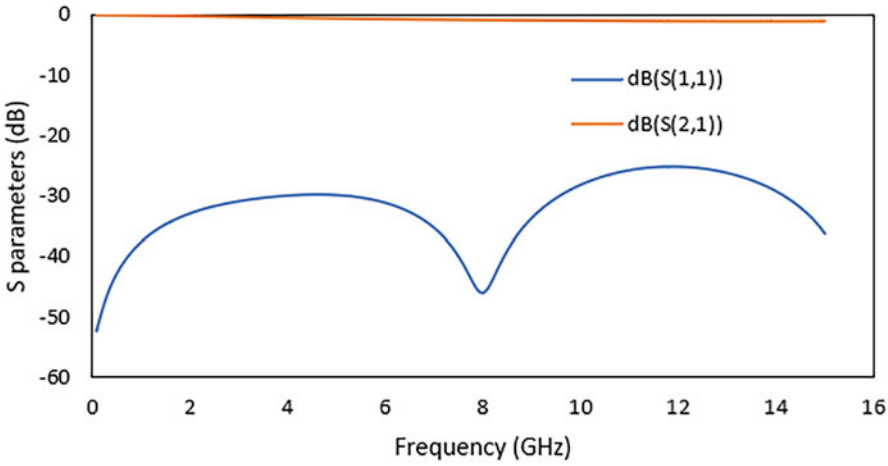

Fig. 15 Reflection and Transmission coefficient of microstrip line

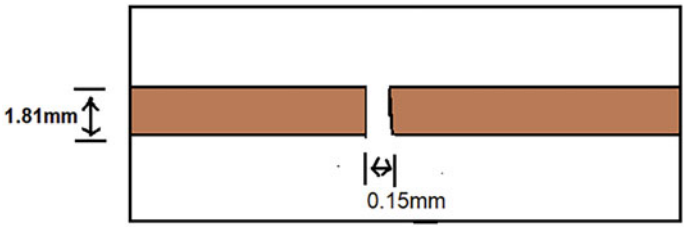

Fig. 16 Transmission line with end gaps between at a spacing of 0.15 mm

End Gaps are introduced in between the transmission; the gap width is 0.15 mm as shown in Fig. 16 and the reflection coefficient S_{11} and transmission coefficient S_{21} are given in Fig. 17. It can be seen that the transmission is interrupted. Now the coupling between the gaps is improved through complementary resonators in the ground plane. The resonator dimensions decide the passband frequencies. Two gaps are placed along the transmission line, and triple ring complementary resonators are placed on the ground plane at the gaps as shown in Fig. 18. The reflection coefficient S_{11} and transmission coefficient S_{21} are given in the Fig. 19. There are two passbands at 3 and 3.8 GHz. The strip width is optimized to 2.25 mm and size of the filter is 7.81 $\times$ 17.5 mm.

Conclusion

The design of metamaterials is vital for the design of metamaterial filter. The metamaterial resonators are designed to introduce negative permeability, and vias are used to have negative permittivity at the selected frequencies. The appropriate location of them along the transmission line also plays a major role in controlling the

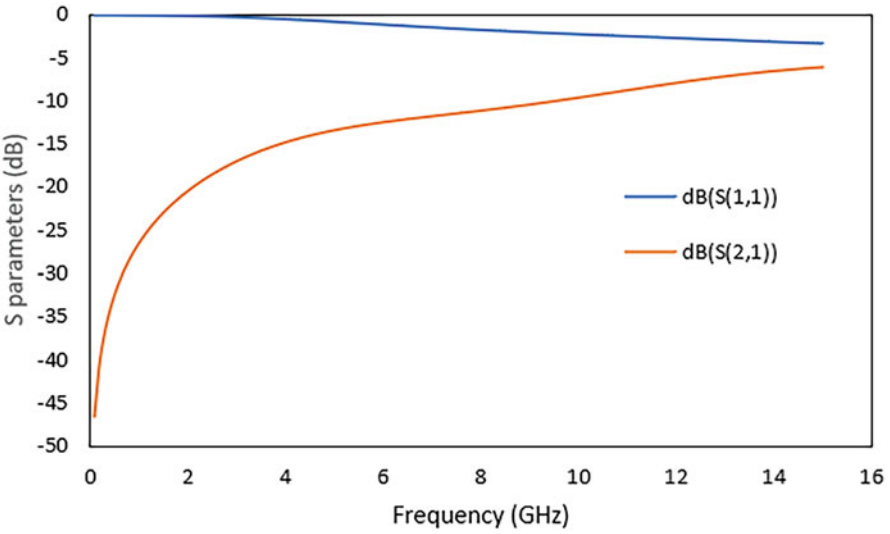

Fig. 17 Reflection and Transmission coefficient of transmission line with end gaps

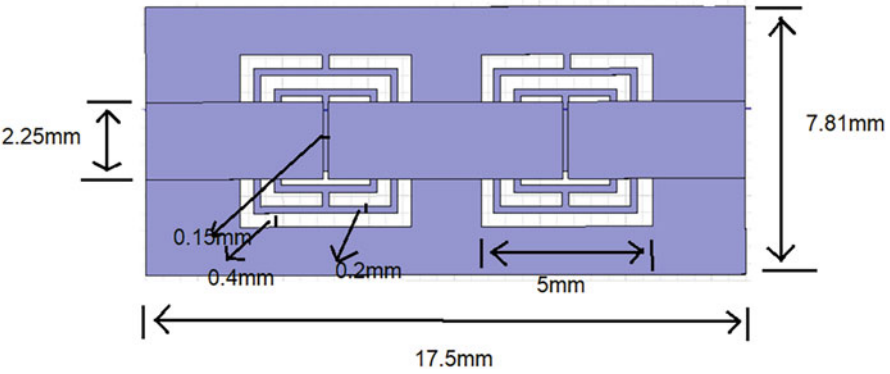

Fig. 18 Structure of dual-band metamaterial-based complementary triple spilt ring resonator

performance of the filters. As the number of resonators increases, the roll of the filter improves, but the insertion of the filter increases simultaneously. The design of planar transmission line is illustrated. The design of bandpass filter with metamaterial-based structures and complementary split ring resonators are discussed. Using the metamaterials, the size of the filters can be reduced. Metamaterial filters have many applications and conventional filter design can be further improved introducing metamaterial structures in the filters. As it is, metamaterials are resonant structures with narrow bandwidth; techniques can be developed to improve the band width of operation. The size of metamaterial resonators is in

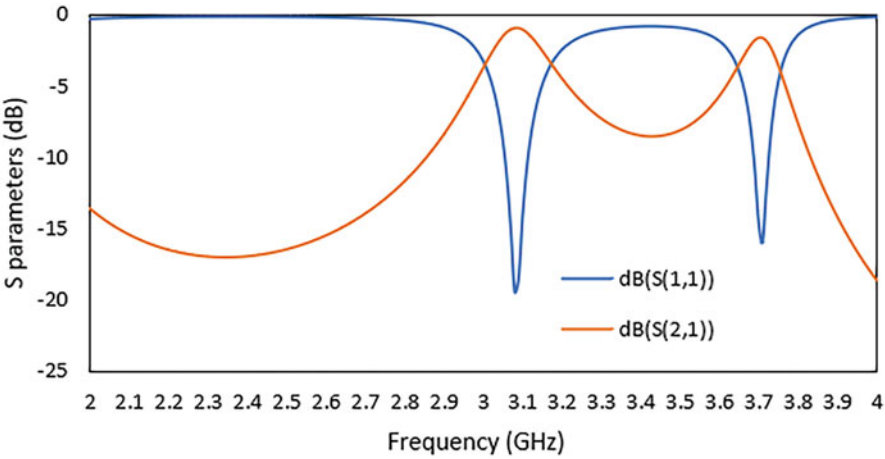

Fig. 19 Reflection and transmission coefficients of dual-band metamaterial-based complementary triple spilt ring resonator

terms of the wavelength of operation and higher frequencies have compact structures, which reduce the size of the filters drastically. The designs presented in the chapter can also be done for design of bandstop filters. The combinations of various resonators can be attempted to obtain novel designs. Metamaterials will revolutionize the filter designs with compact and high-performance structures.

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Part III

Wireless Communication Systems

Compact and Broadband Printed Antennas 12

G. Shrikanth Reddy

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Abstract

In recent years device industries has shown substantial interest in low-profile and conformal circuits to meet the requirements of evolving wireless technologies and standards. There is an increasing interest towards development of integrated/on chip antennas especially for wireless devices. Achieving broad impedance and radiation bandwidth from a compact antenna has been a challenging task. There are several research groups who pioneered in development of the compact broadband antennas for wireless devices. This chapter discusses techniques used to design compact and broadband antennas. The design techniques discussed in this chapter includes design of broadband printed microstrip and slot antennas, Metamaterial inspired antennas, Antennas integrated with periodic structures, and composition of resonators for compact antenna design. Some of the examples presented in this chapter also pertains to the millimeter wave

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antennas where metamaterial inspired configurations are used. Here, the term broadband antenna is associated with those antennas which provides consistent impedance bandwidth i.e., the return loss <10 dB (Voltage standing wave ratio-VSWR <2) and consistent radiation bandwidth i.e., characteristics such as gain, efficiency and radiation pattern are consistent for a given impedance bandwidth. Most of the antennas discussed here are printed antenna i.e., they are printed on a low-loss dielectric substrate with specific permittivity. As the value of substrate's permittivity contributes towards antenna design, its selection for the antenna design is done based on end use/application of the antenna.

Keywords

Compact antenna · Broadband antennas · Metamaterial Inspired antennas · periodic structures · and resonators

Introduction

Conventional wireless devices such as mobile phones, laptops, i-pads, routers, and dongles employs printed antennas of different type. These antennas in general occupies maximum surface area on a printed circuit board (PCB). Hence, antenna designers take a tradeoff between antenna's radiation performance and size to increase antenna's integrability (conformality) with rest of the circuit within a wireless device. The importance of compact antennas will further increase as in coming years the communication device industry will either see a further reduction in the device size or integration of multiple application within one device. To to keep up with evolving industrial standards it will be necessary that the system integrable antenna should exhibits broadband characteristics with a substantial level of conformality.

Antenna size reduction and simultaneous enhancement in bandwidth are major challenges. In recent years, several broadband and compact antennas have been reported. The major work in terms of compactness and bandwidth enhancement is seen in the form of Compact broadband Microstrip antennas, Metamaterial inspired antennas, antennas loaded with periodic structures and integration of radiative resonators. In this chapter the term broad-band antenna is commonly referred to an antenna with consistent impedance bandwidth and radiation characteristics for a broad operating bandwidth. This chapter is divided into four sections which are based on different antennas design techniques and examples. In section "Compact and Broad-band Printed Microstrip Antennas" microstrip or slot based printed antennas such as printed monopole and Vivaldi antennas are presented. These antennas exhibit wideband to ultra-wideband impedance bandwidth by means of closely spaced resonance. Section "Metamaterial Inspired Compact Antennas" discuss about metamaterial inspired antenna design with some examples on millimeter wave antennas. Here, millimeter wave antennas of both broadband and wideband configuration are presented which caters 60 GHz applications. The design of

millimeter wave antennas is inspired by unit cells elements such as Split-ring resonators (SRR), and periodic structures. In section "Metamaterial Inspired Compact Antennas" the techniques developed for millimeter wave configuration are further scaled towards lower frequency to validate the impedance and radiation characteristics. Apart from millimeter wave antennas, resonator based electrically small antennas (ESA) is also discussed briefly in section "Resonator Based Compact Antenna." In this section the discussed ESA is designed by integrating multiple resonant configurations. The ESA design using presented technique can be a single broadband or a multi band configuration. In conclusion this chapter provides a brief overview of the trends in compact and broadband antenna design which cater the requirements of wireless device industry.

Compact and Broad-band Printed Microstrip Antennas:

Printed microstrip antennas (PMA) are one of the most popular antennas used widely in wireless devices. PMA can be categorized into multiple subcategories such as radiating patch antennas, printed monopole or dipoles, and printed slot type like Vivaldi antennas. These antennas are generally multimode resonators where all the higher order (or mode) resonances are closely placed. The broadband behavior PMAs can be understood by observing their design methodology and field distribution. The early reported work on compact broadband PMAs (Wong 2002) involves, Slots and shorted edges in a microstrip antenna configurations (Hsu and Wong 2001; Kuo and Wong 2001). Further, loading of antennas with certain lumped elements also improves the impedance bandwidth (Wong and Lin 1997). The bandwidth of the radiating patch antennas is increased by placing parasitic elements near radiating or non-radiating edges of the patch antenna as reported in (Girish and Ray 2003). These techniques improve the impedance bandwidth in a limited fashion as the primary radiating elements have very narrow bandwidth. This narrow bandwidth is due to patch and ground plane induced capacitive reactance. To design a printed broad band microstrip antenna it is important that the input impedance of the antenna is less reactive and the antenna configuration support multi- mode resonance (closely spaced). To generate multimode resonance in printed microstrip antenna several techniques have been reported. Among these techniques partial ground plane backed printed monopole and flared exponential slots configurations are the most used antenna design techniques. In this section printed monopoles and Vivaldi antenna configuration are discussed briefly.

Broad-band Printed Monopoles

The radiating patch antennas are usually narrow band due to high capacitive contribution by radiating patch and the ground plane. The printed monopole configuration reduces this capacitive contribution by eliminating the portion of the ground plane behind the radiating patch. The radiation characteristics of the printed

monopoles are generally omnidirectional. The wide band mechanism of planar monopoles is reported/ presented in (Girish and Ray 2003; Agarwal et al. 1998). As discussed in (Balanis 2005; Ray and Ranga 2007; Mishra and Mukherjee 2012; Ray 2008) the wide band nature of the planar monopole can be explained in two ways:

(a) Bandwidth of a monopole can be increased when its effective diameter is increased (Agarwal et al. 1998). Therefore, the planar monopole can be considered as cylindrical monopole with large effective diameter.
(b) The planar monopole can be considered as microstrip antenna with very thick substrate or air suspended metallic patch antenna. Since in metallic patch various higher order modes with wide bandwidth can get excited, the shape of the planar monopole can be modified or optimized to bring all the modes together to achieve a wide impedance bandwidth.

The printed monopole antennas reported in (Girish and Ray 2003) are kind of planar metallic plate placed vertically on a horizontal (flat) metallic ground at a small distance 'P' and fed with the coaxial transmission line. The transition of simple monopole to the printed monopole is presented in Fig. 1.

The printed monopole presented in Fig. 1 can be designed using simple empirical equation presented in (Girish and Ray 2003). These relations can be designed as:

$$2\pi RL = WL \tag{1}$$

$$\text{This gives} : R = W/2\pi \tag{2}$$

The relations presented in Eqs. 1 and 2 corresponds to correlation between a wire monopole and a rectangular monopole. Similar correlation is devised for other

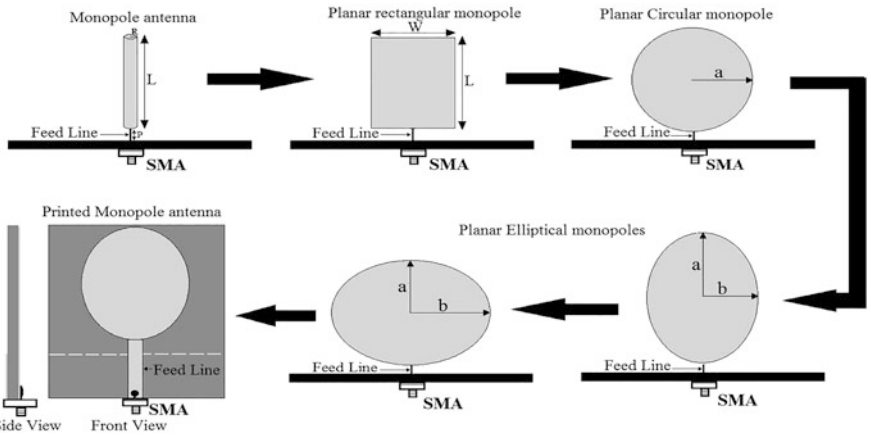

Fig. 1 Evolution of planar monopole antenna

configurations of monopoles as shown in Fig. 1 by considering relation between lowest resonant frequency and the principle radiative dimensions:

$$f_L = 7.2/(L_{cm} + R_{cm} + p_{cm}) \tag{3}$$

In Eq. 2 'f_L' represent lowest resonant frequency, L is the length of monopole, R is the radius of the monopole and p is the gap between lowest edge of monopole and the ground plane. By using Eq. 3-dimensional relation for circular and elliptical configurations are given as:

$$\text{For circular monopole} : L = 2a, r = a/4 \tag{4}$$

$$\text{For elliptical monopoles} : L = 2a, r = b/4 \tag{5}$$

$$\begin{gathered} Or \\ L = 2b, r = a/4 \end{gathered} \tag{6}$$

$$\text{For printed circular monopole} : f_L = 7.2/((L_{cm} + r_{cm} + p_{cm}) \times K) \tag{7}$$

In Eq. 7 the 'K' is the normalizing factor usually depends on the dielectric substrate used to design the monopole (Ray 2008). Using these simple relations, the printed monopoles can be designed. Apart from using the partial ground configurations the printed monopoles can be feed using CPW lines or coaxial feed configurations depending on the device or circuit requirements.

The working of the printed planar monopole antenna can be understood with a simple example. In Fig. 2 a simple printed monopole antenna is presented with a partial ground configuration. Here the dimensional values are given as : $L_S = 25$, $W_S = 28$, a = 7, G = 7, b = 9, L = 9, W = 2, t = 0.5 (All dimensions are in mm), the substrate material used here is RT Duroid 5880 $\varepsilon_r = 2.2$.

In Fig. 2 it can be noted that the first resonance occurs at the frequency for which the vertical length of the monopole (length considered above the ground plane) is quarter wavelength. This monopole is a multimode resonator where both radiator and the ground plane configuration contributes towards the excitation of the multiple modes. It may be noted that the ground plane width is twice the dimeter of the disc. The nature of the radiation pattern for a printed monopole can be understood through its field distribution as discussed in (Liang). As reported in (Liang) the circumferential/peripheral length of the radiator and the width of the ground plane ('Ws' in Fig. 2) are the major contributor towards the wide impedance bandwidth of a printed monopole antenna. The surface currents along these dimensions generally give rise to either oscillatory fields or standing wave type fields. These fields further contribute towards the radiation characterizes of the printed broadband monopole (Fereidoony et al. 2011).

The dimensions of the printed monopole can be further reduced by optimizing the circumferential length and the width of the ground plane. In (Reddy et al. 2014a) the surface area of the presented monopole is 24 × 17 mm^2, where the vertical length of the monopole is less than quarter-wavelength for lowest operating frequency. In this

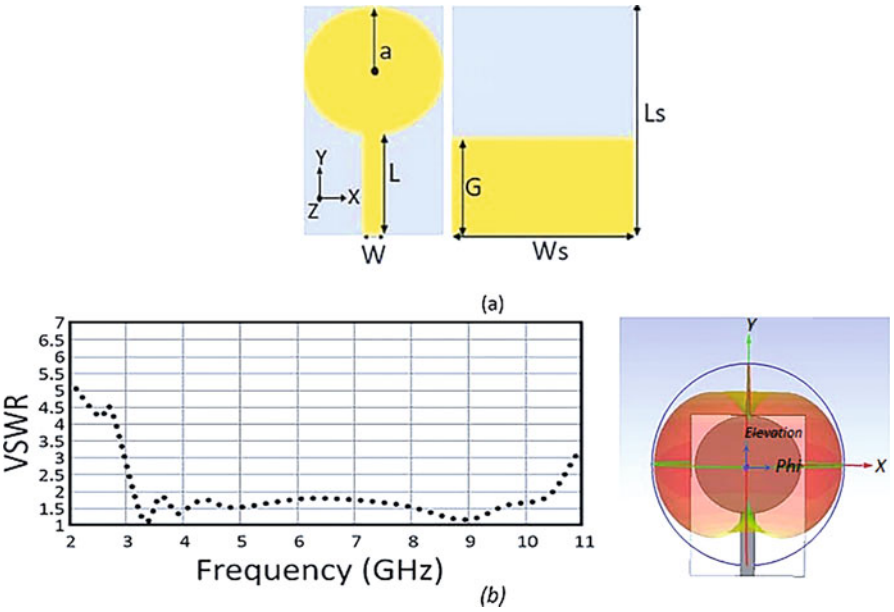

Fig. 2 (**a**) Printed circular monopole antenna and (**b**) bandwidth and radiation characteristics response

work a semi-circular disc is used which is half the size of a full disc. As shown in Fig. 3, the half disc monopole combined with a spiral resonator is designed to produce a dual-band response. This antenna is designed on a duriod substrate with permittivity of 2.2 with thickness of 0.787 mm.

It can be observed from Fig. 3 that the bandwidth of the antenna depends on the electrical length of the configuration. Here, semicircular disc, spiral resonator and the partial ground plane are responsible for wide impedance bandwidth. The printed monopole antennas serve best for wide band applications such as ultrawide-band communications and imaging radars. In Fig. 3 the compact configuration is achieved by modifying the main radiating element (which is directly connected with the feed line). However, the ground plane dimensions can also be modified to have a wide-band transition in impedance bandwidth.

As presented in Fig. 4, the printed monopole antenna uses asymmetric ground plane to achieve a wide-impedance bandwidth along with directional radiation characteristics (Reddy et al. 2014b). It can be observed in Fig. 4 that the ground plane of the CPW feed is modified in two stages. In stage B, the ground plane is transformed into a quarter circle with radius 'R_2'. This transformation extends the surface current path along the edges of the ground plane, which further extends the lower bound of the resonant frequency. Further, in stage C the vertical height for one of the sides of the ground plane is extended to accommodate a circular curve of radius 'R_3'. This asymmetric ground plane (stage C) act as reflector for the radiating

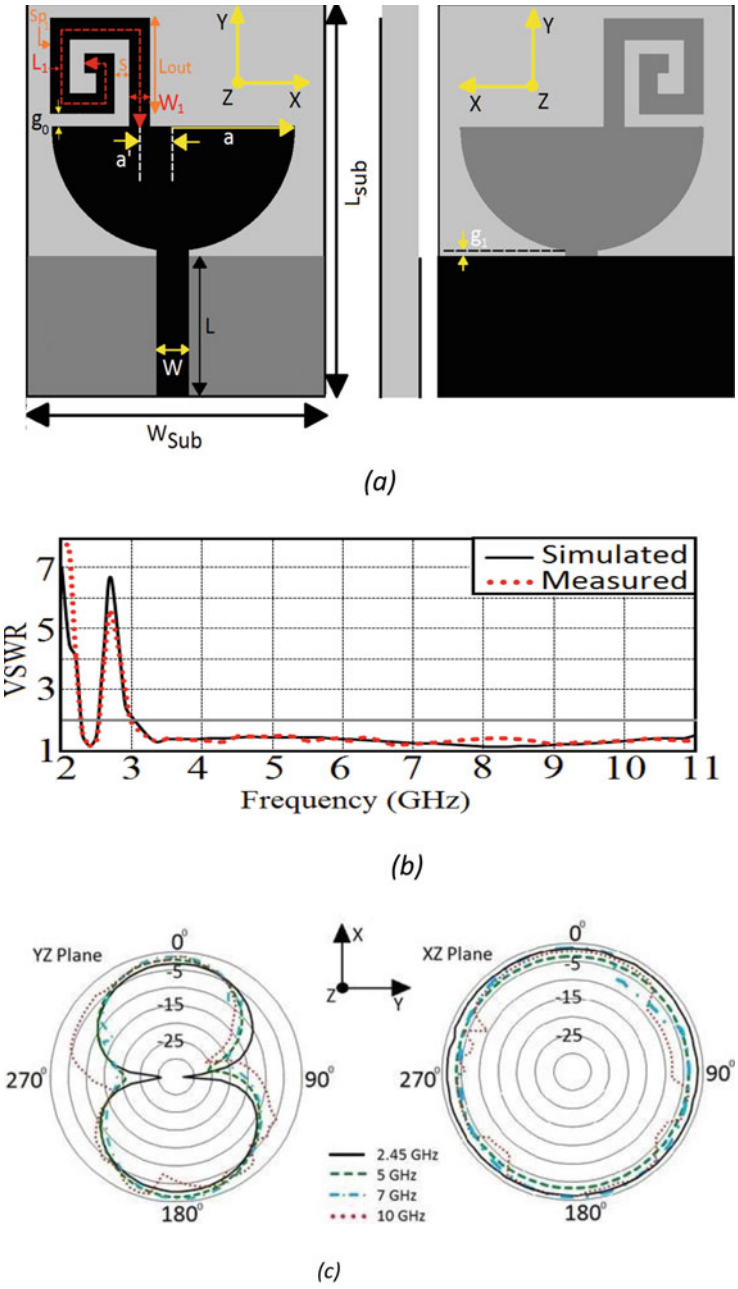

Fig. 3 (**a**) Front, Side and back view of Semi-circular printed monopole, (**b**) VSWR characteristics and (**c**) Radiation pattern characteristics ($L_1 = 31.5$, $g_o = 0.5$, $S = 1$, $L_{out} = 7$, $w_1 = 1$, $a = 8$, $a' = 2$, $L = 7.5$, $W = 2.2$, $L_{sub} = 24$, $W_{sub} = 17$, $g_1 = 0.2$: all dimensions are in mm)

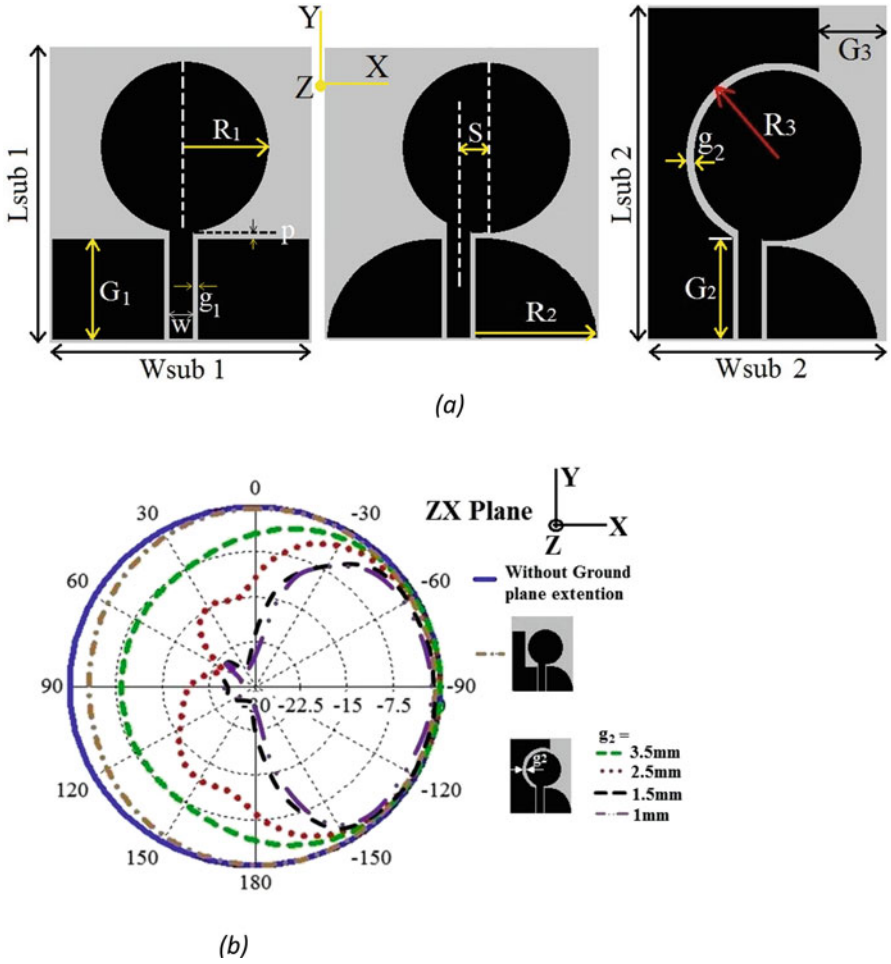

Fig. 4 (**a**) Sequential modification in the ground plane of a printed monopole antenna, (**b**) Variation in the radiation characteristics due to ground plane modification (antenna is designed on FR4 substrate with εr = 4.4, tanδ = 0.025 and 1.6 mm thickness, $R_1 = 11$, $p = 1.5$, $g_1 = 0.5$, $w = 2$, $G_1 = 10$, $S = 4$, $Lsub_1 = 35$, $Wsub_1 = 30$, $R_2 = 13.5$, $R_3 = 12.5$, $g_2 = 1.5$, $G_2 = 12$, $G_3 = 7$, $Lsub_2 = 44$, $Wsub_2 = 28$: all dimensions are in mm)

element due to its extended length, and it also contributes in the generation of multiple closely spaced modes. Since modes are closely spaced, the impedance and the radiation characteristics are consistent for complete impedance bandwidth (Reddy et al. 2015). The surface area of the modified configuration (stage C) is 44 × 28 mm^2 with 1.5 GHz as its lowest resonant frequency. Due to its directional radiation characteristics this antenna is suitable for Ultrawide-band communication and diversity applications. For antenna configurations presented above it can be observed that printed monopole antennas provide wide impedance bandwidth with consistent radiation characteristics.

Broad-band Vivaldi Antenna Configurations

Vivaldi antennas generally comes within the category of directional ultra-wideband antennas. These antennas are travelling-wave type (end-fire) broadband antennas. Vivaldi antennas provide multiple advantages such as wide impedance bandwidth, directional radiation characteristics, and low- profile/compact. The Vivaldi antenna act as resonant configuration at low frequencies, whereas for higher frequencies it behaves as a traveling wave radiator. The inner edges of the flared exponential curve provide guided path for the travelling wave, the reason for wide operating bandwidth is the travelling wave currents which have phase difference of 180°.

Most of the Vivaldi antenna configuration are printed on PCB and fed using microstrip line feed. To improve impedance bandwidth some of the Vivaldi configurations also uses 'balun'. Figure 5 represents a basic configuration of a Vivaldi antenna. Here, λs ('s' represents slot) and λm ('m' represent microstrip line) are the wavelength at center frequency of the required bandwidth. The exponential tapering as shown in Fig. 5 is defined as: (Balanis 2005):

$$y(x) = \pm Ae^{px} \tag{8}$$

where, y corresponds to the half area (or half separation) of the tapered slot. 'A' represent the half of the starting (or opening) width Wmin. 'p' is the rate at which tapering increases along 'x' direction. As discussed in (Balanis 2005) the taper rate 'p' plays a major role in deciding bandwidth and beamwidth of the antenna. An increased 'p' along with 'L' results in a wider bandwidth and beamwidth (in E-plane). However, this also increase the reactive nature of the impedance. It is generally observed that the value of 'L' is typically more than half-wavelength corresponding to lower frequency. Whereas the widths i.e., Wmin and Wmax corresponds to the highest and lowest resonant frequencies, respectively. The bandwidth of the Vivaldi antenna can be tuned using all the dimensional parameter. The major advantage of

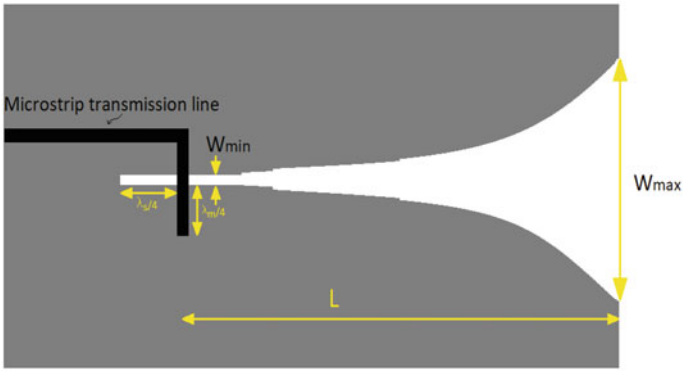

Fig. 5 A basic microstrip line feed Vivaldi antenna

Vivaldi antennas is its integrability with microwave integrated circuits, low profile, broad-bandwidth, and directional radiation characteristics. However, compare to printed monopoles the surface area covered by Vivaldi antenna is still more as the bandwidth is directly proportional to the flaring length.

Metamaterial Inspired Compact Antennas

Metamaterial as we know belong to class of artificial material where permittivity and permeability together contribute towards negative refractive index (Caloz and Itho 2005; Cui et al. 2010). The applications of metamaterial go beyond the limits of left-handed (LH) or composite left-handed-Right handed (CLHRH) mediums. The concept of metamaterial transmission lines (Caloz and Itho 2005) are now extended to metamaterial inspired antennas which brings in a class of compact antennas. In (Erentok and Ziolkowski 2008) multiple examples of compact metamaterial antennas are reported. Most of these antennas are designed by integrating lumped elements or equivalent of lumped elements to device a radiative resonator. One can achieve resonant frequency by appropriate integration of L and C elements (or its equivalent) within a radiating system. To model a metamaterial inspired resonator most conventional technique is to integrate a capacitive loaded loop (CLL) and an inductive element (either a metallic wide or its printed equivalent) (Dash et al. 2020). As reported in (Dash et al. 2020) the designed ESA consists of multilayered hemicylindrical shaped structure which provide a composite LC resonator and an multiturn solenoidal inductive loop which is a primary radiator for the reported configuration.

Split-Ring Resonator (SRR) Loaded Antenna

The metamaterial inspired antennas can provide a viable solution towards the compact design of millimeter wave antenna. These antennas antennas can be designed using a simple combination of split ring resonator (SRR) and metallic strip as shown in Fig. 6.

In Fig. 6a simple composite configuration of split ring resonators (SRR) separated by a section of Jerusalem cross is presented. This configuration is fed with a differential feed. Here, the SRR structure along with the ‘I’ section provides a resonance at 61.3 GHz. The presented configuration provides an impedance bandwidth of 2 GHz. Here the SRR configuration are capacitively coupled with the ‘L’ shaped feed line resonator. Most of the reported metamaterial inspired configuration suffers from narrow bandwidth. This is due to a limit in the number of modes getting excited by resonators usually used in the radiating elements. To achieve a broad-bandwidth response it is important that the designed antenna configuration exhibits a multimode response where modes are closely spaced as discussed in previous section.

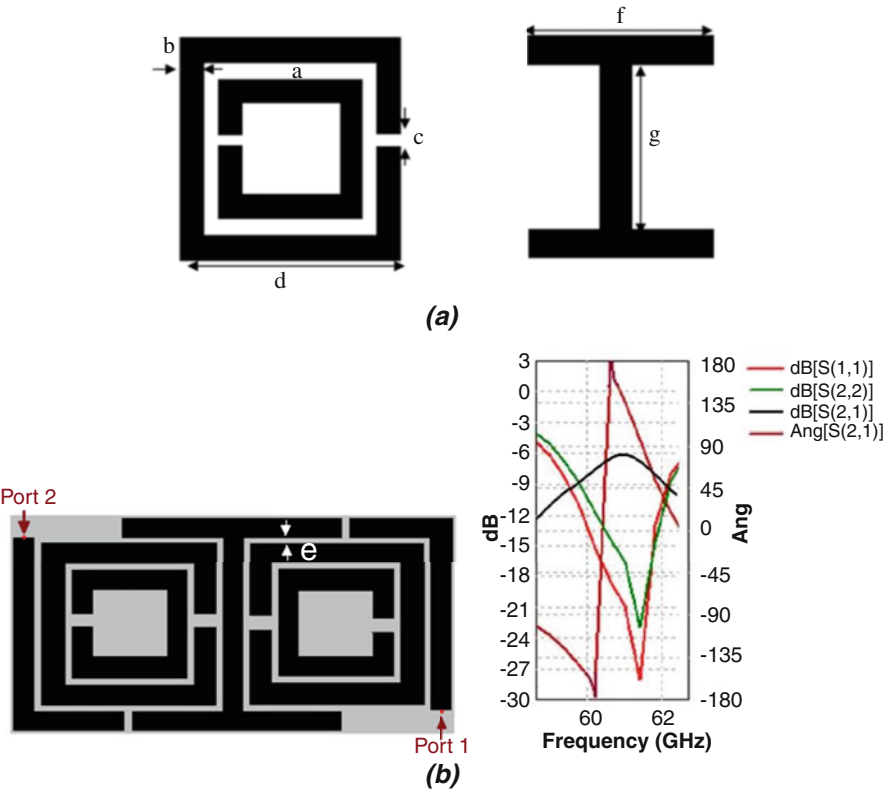

Fig. 6 (**a**) unit cell representation of the antenna, (**b**) composite millimeter wave antenna configuration and its simulated S-parameter (Dimensions: a = 0.025 mm, b = 0.1 mm, c = 0.05 mm, d = 0.9 mm, f = 1.15 mm, g = 1.05 mm, e = 0.025 mm, configuration is simulated using Duriod 5880 substrate with thickness 0.4 mm)

Antenna Loaded with Periodic Patches

The multimode response exhibited by printed monopole antennas can be extended for to design a broadband antenna at millimeter wave frequencies. As presented in Fig. 7, the antenna consists of a partial ground plane loaded with periodic patches. The partial ground enables the wide band response whereas the periodic patches will serve the purpose of surface wave cancellation along the intermediate slots/gaps created between the periodic patches. Here, the periodic patches in the ground plane are placed such that their center-to-center distance is less than the guided wavelength. The presented structure is designed on the FR4 substrate with dielectric permittivity of 4.4. This antenna is designed to cover both 60 GHz band and 70 GHz band (with potential application in automotive radar systems). The structure is designed and simulated using CST Microwave Studio. It can be concluded form above section that the bandwidth of the antenna highly depends on the nature of resonators used to design the antenna.

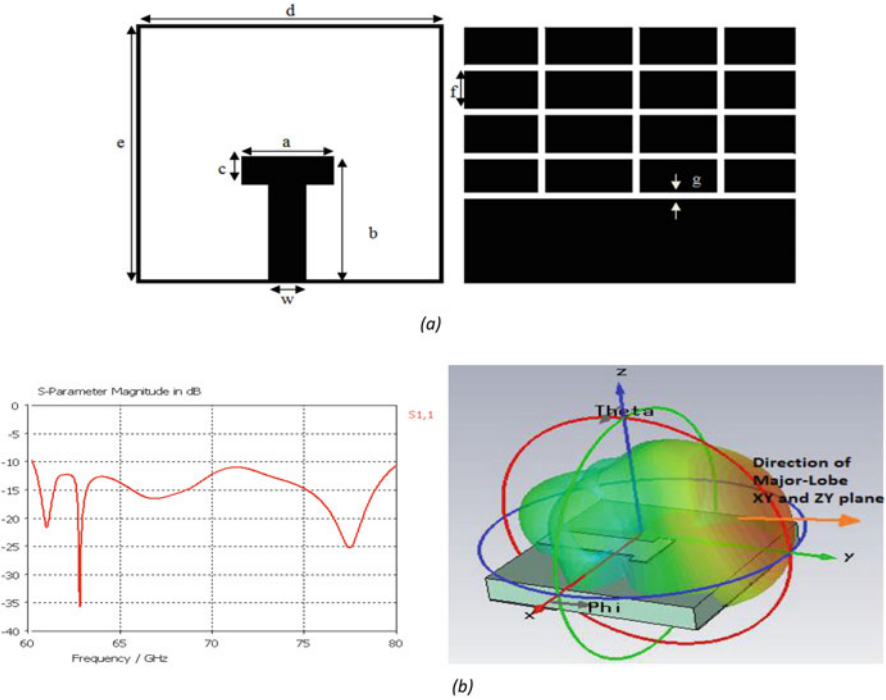

Fig. 7 (**a**) A printed monopole configuration loaded with partial ground and parasitic patches. (**b**) Reflection parameter response and radiation pattern (Dimensions: a = 0.36λg, b = 0.7λg, c = 0.1λg, d = 1.2λg, e = 1.2λg, f = 0.22λg, g = 0.02λg, u = v = 1.2λg, y = 0.6λg, λg = guided wavelength at 61 GHz)

The fabrication and testing of antennas at millimeter waveband are the most challenging task, as dimensions at millimeter wave band becomes critically small. To validate the concept presented in periodic patch loaded antenna, an LS band antenna is designed and reported in (Reddy and Alex 2012). Here the reported configuration utilizes the similar concept of periodic patches loading on partial ground configuration (Fig. 8).

The impedance bandwidth of the proposed antenna configuration is from 1.5 GHz – 2.5 GHz with radiation pattern along Y direction (major lobe). As reported in (Reddy and Alex 2012) this antenna is tested for its impedance and radiation characteristics.

Resonator Based Compact Antenna

In line with metamaterial inspired antennas, there are another set of antennas which are design using composition of resonators such as meander-line resonators, loop resonators and conventional quarter-wavelength resonators (Sharma et al. 2015;

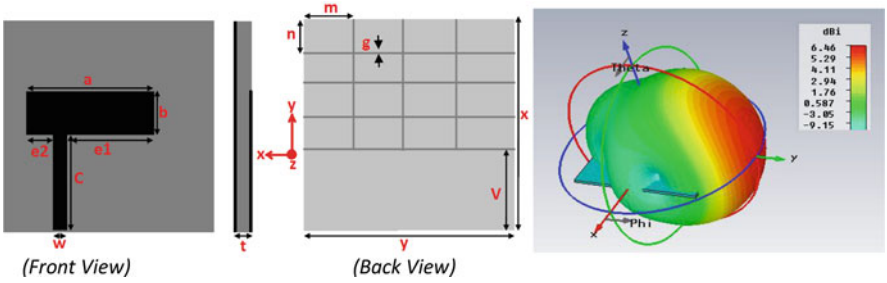

Fig. 8 Periodic patches loaded L-S band antenna with its radiation characteristics, (Dimensions: a = 35 mm, b = 12 mm, c = 34 mm, e1 = 20 mm, e2 = 12 mm, w = 3 mm, t = 1.6 mm, n = 9 mm, m = 25 mm, g = 0.1 mm, v = 27 mm, x = 73 mm, y = 66 mm)

Sharma and Chaudhary 2015). For a simple configuration a composite set of multiple resonators can provide a viable solution towards compact antenna design. Meander line resonators are one of the most basic configurations used to design resonant circuit (Hong and Lancaster 2011). The successive turns in the meander line configuration provides equivalent L and C configuration suitable to constitute a resonant circuit. The resonant frequency for a meander-line resonator can be tuned easily by optimizing the number of turns or the dimensions (length and gaps) of the turns. One such example towards compact meander-line antenna is reported in (Padhi et al. 2020). This reported configuration employs a meander line resonator with a triangular loop to produce a dual band operation as shown in Fig. 9.

The antenna configuration presented in Fig. 9a belongs to a class of electrically small antennas (ESA), which satisfies the ESA criteria laid down in (Wheeler 1975; Chu 1948). The reported antenna presents a dual band operation by using the resonant behavior of a quarter wavelength resonator and meander loop resonator. Over-all surface area of this antenna is 0.24 $\lambda_o \times 0.09\ \lambda_o \times 0.009\ \lambda_o$ which is quite compact when compared to the wavelength of the lowest resonant frequency. This configuration proves that a compact antenna with increased impedance bandwidth can be devised by proper composition of multiple resonators. However, compare to the printed monopoles ESA offered less bandwidth. ESAs can be modeled to cater the specific device requirements (Chen et al. 2020) and hence its more popular among communication device industry.

Conclusion

In this chapter several broadband and compact antenna design configurations are discussed. Most of these antennas are integrability to a system/circuit. Antennas discussed in this chapter ranges from compact broadband printed monopoles to Vivaldi antennas and metamaterial inspired antennas to electrically small antennas. Some of the configuration possesses partial ground or resonant ground which further adds to the pattern and impedance bandwidth. In some configuration multiple

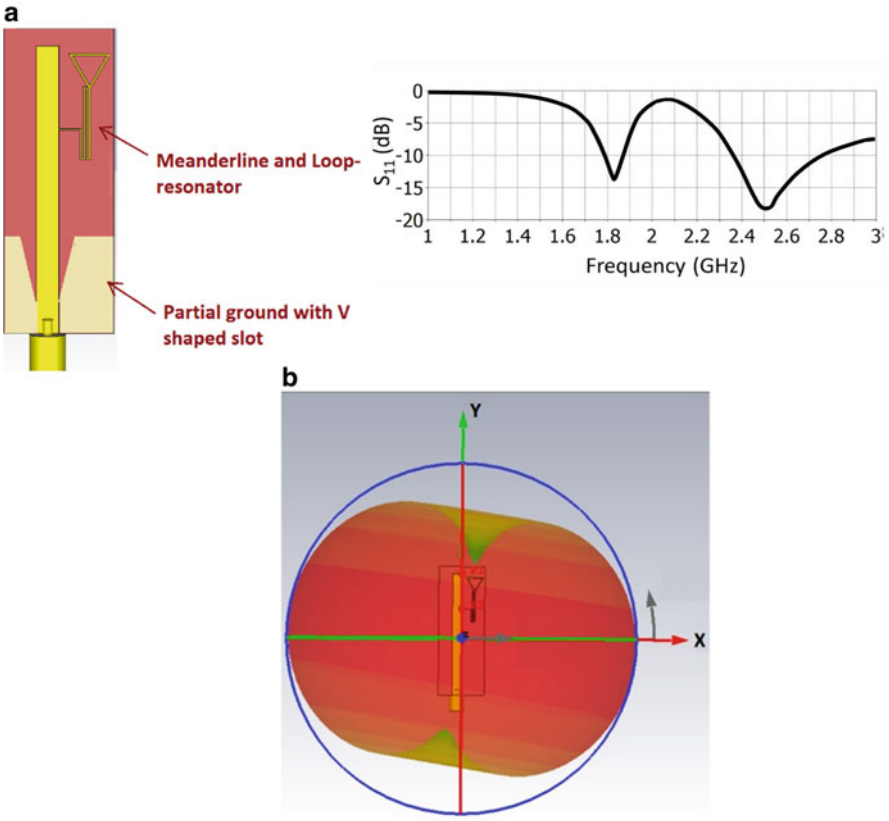

Fig. 9 (**a**) Monopole antenna loaded with meander line and loop resonators and its S parameter response (**b**) Radiation pattern characteristics at 1800 MHz

resonators are added to have either multiple pass-band response or to have wide transition in impedance bandwidth response. The metamaterial inspired antennas discussed in this chapter can possesses both broadband and narrow band response depending on the nature of primary resonator or design methodology. An observation can be drawn from the presented design techniques that the impedance bandwidth and radiation bandwidth can be increased by increasing the modes in a configuration and ensuring that those modes are closely placed.

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Beam Steering of Antenna Array Using Phase Gradient Metasurface

13

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Abstract

Metasurfaces are thin engineered surfaces consisting of either metallic or dielectric resonators arranged in a two-dimensional array with sub-wavelength grid sizes. In recent years, metasurfaces have attracted significant attention from researchers and scientists attributed to their extraordinary capabilities of controlling the propagation and generation of electromagnetic wave. Across the electromagnetic (EM) spectrum from microwave to visible frequency, metasurfaces have been demonstrated to generate a beam of the desired shape, deflect the radiation to the particular direction, spectral and temporal manipulation of the beam with unprecedented control. These features pave the way toward diverse practical applications in the field of sensing, medical imaging, and quantum computing. Furthermore, with the advent of 5G technology, metasurfaces are deployed in wireless communication systems. Beam forming and steering that beam at desired direction is one essential criterion of the 5G communication system. By adjusting the phase of the scattered wave at each of the unit cells of the metasurface, the overall scattered beam can be directed toward any desired direction. These metasurfaces are categorized to a special class of metasurface referred to as "phase gradient metasurfaces." In this chapter, we will discuss in detail the phase gradient metasurface for the beam steering application.

Phase Gradient Metasurface and Theoretical Background

Although phase gradient metasurfaces emerge as an essential ingredient of the cutting edge 5G wireless communication technology, the background theory can be traced back to the sixteenth century, called the Huygens' principle named after scientist Christiaan Huygens. This mechanism is illustrated in Fig. 1a. According to this Huygens' principle, while electromagnetic wave traversing from one medium to another medium, each point on the interface plane acts as a secondary source of the wave. Hence, each of these secondary wavelets emit wave of the spherical wavefront to the forward direction. The new wavefront emerges as the tangential surface of all these secondary wavefronts. This emerging wavefront can take the shape of any predetermined waveform if the phase values of the secondary waves are manipulated accordingly. An example of this wavefront shaping is depicted in Fig. 1b. The example shows that, if the interface between two media is engineered in such a way that the secondary wavefronts of the middle portion of the interface plane are delayed more than those of the edges, the emerging wavefronts bend and get focused on a point. This behavior of the emerging wavefronts is exactly the same as those wavefronts refracted through a convex lens as shown in Fig. 1c.

The more rigorous mathematical foundation of the Huygens' principle was reported in the seminal paper named "Light Propagation with Phase Discontinuities: Generalized Laws of Reflection and Refraction" authored by Nanfang Yu et al. The authors show that Snell's law of reflection and refraction can be generalized using Fermat's principle (Yu et al. 2011). Here, the term "generalized" refers to the fact that

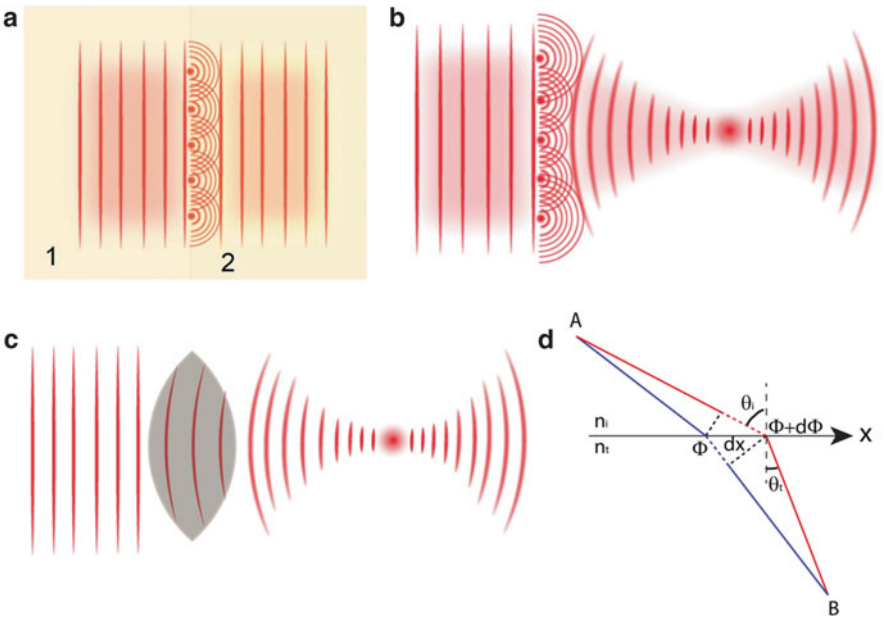

Fig. 1 (**a**) Illustration of Huygens' principle using a wavefront traveling from medium 1 (denoted by 1) to medium 2 (denoted by 2), (**b**) convergence of the wavefront traveling through an engineered interface between two media, (**c**) convergence of the wavefront refracted through a convex lens, (**d**) illustration of the Fermat's principle (Yu et al. 2011)

reflected or refracted light ray can be directed to a particular direction that is determined by Fermat's principle. Fermat's principle states that the light ray takes the least time to traverse from one point to another point. Alternately, the Fermat's principle can be stated that the derivative of the accumulated phase along the actual optical path ($\int_a^b d\varphi\left(\vec{r}_s\right)$; light ray traversing from point a to point b) with respect to the infinitesimal variations of the path will be zero.

To derive the generalized Snell's law, we illustrate here the Fermet's principle in Fig. 1d. Let us consider that a light ray is traveling from point A in the medium 1 to point B in the medium 2. Now, in the path of light propagation, we can introduce an abrupt shift of phase value $\Phi\left(\vec{r}_s\right)$ at the interface plane between the two media; here, $\Phi\left(\vec{r}_s\right)$ is the value of the phase shift at a point whose position vector is $\vec{r}_s$, on the interface plane. Hence, the total phase shift in the light traversing the path from point A to point B would be $\Phi\left(\vec{r}_s\right) + \int_a^b \vec{k}.d\vec{r}$; here, $\vec{k}$ is the wave vector of light propagation and equal to $2\pi/\lambda$, where λ is the wavelength. This total phase value would be the stationary phase for the actual light path. In other words, if between point A and point B, there are two light propagation paths those are infinitesimal closed, the difference between accumulated phase values along the respective paths would be zero. Mathematically, this statement can be expressed by the following equation,

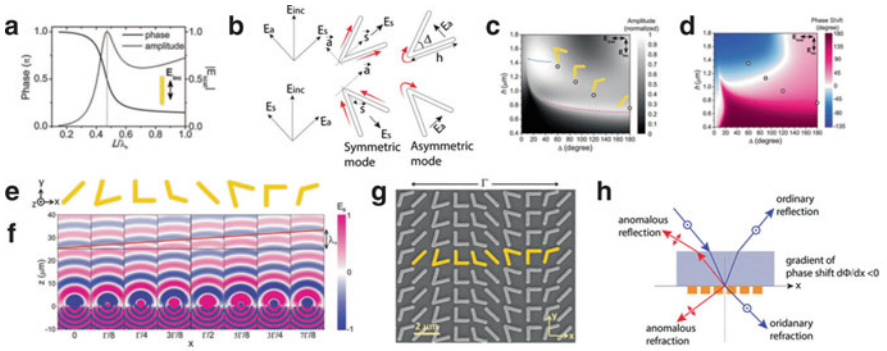

Fig. 2 (**a**) Amplitude and phase spectra of a dipole resonator, (**b**) symmetric and anti-symmetric resonance modes of the V-shaped resonator, (**c**) two-dimensional plot of the amplitude of the scattered radiation with respect to h and Δ, (**d**) two-dimensional plot of the phase of the scattered radiation with respect to h and Δ, (**e**) set of the eight unit cell resonators, (**f**) simulation results of electric field scattered from eight unit cells, (**g**) the Huygens' metasurface, (**h**) illustration of schematic of Huygens' metasurface with ray-optics (Yu et al. 2011)

$$[kn_i \sin(\theta_i)dx + (\Phi + d\Phi)] - [kn_t \sin(\theta_t)dx + \Phi] = 0 \tag{1}$$

In this equation, θ_i and θ_t are the incidence and refraction angles, respectively. The $d\Phi$ is the phase difference introduced between two infinitesimal closed points, separated by the distance dx , as depicted in Fig. 2; n_i and n_t are the refractive indices of two media, respectively. If we consider the phase gradient $d\Phi/dx$ of the media interface as constant, the Eq. (1) will lead to the following equation, which is the generalized Snell's law for light refraction.

$$n_t \sin(\theta_t) - n_i \sin(\theta_i) = \frac{\lambda_0}{2\pi} \frac{d\Phi}{dx} \tag{2}$$

In a similar way, the generalized Snell's law of reflection is expressed as follows:

$$\sin(\theta_r) - \sin(\theta_i) = \frac{\lambda_0}{2\pi} \frac{d\Phi}{dx} \tag{3}$$

In this equation θ_r is the angle of reflection. Please note that, in Eqs. (2) and (3), if constant phase gradient $d\Phi/dx$ is set to zero, the generalized Snell's law will be reduced to the conventional Snell's law that we studied in the elementary physics book. However, with the introduction of a suitable constant gradient of phase discontinuity along the interface between two media, the refracted and reflected beam can be deflected to any desired direction. This is evident from the above two equations that both refraction angle θ_t and reflection angle θ_r depend on the phase gradient. Hence, reflected and refracted beam can be steered at desired directions by creating a suitable phase profile at the interface. Now the quest is how we will obtain this phase gradient $d\Phi/dx$ along the interface between two media. In practice, the

process of obtaining this continuous phase gradient is discretized and these discrete values of phase shift are obtained by designing metal or dielectric resonators as those resonators can be engineered to possess suitable amplitude and phase of the scattered radiation. Depending on these metal or dielectric resonators, phase gradient metasurfaces can be of two types – (a) metal or plasmonic metasurface and (b) all-dielectric metasurfaces. In the following section, we describe the design methodology of beam steering antenna array with phase gradient metasurface in detail.

Design Methodology

As hinted in the previous section, the electromagnetic resonator made of either metal or dielectric is the fundamental constituent of the phase gradient metasurface. Two possible phenomena named absorption and scattering occur, while incident radiation impinges on the electromagnetic resonator. Here, we will particularly focus on the scattering phenomenon as we are talking about the steering of the scattered radiation from the phase gradient metasurface. To describe the methodology, we consider here two types of beam steering antenna array as illustrative examples. The first one is called Huygens' metasurface and the second one is called microstrip or printed reflectarray antenna. Most of the beam steering antennas are either of these two types.

Huygens' Metasurface

As mentioned in the previous section, the phase gradient is achieved by suitably placing different sized resonators possessing proper phase shifts. In order to get complete control over the wavefront, the phase shift is required to cover the full range of 0-to-360°. Moreover, the amplitude of the scattered radiation needs to be constant across the resonator sizes. The phase as well as amplitude of the scattered radiation vary with the frequency. The variation of phase becomes appreciable at the vicinity of the resonance frequency. For example, a dipole resonator resonates at a frequency when dipole length is equal to half of the corresponding wavelength of that frequency ($L = \lambda_0/2$). This is called $\lambda_0/2$ resonance. At this resonance frequency, slope of the phase becomes the maximum and scattered radiation attains the maximum amplitude. As the resonance of dipole depends on its length, the resonance shifts toward lower and higher frequencies in accordance with the increase and decrease of the length, respectively. Hence, phase as well as amplitude, both vary with the length of the dipole. Now, if we plot these varied phase and amplitude values with respect to the varied lengths at one particular frequency, we will obtain the graphs as shown in Fig. 2a. However, as we can clearly observe that the varied length of the dipole resonator cannot cover the full phase range of 0-to-360°. Moreover, the scattering amplitude is not constant over all the lengths. This issue can be resolved by introducing resonators having multiple (double or triple) resonance features. The first demonstrated Huygens' metasurface consisted of an array of

V-shaped resonators those exhibiting such double resonance properties (Yu et al. 2011). The V-shaped resonator has two arms of the same length h and they are connected with each other at an intermediate angle Δ, as depicted in Fig. 2b. Two different resonance modes are excited in the V-shaped resonator while illuminated with electromagnetic radiation. These modes are named as symmetric and anti-symmetric mode, respectively. In case of the symmetric mode, each arm behaves like an individual dipole resonator, hence, the $\lambda_0/2$ resonance is observed in each arm and $h = \lambda_0/2$. In case of the anti-symmetric mode, the current distribution along two arms occurs in such a way that the whole V-shaped resonator acts as a single dipole; hence, in this case $2h = \lambda_0/2$. As illustrated in Fig. 2b, both resonance modes can be excited for an arbitrary polarized radiation. More precisely, while the incident wave is polarized at an angle of 45° with respect to both $\vec{a}$ and $\vec{s}$ vectors, both modes are excited, and the field component polarized perpendicular to that of the incident radiation is present in the scattered radiation in a substantial amount. This cross-polarized scattered wave can possess a complete 360° phase range as well as large scattering amplitude due to the modal properties of V-shaped resonators and two degrees of freedom (h and Δ) in the resonator geometry. Figure 2c, d present the two-dimensional plots of phase and amplitude distribution, respectively, with respect to the varying arm lengths (h) and intermediate angles (Δ). From these plots, four resonators were chosen and shown with four respective black circles as marked on the amplitude and phase distribution. As shown in Fig. 2d, these resonators from left to right provide an incremental phase shift of 45° for the cross-polarized scattered radiation. Moreover, these selected resonators have large scattering amplitudes. Now, by taking mirror structures of these four V-shaped resonators, four new resonators can be created. Each of these newly created resonators can generate cross-polarized scattered radiation with 180° out of phase with respect to that of its corresponding mirror structure. This is due to the current flowing in two mirror-shaped resonators are of opposite polarity with respect to each other. Hence, a set of total eight resonators are obtained from the process of unit cell design. The amplitude of the cross polarized scattered radiation from each of eight resonators is high and they cover total a 360° phase with a step of 45°. Now, we will discuss how these eight resonators are used to build a Huygens' metasurface. Each resonator is called the unit cell of the metasurface and often the unit cells of metasurface are also called meta-atoms. Figure 2e illustrates that arranging these eight resonators in a row along the x-direction creates a discretized version of phase gradient $d\Phi/dx$ that is equal to 360°/Γ, where Γ is the total length of the eight unit cells along the x-axis. This set of eight unit cells arranged along a row is called super-cell. According to the generalized Snell's law described in Eq. (2), the scattered beam would be deflected at an angle of θ_r depending on the value of θ_i and $|d\Phi/dx| = 360°/\Gamma$. In this particular example, θ_i is considered as 0°, as incident radiation impinges on the metasurface normally. The scattered radiation from successive unit cells is delayed from right toward left as there is 45° phase difference between scattered radiation from two successive unit cells. As showing in Fig. 2f, the red line indicates that the scattered beam is tilted by an angle. By changing $|d\Phi/dx|$, the scattered beam can be steered at any desired direction. The whole metasurface shown in Fig. 2g is consisted of the

array of the super-cells and each super-cell contains this set of eight unit cells. As a result, the collective scattering from all these super-cells is directed to the particular direction θ_r. Figure 2h illustrates the schematic of beam deflection using Huygens' metasurface. The rays drawn in red lines represent the anomalous reflection and refraction resulting from the gradient of phase shift along the x-axis.

We can generalize this design methodology of Huygens' metasurface in the successive steps described below:

1. The metasurface consists of the array of the super-cells.
2. Each super-cell contains N number of unit cells. Let us consider, these N unit cells are arranged along the x direction.
3. Let us consider, the unit cell dimensions are P_x and P_y in two transverse directions x and y, respectively.
4. Hence, the super-cell dimensions would be NP_x and P_y in the x and y, respectively.
5. The phase difference between two successive unit cells would be $360°/N$.
6. The phase gradient in a super-cell would be $360°/NP_x$.

Therefore, we need to come up with N different unit cells each of which possesses distinct phase value ranging from 0-to-360° with a step between two successive unit cell-phases $360°/N$. While designing the unit cell, we also need to take care of the fact that the scattering amplitude from each unit cell must be almost constant. The constructed Huygens' metasurface can deflect the incident radiation to any desired direction that leads to the beam steering metasurface. In the next section, we will discuss the design methodology of the microstrip reflectarray antenna.

Microstrip Reflectarray

The working mechanism of the microstrip reflectarray relies on the same Huygens' principle. However, reflectarray antenna and microstrip reflectarray has been demonstrated in radio frequency since 1950. Microstrip reflectarray consists of an array of variable-sized microstrip patch resonators on a grounded dielectric substrate. The whole array is illuminated with a feeding horn antenna. A representative image of microstrip reflectarray is given in the following Fig. 3a. A schematic image of the microstrip reflectarray is shown in Fig. 3b. As illustrated in Fig. 3b, the array elements are illuminated with the radiation from a feeding horn. It is assumed that the antenna elements are in the far-field zone of the feeding horn. Let us consider the reflectarray as a transmitting antenna without losing the generality. Hence, the feeding horn antenna is connected to a RF transmitter. In this situation, electromagnetic radiation from the horn antenna impinges on each individual element at different incident angles. From the horn antenna to reflectarray elements, the wave propagates spherically as illustrated in Fig. 3b. The incident field on each individual element can be considered as a local plane wave with a phase proportional to the spherical wave propagation path from the phase centre of the feeding horn to that antenna element. Now, in order to convert to this spherical wave to the plane wavefront, the field must be reflected from

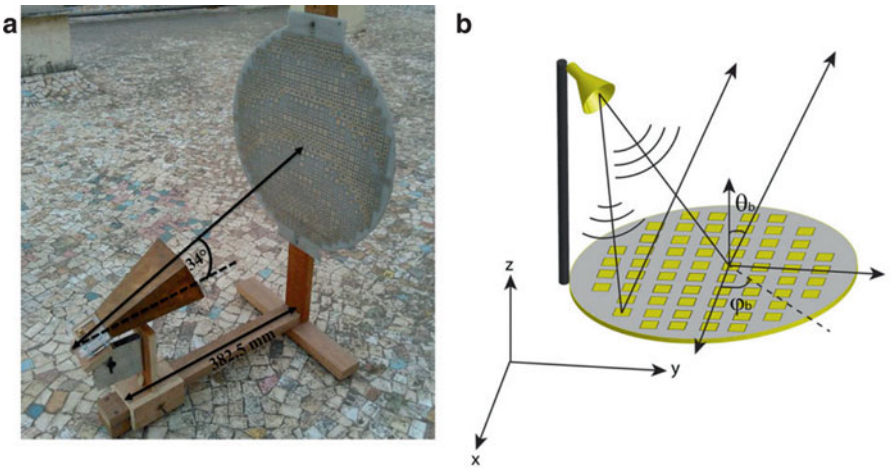

Fig. 3 (**a**) A reflectarray antenna with offset feed horn antenna (Pattanayak et al. 2018), (**b**) schematic of the reflectarray with feeding horn

each element of the array with a suitable phase shift, so that all the reflected beams from corresponding elements go parallel to each other and interfere constructively to form a highly focused beam in the far field region. Hence, the main point here is that each individual element should possess an appropriate reflection phase and that reflection phase needs to be adjusted independently to form a progressive phase distribution of reflected wave on the planer surface of reflectarray. As a result, the reflected wave from the reflectarray surface forms a narrow beam that can be steered at any desired direction by suitably adjusting the reflection phase profile.

The phase of the reflected field at each unit cell of the reflectarray antenna is equal to the sum of two phase components (Huang and Encinar n.d.). These are (1) phase accumulated due to propagation of incident field from feed antenna to unit cell and (2) phase-shift introduced by each unit cell to generate pencil beam at the desired direction. Hence, if we consider the reflectarray is lying on the XY-plane of the coordinate system, the resulted phase shift $\phi(x_i, y_i)$ at the i-th unit cell centered at (x_i, y_i) is expressed as:

$$\phi(x_i, y_i) = -k_0 d_i + \phi_R(x_i, y_i), \tag{4}$$

where $k_0 = 2\pi/\lambda$ is the free space propagation constant, d_i is the distance from feed horn to the i-th element of the array, and $\phi_R(x_i, y_i)$ is the phase of reflection coefficient for the i-th element centered at (x_i, y_i). On the other hand, to deflect a far field radiated beam at (θ_b, φ_b) direction, the progressive phase distribution of the reflectarray surface needs to satisfy the following equation that originated from the antenna array theory,

$$\phi(x_i, y_i) = -k_0 x_i \ \sin\theta_b \cos\varphi_b - k_0 y_i \ \sin\theta_b \sin\varphi_b \tag{5}$$

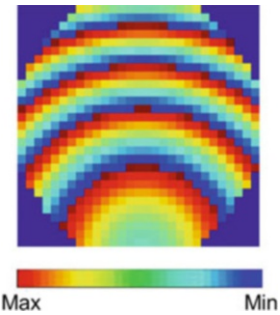

Fig. 4 Phase distribution of a reflectarray (Pattanayak et al. 2018)

By equating the above two expressions, we can derive $\phi_R(x_i, y_i)$ as follows:

$$\phi_R(x_i, y_i) = k_0(d_i - (x_i \cos\varphi_b + y_i \sin\varphi_b)\sin\theta_b) \tag{6}$$

$$d_i = \sqrt{(x_i - x_f)^2 + (y_i - y_f)^2 + (z_i - z_f)^2} \tag{7}$$

where (x_f, y_f, z_f) is the coordinate of the phase center of the feeding horn.

The above equation calculates the required phase shift at each element of the array while the respective values of d_i, (θ_b, φ_b) and (x_i, y_i, z_i) are provided. Figure 4 illustrates a phase distribution of a circular reflectarray which was aimed to design to deflect the main beam at ($\theta_b = 0°$, $\varphi_b = 0°$) with an offset feed horn antenna as the reflectarray shown in Fig. 3a.

After calculating the phase distribution, the next step is to design the unit cell to obtain the reflection phase values ϕ_R required according to the equation. Unit cell design is similar to the case of Huygens' metasurface that has been described in the previous section. The only difference is that the backside reflectarray antenna is covered with a continuous metallic ground plane. The reflected wave amplitude is always high due to this backside reflector. Hence, while designing the unit cell of a reflectarray antenna, we need not be worried about the amplitude of the reflected wave; we only concentrate on the reflection phase graph. The required reflection phase can be any value in the range from 0° to 360°. This phase range is obtained by varying one or multiple geometrical parameters of the unit cell. These unit cells are actually metallic or dielectric resonators on the grounded dielectric substrate. An illustrative example is given here to discuss the unit cell simulation in order to obtain the reflection phase graph. Here, a square patch is considered for this example to obtain the required phase graph for a reflectarray antenna. At the resonance frequency, reflected wave amplitude becomes minimum, and phase crosses at zero value, as shown in Fig. 5a. As the resonance of microstrip patch depends on its length, the resonance shifts toward lower and higher frequency in accordance with the increase and decrease of the length, respectively. As a result of that, for different values of the length, we get different phase spectra. As shown in Fig. 5b, two different phase spectra result from two different lengths, respectively. Now, if we

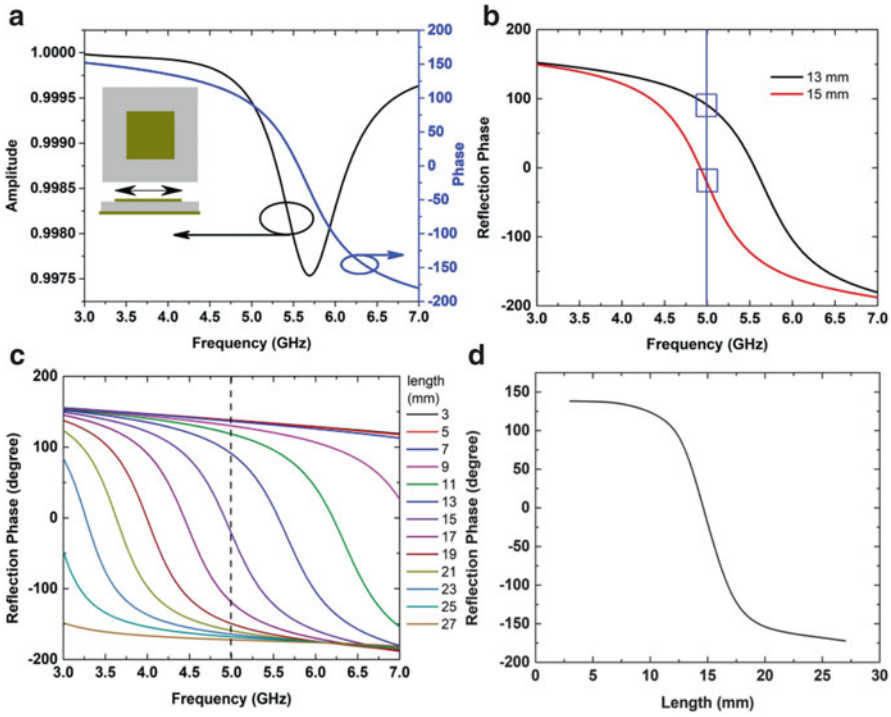

Fig. 5 (**a**) Reflection amplitude and phase vs. frequency of a square unit cell, (**b**) reflection phase graphs vs. frequency for two different lengths of the patch, (**c**) reflection phase graphs vs. frequency for a range of patch lengths, (**d**) reflection phase vs. patch length at operating frequency 5 GHz

concentrate on a particular frequency of our interest, we will see the phase values changes with varying lengths. This particular frequency is marked with a blue vertical line in Fig. 5b. These two different phase values are marked with respective blue squares. The reflection phase spectra for a range of different length values are shown in Fig. 5c. Now, if we plot these varied phase values with respect to the varied lengths, we will obtain the graphs plotted in Fig. 5d.

Once the unit cell design is ready, the next step is to map those unit cells properly according to the phase distribution of the reflectarray. This mapping of unit cells is illustrated in the following Fig. 6. The reflection phase profile obtained from Eq. (7), provides the information about the required phase shift at any particular position on the reflectarray. The total area of reflectarray is discretized with a number of grids. The length and width of each grid are the same of those of the unit cell.

Corresponding unit cell dimension for that required phase value is obtained from the reflection phase graph. Once, the unit cell dimension is known for a particular position, that unit cell is placed at that position. In this way, the unit cells of suitable sizes are properly placed on the corresponding grids of the reflectarray.

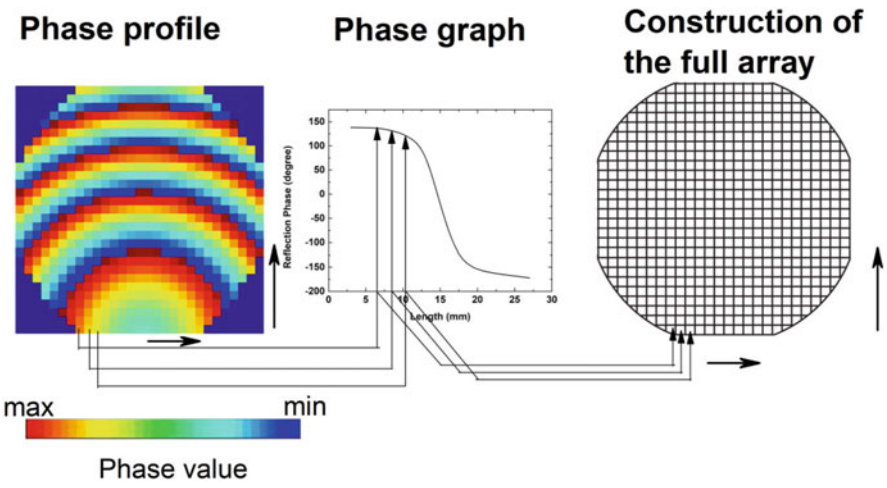

Fig. 6 Mapping of the unit cells according to the phase distribution. Phase profile determines the required phase shifts at different co-ordinates on the array. According to the required phase shift at a position, corresponding unit cell are placed on the array. This correspondence between unit cell dimension and required phase at a position is governed by the reflection phase graph

The first main criterion to design the unit cell is to cover at least the full phase range of 0-to-360°. Hence, phase range is an important parameter for the unit cell design (Bozzi et al. 2003). The phase range is expressed mathematically as follows:

$$\Delta = \phi_{Rmax} - \phi_{Rmin}, \tag{8}$$

where, ϕ_{Rmax} and ϕ_{Rmin} are the maximum and minimum reflection phase values, respectively, obtained at the operating frequency of the reflectarray. In case, the unit cell is not able to cover the full phase range, some of the required reflection phases may fall outside of the phase range Δ. This will lead to the phase error that in turn will affect the performance of the reflectarray. Therefore, the designed unit cell should cover at least 360° phase range. The second important criterion for unit cell design is phase sensitivity (Bozzi et al. 2003), which is expressed as follows:

$$\sigma = max \left| \frac{\partial \phi_R}{\partial A} \right|, \tag{9}$$

where A is the varying geometrical parameter. The high value of σ indicates the sharp variation of the reflection phase graph with respect to the varying parameter A. In that scenario, a small change in the value of A due to the fabrication error will lead to a high variation of the reflection phase. For example, if the maximum phase sensitivity of a phasing element is 1000°/mm, it can cause a phase error of ±50° with 0.1 mm etching tolerance (Yoon et al. 2015a). Hence, higher phase sensitivity compared to etching tolerance causes phase error. This will degrade the performance of the reflectarray. Hence, the value of σ should be as small as possible.

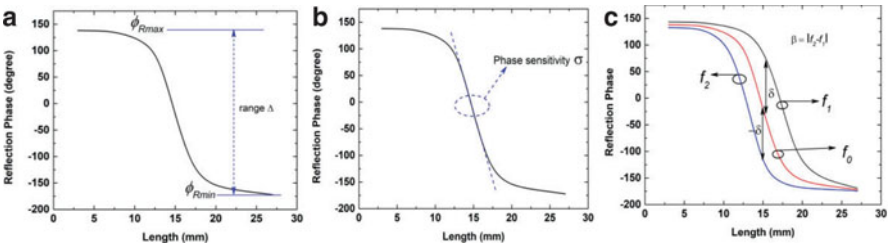

Fig. 7 (**a**) Phase range, (**b**) phase sensitivity and (**c**) bandwidth of a reflection phase graph

Apart from these two main criteria, another important factor of unit cell design is bandwidth (Bozzi et al. 2003). The reflection phase graph is determined at a particular frequency as illustrated in the previous Fig. 5d. Bandwidth is determined from the fact that how much variation of the reflection phase graph occurs when the operating frequency is changed. For broadband operation, it is desirable to have minimal change in the reflection phase graph for change in operating frequency. Mathematically, bandwidth is expressed as follows:

$$\beta = | f_2 - f_1 | \tag{10}$$

where, f_1 and f_2 are the two different frequencies, at which the reflection phase differs from the phase graph at operating frequency f_0 by a certain limited amount, say δ°.

The phase range Δ, phase sensitivity σ, and bandwidth β of a reflection phase graph are illustrated in Fig. 7a, b, c, respectively.

All these design criteria depend not only on the shape of the resonator but also the resonator size, permittivity and thickness of the dielectric layer as well

System Level Description of Beam Steering Antenna

As explained so far, the beam steering antenna is comprised of individual phasing elements arranged in a two-dimensional array in such a way so that scattered radiations from all these elements constructively interfere to generate a pencil beam toward the desired direction. A schematic diagram of the beam steering antenna array is depicted in Fig. 8. The radiated beam of desired features is the main design criteria to realize such an antenna. In this section, we are describing the system level performances of the beam steering antennas.

Radiation Pattern

The radiation pattern is a quantitative description of how radiated power from the antenna is distributed at different angular directions in three-dimensional space (Visser n.d.). As depicted in Fig. 8, the angular region consisting of the strongest

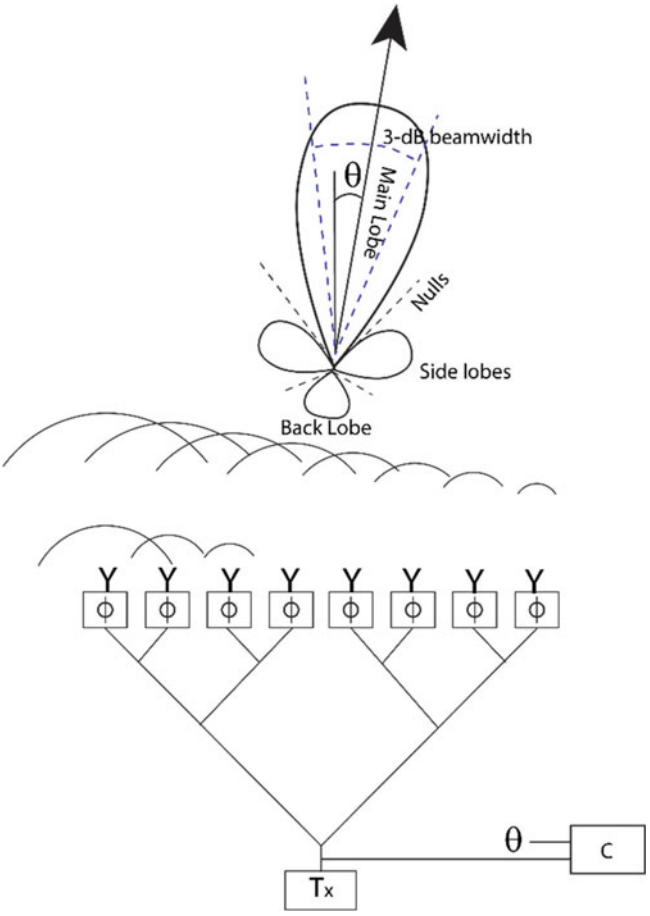

Fig. 8 System level description of the beam steering antenna. Tx and c represent the transmitter and controlling unit, respectively

radiation is called the main lobe or main beam. Apart from the main lobe, other angular regions of locally strong radiation are called the side lobes. The side lobe which is at 180° angular position with respect to the main lobe is called the back lobe. At some specific directions, no power is radiated at all. These are called nulls.

Half-Power Beam Width

HPBW is defined as the angular distance between the two points on a cut-plane of the main beam, at which radiated power is half of the maximum radiated power. As, $10\log_{10}0.5 \approx 3$ dB, this is also called as 3-dB beam width.

Directivity and Gain of Antenna

Very often we need to compare performance-wise one antenna with another antenna. In this comparison process, an isotropic radiator is taken as the reference. As the name suggested, an isotropic radiator radiates equally in all the directions in the three-dimensional space. However, in practice, an isotropic radiator is extremely hard to realize. All practical antennas radiate more power to some preferred directions than other directions. This feature of radiating power in some preferred directions is called directivity. Directivity function $D(\vartheta, \varphi)$ at some specific direction (ϑ, φ), is defined as the ratio of power radiated in the direction of (ϑ, φ) by the antenna and the power radiated in that same direction by an isotropic radiator. $D(\vartheta, \varphi)$ is expressed as below:

$$D(\vartheta, \varphi) = \frac{P(\vartheta, \varphi)}{P_t/4\pi} \tag{11}$$

where, $P(\vartheta, \varphi)$ is the power radiated by the antenna in direction (ϑ, φ) per solid angle and P_t is the total radiated power. If this total radiated power P_t is radiated by an isotropic radiator, the radiated power in any direction is $P_t/4\pi$. Hence, directivity is a comparative measurement that quantifies how much of total radiated power is confined into the angular region of the main lobe (Visser n.d.).

The directivity D is defined as the maximum value of the directivity function. Hence, D is expressed as follows:

$$D = max\,[D(\vartheta, \varphi)] \tag{12}$$

Directivity is measured in dBi, or decibels over isotropic. For an isotropic radiator D is 0 dBi.

In reality, it is difficult to know the total radiated power by antenna. Therefore, an alternative function gain is introduced as follows to replace the total radiated power P_t by total input power P_{in}:

$$G(\vartheta, \varphi) = \frac{P(\vartheta, \varphi)}{P_{in}/4\pi} \tag{13}$$

P_{in} is input power supplied to antenna which is easily accessed. The difference between P_t and P_{in} is due to the fact that some portion of input power is lost due to ohmic loss because of the finite conductivity of the metal part of antenna. The gain G is defined as the maximum value of the gain function $G(\vartheta, \varphi)$.

$$G = max\,[G(\vartheta, \varphi)] \tag{14}$$

The ratio between antenna gain and directivity is called the power efficiency $\eta = G/D$. When total input power to antenna is equal to the total radiated power, power efficiency η will be unity. In reality efficiency is always less than unity as some power is lost due to ohmic loss, as stated earlier.

Equivalent Isotropically Radiated Power (EIRP)

EIRP is defined as the product of the total radiated power P_T and antenna gain G. Hence, $EIRP = P_T G$. The unit of EIRP is given in dBi, or decibels over isotropic.

Effective Aperture and Aperture Efficiency

Let us assume an antenna in the receiving mode and a plane wave is coming from the direction of the antenna directivity. Let us also assume that the polarization of the incoming wave and antenna are aligned. In this situation, consider the antenna is receiving P_T amount of power and power density at the antenna position is S Watt/m^2. The ratio between these two quantities is defined as the effective aperture, expressed below as:

$$A_e = P_T / S \tag{15}$$

The ratio between effective aperture A_e and physical aperture of the antenna is defined as the aperture efficiency.

Huygens' Metasurface and Reflectarray as Beam Steering Antenna Array

In this section, we provide a literature review to present an overview of the reported Huygens' metasurfaces and microstrip reflectarrays those can be used as potential beam steering antennas. This section is divided into two main subsections to distinctly present the passive and active beam steering antenna arrays. In case of passive beam steering antenna array, radiated beam is steered by either rotation of the full array or changing the position of the feed antenna. On the other hand, electronic components are incorporated with the active beam steering antenna array to steer the beam electronically.

Passive Beam Steering Antenna Array Using Gradient Metasurface

Several beam steering antennas based on phase gradient metasurfaces have been realized in recent years. An anisotropic Huygens' metasurface was demonstrated that was combined with a linearly polarized circular horn antenna in order to mechanically reconfigure the direction of the main beam of the horn antenna (Wan et al. 2017). The Huygens' metasurface consists of an array of the three layers of Jerusalem cross resonators integrated in F4B dielectric substrate. By rotating this metasurface, the direction and polarization of the main radiated beam of the horn antenna can be made reconfigurable. The proposed array can be used in base-station communication and radar detection. Another Huygens metasurface-based transmit-array

was proposed to operate in the cutting edge 5G communication spectrum from 24.25 to 27.5 GHz. The proposed array consists of two antisymmetric conducting semi-circle arcs on both surfaces of a thin dielectric substrate (Xue et al. 2020). The experimental results revealed the excellent radiation performance including the peak gain of 30.7 dBi at 26 GHz, the aperture efficiency of 42.25%, and the 3 dB bandwidth of 15.7%. A similar type of coupled-resonator approach-based Huygens transmit-array was reported in Sakurai et al. (2020), where a pair of identical elliptical metal patches with an elliptical hole separated by a thin dielectric substrate was used to realize millimeter-wave beam forming at 60 GHz for linearly polarization. The Huygens transmit-array was placed on the top of a high gain 2D slot-array antenna, in its near-field. The slot array antenna was a corporate feed antenna consisted of 16 × 16 array of radiating slots with a standard waveguide input placed at the bottom of it. The beam steering principle is illustrated in Fig. 9a, where the Huygens' transmit-array acts as a broadband phase plate. Apart from the beam steering, proposed array was also demonstrated for beam diffraction and expansion. In Oyesina and Wong (2020), two Huygens' box antennas (HBA) were proposed to realize highly directive and fully steerable far field beam for microwave and millimeter-wave wireless application. The Huygens' box antenna (HBA) is an enclosure wrapped with the Huygens' metasurface elements. Two Huygens' box antennas were demonstrated in this paper. The first HBA is comprised of a single-layer Huygens' metasurface element enclosing a free-space region within a parallel-plate waveguide. Transverse electromagnetic (TEM) traveling wave is excited inside the cavity and the desired radiation is formed by leaking the cavity wave through a perforation on the top plate. By filling the cavity with the proper dielectric material and controlling the positions of current strips between two conducting parallel plates, the desired direction of the leaked beam is obtained. The second HBA model consists of air cavity between two layers of Huygens' metasurface. TM_1 (transverse magnetic) traveling wave is excited inside the cavity. The elevation direction of the radiated beam is controlled by the height of the cavity. The desired azimuthal angle of the radiated beam is obtained by the direction of wave travel inside the cavity which is in turn controlled by the current strip pairs between the

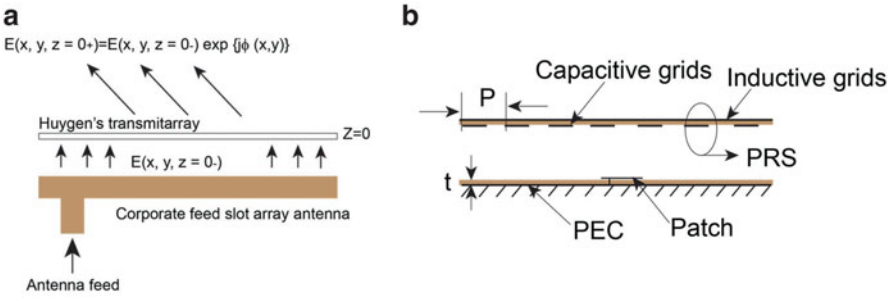

Fig. 9 (**a**) The schematic of the beam steering principle of the proposed Huygens' transmitarrays on top of a corporate feed 2D slot array structure in Sakurai et al. (2020). (**b**) Schematic view of the proposed beam steerable Fabry-Perot (FP) cavity antenna in Ratni et al. (2017)

plates. The proposed antennas can be used for the highly directive beam-forming applications in millimeter-wave frequency band. In Ratni et al. (2017), a Fabry-Perot (FP) cavity antenna was demonstrated that consists of a phase gradient metasurface-based partial reflector and a ground plane. The schematic view of the proposed antenna is shown in Fig. 9b. In order to realize the metasurface, the authors used a composite metamaterial consisting of both capacitive and inductive grids on the two opposite faces of the dielectric substrate, as depicted in Fig. 9b. The capacitive and inductive grids are respectively formed using a two-dimensional array of the metal patches and mesh stripes. Phase modulation of the partially reflecting surface is realized by varying the inductive grid widths. Four antennas with four different inductive grid widths were fabricated to steer the radiated beam at four corresponding directions. A microstrip patch antenna with coaxial feed is used as the primary source for this Fabry-Perot cavity antenna. This proposed antenna can be further upgraded by incorporating active electronic elements to obtain the Fabry-Perot cavity based electronic beam steering antenna.

Likewise, Huygens' metasurface, reflectarray antennas were demonstrated as beam steering antennas as well. As discussed in the section of design methodology, the unit cell design of the reflectarray antenna needs to fulfill two main criteria. The first one is that the unit cell should cover at least 360° phase range to avoid any phase error. The second one is the slope of the reflection phase graph should be as small as possible to avoid any error due to fabrication tolerance. However, there is a design trade-off between these two criteria. For example, if a single square metallic patch is used for unit cell, the phase sensitivity would be comparatively lower, but it cannot cover the full 360° phase range. On the other hand, if a single square ring or multiple square rings is used as the unit cell, it can cover full 360° phase range due to multiple resonances, although phase sensitivity would be comparatively higher than square patch (Rajagopalan et al. 2012; Bialkowski and Sayidmarie 2008). Hence, it is evident that attempting to increase phase range often causes higher phase sensitivity in the reflection phase graph. This issue of the trade-off was resolved in Yoon et al. (2014), three different resonators, i.e., single square patch, single ring and patch loaded ring were used as unit cells to cover the full 360° phase range while simultaneously maintain the lower slope of the reflection phase graph. Another promising design was demonstrated in Yoon et al. (2015a), where a single square ring was employed as the phasing element. In that report, in addition to conventional methods, i.e., fixed width method and fixed ratio method, two new design methods were introduced: the first method was to increase the outer length and decrease the inner length simultaneously and the second method was to increase the inner length with fixed outer length. Hence, the proposed design keeps the reflection phase sensitivity lower without sacrificing the full 360° phase range. Apart from these issues of phase range and phase sensitivity, another major disadvantage of the reflectarray is narrow gain-bandwidth (Pozar 2007). This is mainly due to the narrow bandwidth of the microstrip phasing elements. In order to solve this problem, several new broadband phasing elements are reported in recent years such as multi-layered structures (Encinar and Agustín Zornoza 2003; Encinar 2001), patches coupled to delay lines through aperture (Carrasco et al. 2008), patches attached with stub lines

(Li et al. 2012), fractal structure (Sayidmarie and Bialkowski 2011; Costanzo and Venneri 2014) and single layered multi resonant elements (Zarghani and Atlasbaf 2015; Yoon et al. 2015b; Han et al. 2017). Most of these reflectarray designs involve either multiple-layer fabrication process or thick substrate added with extra foam layer that substantially increases the material and manufacturing costs. Further, the fabrication of the stacked layers requires proper alignment between successive layers. Apart from these design approaches, another promising design approach to get broadband reflectarray is shrinking down the grid size or periodicity of the array from $\lambda/2$ to lower values (Pozar 2007; Nayeri et al. 2010; Ethier et al. 2012) so that, more parallel behavior for the phase curves at different frequencies can be realized as compared to unit cell size of $\lambda/2$. This sub-wavelength unit cell approach is different from conventionally designed reflectarrays that employ near resonant ($\lambda/2$) variable sized patches. However, these reported sub-wavelength phasing elements cannot achieve full 360° reflection phase range due to etching tolerance on small gap-size. As a result, antenna gain decreases and side lobe level increases. Hence, sub-wavelength unit cells improve gain bandwidth at the cost of antenna gain which is undesirable. This problem is addressed and solved in Niroo-Jazi et al. (2015), where multi-resonant approach and is employed as sub-wavelength unit cells. But attempting to increase phase range often causes higher phase sensitivity in reflection phase graph. In Qin et al. (2016), double square meander-line rings were employed as sub-wavelength unit cells those covered full 360° phase range without sacrificing the lower phase sensitivity and demonstrated reflectarray exhibited broadband performance as well. Another sub-wavelength reflectarray was demonstrated using two different types of phasing element, i.e., single square ring and complementary square ring to obtain broadband performance to cover full phase range of 360° with extremely low phase sensitivity (Pattanayak et al. 2018).

However, all these mentioned Huygens' metasurfaces and reflectarray antennas were demonstrated as the passive antennas. That means once fabricated, the antenna performance cannot be altered electronically. Hence, these antennas radiate highly-directive far-field beam at a particular direction fixed by the design. This highly directive pencil beam can only be steered at different directions by tilting and rotating the antenna in mechanical way. This mechanical rotation can be of different types. For example, the phase agility can be obtained using mechanical actuation that in turn changes the orientation or physical dimension of the phasing elements. For these purposes, the required micro-actuators and micromotors are fabricated using micro-machining techniques. As demonstrated in Iiuang (1995), Phillion and Okoniewski (2008), micromotors were placed at the backside of the individual elements in order to realize the mechanical phase tuning. In Fusco (2005), 10° of beam scanning was experimentally realized by rotating the patch radiators using miniaturized motor. Moreover, the low-cost stepper or linear motors were utilized to rotate the two-phase gradient metasurfaces those were placed at the near-field region of a fix beam antenna. This is referred to as near-field meta-steering system (Singh et al. 2020). In such system, metasurfaces can be placed at the height equal to 8–10 times of operating wavelength from the fix beam base antenna (Gagnon and Petosa 2013). These metasurfaces have the ability to rotate along their planes and this

rotation allows to obtain the steerable beam. However, such system produces undesired grating lobe while steering the radiated beam. This issue of undesired grating lobe was addressed and solved in Singh et al. (2020), by applying Floquet analysis and multi-objective particle swarm optimization to optimize a supercell, that is comprised of phase transforming cells arranged in a way to linearly increase the transmission phase shift. These supercells were further arranged to form the phase gradient metasurfaces. It was shown in the paper that, by increasing the transmission phase gradient of the two-phase gradient metasurfaces, unwanted lobes can be decreased below −20 dB for all the beam directions. Experimental beam steering was realized for different directions within the solid angle with 96° apex angle. The system was further improved using the metasurfaces with steeper phase gradients that can steer beam at wider range of angles. Moreover, in the improved version of the system, lesser mechanical rotation is required to realize a given scan range and the beam steering is also faster. The schematic views of the proposed metasurface and the rotation of two such metasurfaces for beam steering are respectively depicted in Fig. 10a, b.

A low-profile steerable beam antenna at 10.10 GHz was demonstrated by placing a graded index metasurface lens above a microstrip patch radiator (Singh et al. 2019). Two different types of ultrathin metasurface named linear graded index metasurface and radial graded index metasurface are used to obtain the steerable

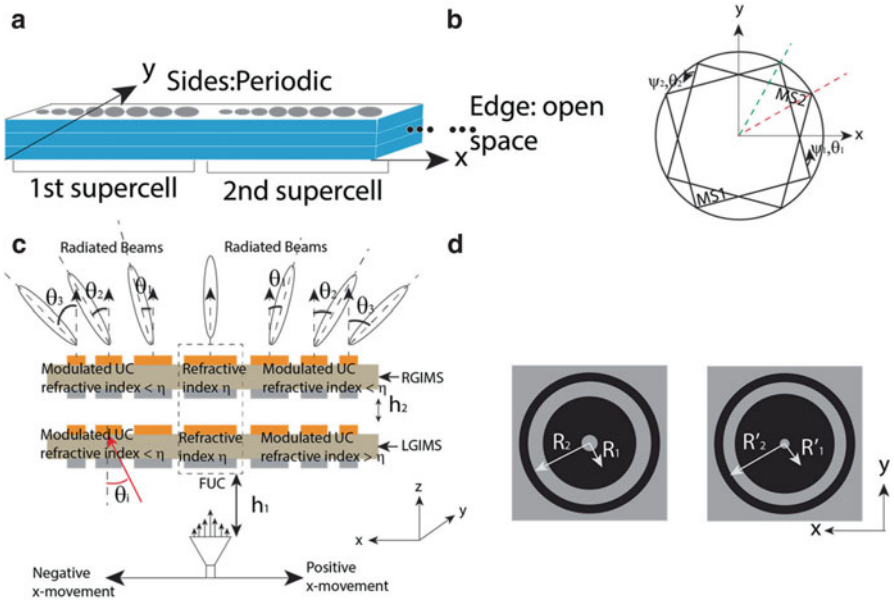

Fig. 10 (**a**) Schematic of 2-D metasurface obtained by periodic repetitions of supercells along the x-axis, (**b**) a pair of rotating metasurfaces for beam steering demonstrated in Singh et al. (2020). (**c**) The schematic of the flat top beam antenna with steered radiated beams (**d**) top (left) and bottom (right) views of the fundamental unit cell of the proposed metasurface (Singh et al. 2020)

beam. When only linear graded index metasurface is placed above the microstrip patch radiator at an optimum height, beam steering can be realized in a conical plane of 84° vertex angle. Furthermore, radial graded index metasurface is placed above the former configuration to realize a flat top beam antenna with stable gain. Figure 10c illustrates the proposed antenna schematic along with the ray trace model. The top and bottom views of the fundamental unit cell of metasurface are depicted in Fig. 10d. The movement of two metasurfaces leads to wide-band continuous beam scanning in a conical plane of 100° vertex angle with maximum 1-dB gain variation.

Mechanical actuators were also used to achieve phase tuning. Among the various types of commercially available actuators electromechanical and electrostatic-type actuators were utilized to obtain beam steering arrays. Micro-electrostatic actuators were utilized in Gianvittorio and Rahmat-Samii (2006), in order to adjust the height of the individual patch element from the ground plane. By applying the DC bias voltage between the ground plane and patch element, the patch radiator can be vertically moved that leads to the change in the fringing field. Therefore, a phase variation is realized by adjusting the control bias voltage. A multilayer reflectarray was reported in Shaker et al. (2008), where the entire mid-layer can be translated using linear-motion mechanical actuator to obtain the beam switching at 0 and $\pm 30^{\circ}$.

Apart from mechanical rotation, the wide-angle beam scanning performance can be obtained by tuning the phase distribution of the array aperture by varying the spatial delay realized by displacement of the phase centre of the feeding antenna as described in Rahmat-Samii (n.d.). In another report, the beam scanning performance was studied by laterally displace the feeding antenna. Apart from these approaches of feed displacement, multiple feeds are used to scan the radiated beam as well. This approach was adopted from the concept of the bifocal lens. Bifocal lens has wider scanning ability because of quite well focusing ability at the nearby of the two focal points as well as between these two points. A scan range of about 27 degrees (±13.5°) was demonstrated in Menzel et al. (n.d.) using a bifocal folded dual-reflectarray antenna with seven element feed array. Furthermore, the scanning range was improved to 49° by designing a reflectarray system that has restricted aperture for each beam. A bifocal reflectarray with double offset feeds was demonstrated in order to obtain wide-angle beam scanning (Rodriguez-Alvarez et al. 2012). The beam scanning performance was significantly improved with these designs at the cost of the increased design complexity attributed to the requirements of two surfaces for the phase correction. In order to resolve this issue, a single-reflectarray system with two focal points was demonstrated in Nayeri et al. (2013). In this design, the two feed antennas can be turned on either simultaneously to obtain a multi-beam performance or separately to obtain a beam-switching performance. The proposed reflectarray can steer the radiated beam up to 60°. The beam scanning performance can be further improved by realizing phase distribution using the particle swarm optimization (PSO) method and multi-objective PSO (MOPSO) method.

The drawbacks of the beam steering using mechanical rotation include slow steering, lack of compactness, lack of long-term reliability and having only one desired pattern. All these drawbacks can be eliminated by replacing the passive

Huygens' metasurface and reflectarray with active antenna array having the capability of electronic beam steering. In the next section, we present active beam steering antenna in details those are reported in the recent years.

Active Beam Steering Antenna Array Using Gradient Metasurface

In recent years, active beam steering antennas have attracted considerable attentions both from industries as well as researchers because these antennas enable the electronic way of steering the radiated beam at desired directions. In the recent cutting edge 5G and upcoming 6G wireless communication technology, electronic beam steering antenna arrays are highly demanded in order to improve the efficiency and accuracy of the communication system. There are several advantages of using these active beam steering antenna arrays which include lower profile, fast steering and reliable performance. Although, performance-wise active beam steering antennas are superior to their passive counterpart, the design complexity is naturally increased in order to incorporate and integrate the active electronic components with the antenna elements. Furthermore, several other design aspects of active array need to be considered such as routing of the control lines, power supply management, thermal management, pulsed circuitry, and environmental considerations.

One point needs to be clarified here that these active beam steering antennas are not a separate class of antenna arrays from the above mentioned reflectarrays or Huygens' metasurfaces. The basic working mechanism of active beam steering antenna relies on the same Huygens' principle. The only difference is that in case of active beam steering antenna, the phase value of each element is controlled electronically instead of the dimension of the phasing element. There are several electronic ways to control the phase value of each element of the antenna array. In recent days, several electronic components and modules are easily available commercially. These electronic components and modules can be integrated with the Huygens' metasurface and reflectarrays to make electronic beam steering arrays. The major electronic devices to obtain the reconfigurability in reflectarray antenna are p-i-n diodes, varactor diodes, field-effect transistors (FETs) and micro-electro-mechanical system (MEMS) (Nayeri et al. 2015). These electronic devices can be classified into two types. The first type includes the switching devices like p-i-n diode, MEMS switch and field-effect transistor (FET). These devices provide discrete phase control. Hence, these devices are called as digital phase shifters. The second type includes continuous phase tuning devices, such as varactor diodes and tunable MEMS devices. Therefore, these devices are called as analog phase shifters. In the subsequent paragraphs, we present recently reported active beam steering antenna arrays using these electronic devices.

In Carrasco et al. (2012), an electronically reconfigurable beam reflectarray was designed to operate at X-band of the frequency spectrum. The reconfigurability was obtained using p-i-n diode. The p-i-n diode is a semiconductor switching device which is made to behave as either open circuit, i.e., off state or a short circuit, i.e., on state by controlling the bias dc voltage across its terminals. Due to this feature, p-i-n

diode can be used as an excellent delay-line phase shifter for the phasing elements of the beam steering reflectarray (Carrasco et al. 2012; Menudier and Koleck 2012; Montori et al. 2015; Wang and Tennant 2012, 2014; Li and Abbosh 2012). The reported antenna in Carrasco et al. (2012), is consisted of 244 phasing elements arranged in a circular aperture. In order to reduce the number of the electronic devices and reducing both fabrication cost and complexity, phase control is implemented at the sub-array level, using the gathered elements those are based on the patches aperture-coupled to the delay lines. More specifically, a sub-array containing two-elements is used to control the phase of the reflected field. The antenna exhibits switching beams directed at $\pm 5^{\circ}$ by forward biasing the half of all diodes, while the other half remains reverse biasing. A third beam can be pointed toward 0° if all the diodes are simultaneously kept in reverse bias. These switching angles are reasonable values considering for the geostationary orbit as the whole Earth can be covered with $\pm 9^{\circ}$. Hence, the proposed antenna can be used for direct broadcast satellite (DBS) applications. The schematic views of the different layers of the p-i-n diode-based active phasing element and beam-switching reflectarray are respectively shown in Fig. 11a, b.

In Montori et al. (2015), another p-i-n diode-based reconfigurable reflectarray antenna was presented for the satellite communications at Ku-band. The presented

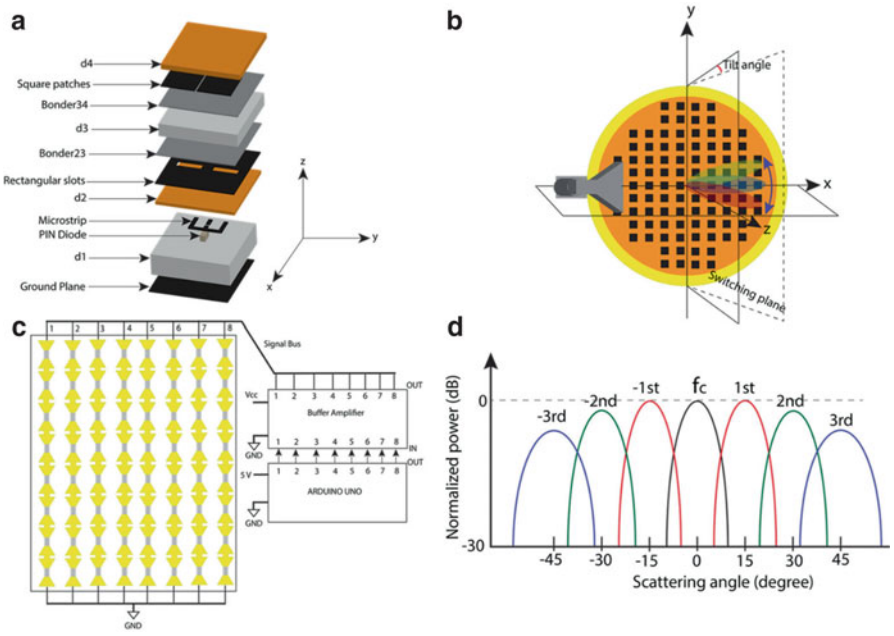

Fig. 11 (**a**) Schematic view of the different layers of the proposed phasing element, (**b**) schematic of the switching beam full reflectarray antenna (Carrasco et al. 2012). **(c)** Schematic diagram of the time-modulated reflector array prototype, (**d**) beam patterns from the TMRA at the fundamental and harmonic frequencies (Wang and Tennant 2014)

antenna architecture is comprised of a pentagonal shaped main reflectarray consisted of p-i-n diode-based reconfigurable single bit elementary cells and a passive rectangular sub-reflectarray. The sub-reflectarray is introduced in order to adopt the dual reflectarray configuration that leads to the reduced dimension with folding and transportation facilities. In this reconfigurable system, two p-i-n diodes (Skyworks DSG-9500) are mounted in antiparallel configuration on the phase shifting layer. The antiparallel configuration allows those diodes to be alternatively activated using a single control signal with the voltage values of the two-opposite polarity. In this way, only one of the two possible paths on the slot line becomes activated. The measured radiation patterns show the electronic beam steering of the antenna in the elevation plane from 30° to 90° with azimuth pointing direction of 0° and in azimuthal plane from 0° to 15° with elevation pointing direction of 40°. In Wang and Tennant (2014), a new type of reflectarray named time-modulated reflector array (TMRA) was introduced and analyzed. In TMRA, instead of using phase shifters, active scattering time of each element is controlled to obtain the desired directive beam. This controlling of scattering time is obtained by utilizing the active switching device like p-i-n diode that can be switched between non-reflecting OFF state and highly reflecting ON state. While in the ON state the element radiates isotropically. In this demonstration, p-i-n diodes were used as the switching devices to control the switching sequence of the elements in order to get the desired beam pattern. Figure 11c shows the schematic diagram of the proposed TMRA. The resulted harmonic scattering patterns at the fundamental frequency as well as first three positive and negative harmonic frequencies are shown in Fig. 11d.

Likewise, p-i-n diodes, RF MEMS switches are utilized to obtain reconfigurable reflectarray. In Bayraktar et al. (2012), a reconfigurable reflectarray antenna was demonstrated to operate at 26.5 GHz. The array is consisted of 10 × 10 antenna elements which are aperture coupled microstrip patches. Reconfigurability was obtained using RF MEMS switches which were monolithically integrated with the transmission lines of the aperture coupled microstrip patch elements. The progressive phase shift between antenna elements was realized by adjusting the length of the transmission line. The single RF MEMS switch per element was used to switch the radiated beam between broadside and 40° directions. The number of beam states can be further increased by integrating more than one RF MEMS switch per element.

Besides p-i-n diodes and RF MEMS switches, varactor diodes are used to obtain active beam steering reflectarray (Lau and Hum 2010; Kishor and Hum 2012; Venneri et al. 2013; Costanzo et al. 2016, 2018; Rodrigo et al. 2013). The varactor diode is a semiconductor p-n junction device whose junction capacitance can be varied by controlling applied bias voltage. As mentioned earlier, varactor diode is a continuous phase shifter that is integrated with the phasing element to continuously reconfigure the reflection phase angle by controlling the capacitance (Boccia et al. 2002). In Lau and Hum (2010), a reconfigurable array was presented to operate at 5.7 GHz, with a two-dimensional beam-scanning ability in the range of 60° by 30°. The unit cell of the reflectarray is consisted of two patches placed respectively on the two different sides of the ground plane and these patches are coupled through a slot cut on the ground plane. Each patch was loaded with a pair of varactor diodes, and a

third varactor diode was placed across the slot cut in the ground plane. These two patches and the loaded slot constitute three tunable resonators that can provide 360° of phase tunability. Measured beam patterns show that the proposed reflectarray has ability to perform the beam scanning with a range of 60° by 30°. In Kishor and Hum (2012), the authors presented another electronically tunable reflectarray each of its element is an aperture-coupled microstrip patch that accepts the linearly polarized wave, and that incident wave is coupled into the input terminals of an electronic phase shifter through one of the two slots on the ground plane. The phase shifter was fabricated on a Rogers substrate and MGV100-20 hyperabrupt GaAs varactor diodes from Aeroflexre were used to tune the reflection phase electronically by applying reverse bias voltage to these diodes through the bias lines. Next, the wave is amplified using a two-port amplifier. Finally, the amplified and phase shifted signal is coupled back to the microstrip patch element through another slot of the ground plane and the radiation comes out as a cross-polarized wave. Two-dimensional beam steering was demonstrated from 0° to 20° in the elevation direction and −20° to 20° in the azimuth direction at 5.7 GHz.

Although, these above mentioned reflectarray elements provide good phase tuning capability, multiple diodes per element were used to realize the phase agility. In Venneri et al. (2013), a reconfigurable phasing element, tuned by a single varactor diode, was considered to realize an active beam steering reflectarray to operate at X-band. The proposed phasing element is a rectangular microstrip patch on a grounded dielectric substrate and aperture coupled to single varactor diode loaded microstrip line, as illustrated in Fig. 12a. The line is consisted of two segments: the first segment starts from the slot and terminates on the varactor and the second one is an open-ended stub starting from the slot. In general, the tunable phase obtained by a single diode usually varies within the range of 180°, due to the capacitive nature of

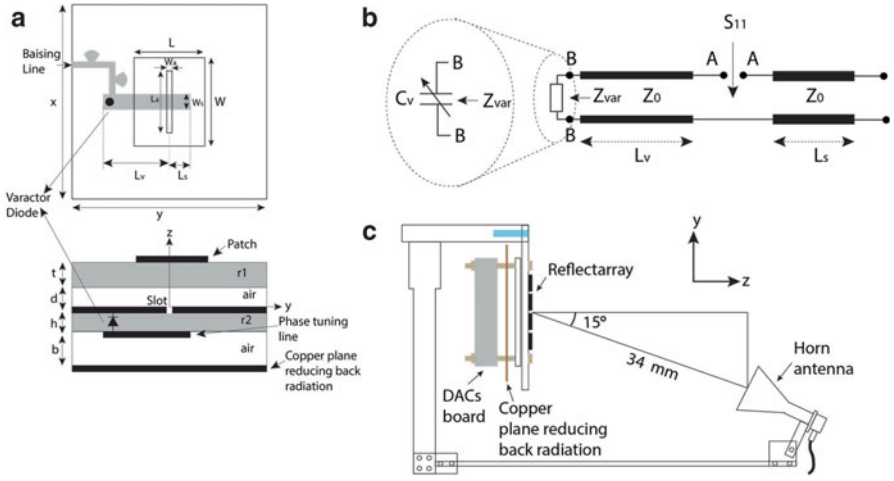

Fig. 12 (**a**) Top and side views of the unit cell schematic, (**b**) varactor loaded transmission line model and (**c**) schematic of the full reflectarray with feeding horn antenna (Venneri et al. 2013)

the varactor reactance. However, the range of the tunable phase can be increased by introducing a proper inductive load in the varactor-based circuit. In this paper, by properly adjusting the length of both line segments, the full phase range was obtained. The equivalent transmission line model, depicted in Fig. 12b represents the varactor loaded phase tuning line. From this equivalent line model, it is evident that the phase the reflected signal (S_{11}) can be tuned by varying the capacitance of the varactor diode. A microcontroller driven circuit board comprised of twelve digital-to-analog converters (DACs) was placed at the back side of the reflectarray. The output lines of the DAC board were properly connected to each varactor diode to deliver the desired bias voltages to all the varactor diodes. The schematic of the full array with feeding horn antenna is shown in Fig. 12c. This reconfigurable reflectarray exhibited a beam scan range of about 40°. So far, all of the above-mentioned beam steering reflectarrays were demonstrated for a particular fixed frequency. It was mentioned earlier that one of the main limitations of the reflectarray antenna is narrow bandwidth. Recently, several efforts are made to realize broadband active reflectarray elements those exhibited bandwidths up-to a few percentages. For example, in Riel and Laurin (2007), a stub-tuned phasing element loaded with a pair of varactor diodes exhibited a 3 dB-gain bandwidth of 3.6% and a phase bandwidth of 2.2% at the frequency of 5.4 GHz. In another demonstration (Liu and Hum 2010), a single layer reflectarray cell consisted of a double square ring integrated with three pairs of varactor diodes possesses a 2.5% phase bandwidth. A rectangular patch radiator aperture coupled to a couple of radial stubs loaded with a single varactor was demonstrated with 1 dB-gain bandwidth of 6% and phase bandwidth of 3% (Costanzo et al. 2016). In order to further improve the bandwidth, double stacked fixed size rectangular microstrip patches tuned by single varactor diode was demonstrated in Costanzo et al. (2018). More specifically, a patch resonator, printed on the top side of the stacked structure, is parasitically coupled to a varactor loaded patch on the lower side of the structure. The proposed reflectarray exhibits a beam steering from 0° up to ±40° with 1-dB gain bandwidth of 9–10% at 11 GHz.

In order to improve the bandwidth, a beam scanning reflectarray along with the frequency reconfigurability feature can be extremely useful as this system can exhibit steerable beam at multiple frequencies. In Rodrigo et al. (2013), the authors demonstrated a reconfigurable phasing element that is capable of controlling the reflection phase dynamically at different frequencies within a range which is significantly larger compared to the bandwidth of the conventional reflectarray element. In order to simultaneously realize the large phase reconfiguration and continuous frequency tuning, a pair of RF switches and a varactor diode are used per unit cell. The varactor diode is used to tune the resonant frequencies continuously by adjusting the capacitive loading of varactor whereas the RF switches are used to broaden the range of reconfiguration. Four switching configurations are realized using these active elements. Each one of these four switching configurations generates a different resonant frequency. The proposed phase element can be used in frequency-hopping RADAR and communication system for imaging and sensing application.

The active integrated electronic components mentioned so far, render the means for high-speed beam steering. However, with the current available technology, the

most of these devices cannot be used effectively at the mmwave and THz spectrum. A promising alternative solution is to utilize the exotic materials to dynamically reconfigure the scattered phase of unit cell of the array. In recent years, reconfigurable reflectarray antennas are developed based on several of such active materials such as, barium strontium titanate (BST) and Liquid Crystal (LC). However, as presented in Karnati et al. (2015), BST-integrated unit cell is unable to obtain the full 360° phase range required for realizing the array without occurrence of any phase error. A total phase range of 263° and 298° are obtained at X- and Ka- bands, respectively. Despite of the inability to obtain full 360° phase range, there are several advantages to use this material such as, fast switching speeds, monolithic integration capability, large power-handling capacity and negligible dc power consumption. The reflectarray unit cell was designed on a CSD (chemical solution deposition) grown thick BST film. This proposed unit cell paved the way to realize reconfigurable reflectarray with such exotic materials. A liquid crystal technology-based reflectarray was realized and presented in Bildik et al. (2015). As illustrated in Fig. 13a, the unit cell of proposed reflectarray is comprised of a patch, a superstrate, a substrate, a ground plane and a liquid crystal cavity. The top of the substrate was completely covered with metal to form the ground plane, and a microstrip patch antenna was printed beneath the superstrate. The liquid crystal cavity was formed by inserting a spacer layer between the ground plane and the superstrate. The orientation of liquid crystal (LC) molecules can be controlled by applying a dc bias voltage between the printed patch and ground plane. When no bias voltage is applied, the LC molecules are oriented perpendicular to the RF (radio frequency) electric field between the microstrip patch and bottom ground plane. As shown in Fig. 13a, the effective permittivity ($\varepsilon_{r,\ eff}$) of the bulk liquid crystal in this case is equal to $\varepsilon_{r,\ perpendicular}$. When a bias voltage ($V_b$) is applied, the LC molecules start to align parallel to the RF electric field, leading to a change in the effective permittivity of bulk liquid crystal, as shown in Fig. 13b. With further increase of the bias voltage, the LC molecules are completely oriented along the RF electric field lines as well as

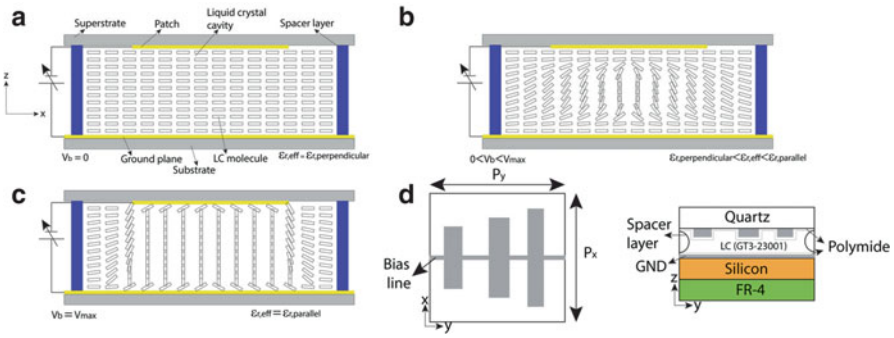

Fig. 13 Cross-section of a liquid crystal cavity between patch antenna element and ground plane (**a**) without a bias voltage, (**b**) with a bias voltage less than saturation voltage and (**c**) with a saturation voltage (Bildik et al. 2015). (**d**) Schematic of the top and side views of liquid crystal reflectarray unit cell (Perez-Palomino et al. 2015)

the bias field lines, as illustrated in Fig. 13c. Hence, $\varepsilon_{r,\ eff}$ is now equal to $\varepsilon_{r,\ parallel}$. Therefore, in order to realize the beam steering, the necessary phase adjustment can be realized by applying a bias voltage to liquid crystal layer. The beam steering ability of the proposed antenna was demonstrated by steering the main radiated beam to $-6°$, $0°$ and $6°$ at 78 GHz with a gain of 25.1 dB. Applications of such antenna can be found in automotive radar and wireless local area communication. Another liquid crystals-based reconfigurable reflectarray was realized in the frequency range from 96 to 104 GHz (Perez-Palomino et al. 2015). The unit cell schematic of the proposed reflectarray antenna is shown in Fig. 13d. The antenna can generate a steerable beam in a plane over an angular range of 55°. Due to such wide range beam scanning capability in millimeter wave range, the proposed antenna can be used for the radar or imaging applications.

Apart from reflectarray, other types of active Huygens metasurface are also demonstrated as the beam steering antenna. In Li et al. (2019), a reconfigurable metasurface lens loaded with the varactor diodes was demonstrated that can provide a DC voltage-tunable transmission phase up to 195°. The steered beam patterns were obtained by adjusting the feed positions. With increased number of the feeding sources, wide-angle beam steering up to $\pm 60°$ can be realized with this cylindrical metalens. The proposed beam steering antenna can be used for multiple input multiple output (MIMO) communication.

In addition, with the beam steering feature, reduction of radar cross section (RCS) and high gain are the two desirable characteristics for the application like radome. For realizing such tunable radome, a p-i-n diode based reconfigurable phase gradient metasurface was demonstrated to switch the radiated beam in the directions $0°$, $\pm 7°$, $\pm 15°$.and $\pm 23°$ at 4.4 GHz (Yu et al. 2019). The unit cell of the metasurface is consisted of three stacked layers and each layer is comprised of a pair of notched rings connected by four PIN diodes. A strip line was used to connect the inner ring with the adjacent outer ring. This line is also used to supply the direct current (DC) bias to the loaded p-i-n diodes. By switching the ON/OFF states of the p-i-n diodes in each of the layers, four states with four different transmission phases can be realized. A microstrip antenna along with a pair of symmetrical slot parasitic patches was used for the primary source of the radiation. This microstrip patch and the parasitic patches are surrounded by an absorbing structure to reduce the radar cross section. Moreover, the proposed antenna is capable of beam tilting at the desired directions.

The above-mentioned active beam steering antennas are becoming essential components for the commercial applications, like satellite Internet, 5G wireless communications and radar sensors for autonomous vehicles. The major issues of deploying these active beam steering antennas at the large scale for commercial purposes are design complexity and manufacturing cost due to the usage of large number of active elements. In order to address this issue, very recently, a peripherally excited phased array architecture is presented in Dorrah and Eleftheriades (2020), where number of active antenna elements are significantly reduced. The proposed design approach is based on the concept of the Huygens' metasurfaces, where an plane wave can be excited inside an otherwise passive two dimensional cavity by the

peripheral active sources. By perforating appropriate slots on the top surface of the cavity, highly directive pencil beams can be directed at desired directions. The proposed architecture can reduce the number of active elements without scarifying directivity of the antenna.

Summary

In this chapter, we present the phase gradient metasurface-based beam steering antennas. Theoretical foundation and design methodologies are discussed in detail. Recent demonstrated antennas are presented to reveal the cutting-edge beam steering technologies. It is found that mostly metallic resonator-based beam steering antennas are designed and demonstrated so far. Realizing such metal-based electrically large beam steering antenna arrays in the higher frequency range (Tera-Hertz) can be problematic as metal-based resonators possess high ohmic loss. Hence, dielectric resonator-based metasurface antenna can overwhelm in the near future. Furthermore, in order to realize intelligent beam steering, the metasurfaces can be designed with the support of machine learning and deep learning algorithms. Hence, the area of gradient metasurface-based beam steering antenna has huge research prospect to explore in near future. Specially, in the Tera-Hertz frequency range beam steering antenna using gradient metasurface can be trending research topic.

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Terahertz Metamaterial FSS for Future Wireless Communication Systems

14

Nidhi Pandit, Rahul Kumar Jaiswal, and Nagendra Prasad Pathak

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Abstract

The continuous growing demand for wireless communication with a high data rate has opened a new gate for the investigation of Terahertz (THz) frequency regime. This is because of the fact that THz frequency regime of electromagnetic (EM) spectrum can support data rate up to terabit per second due to large (several GHz) bandwidth availability which is crucial to satisfy the demand of future wireless communication. The recent advancement in semiconductor technology in the last decade has now enabled slow but steady growth toward THz gap exploration. Despite the fact that several advantages are offered by THz frequency range, such as high data rate and miniaturized circuit size, the major challenge in the direction of THz communications is large path loss which reduces its coverage area. The purpose of this chapter to investigate metamaterial frequency selective surface (FSS) and their anticipated role in successful future THz wireless communication. Starting with the detailed description on demand for THz wireless communication, the chapter discusses motivation toward the design and development of frequency selective surfaces (FSS) and their different applications in wireless communication. Further, it reviews different metamaterial techniques to resolve the anticipated challenges with conventional FSS design and its potential application in future wireless communication.

Keywords

Electromagnetic (EM) wave · Frequency selective surface (FSS) · Future wireless communication · Metamaterial · Terahertz (THz)

Terahertz Wireless Communication: Motivation

Wireless communication brought a new revolution in the world. Also, it is experiencing continuous evolution due to the change in the way of modern age society creates shares and consumes information. For example, mobile communication technology has revolutionized the world in every aspect. Today, smartphones can perform various complex tasks, and access to the world is at one arm's distance. Voice calling to video calling, live video streaming, digital payments, real-time monitoring of houses, industry, and medical patients, and all the other exciting information are just available at the fingertips. Also, wireless communication has been used everywhere, such as defense, satellite, radar, civil fixed point, and mobile communication applications. A brief overview of the applications is provided in Figs. 1 and 2.

High-frequency EM wave has a long history of wireless communication purposes. Electromagnetic spectrum and roll of microwave to THz frequency bands with regard to different communication applications have been depicted in Fig. 2. Microwave frequency signals are nonionizing, low power, and can travel far with a

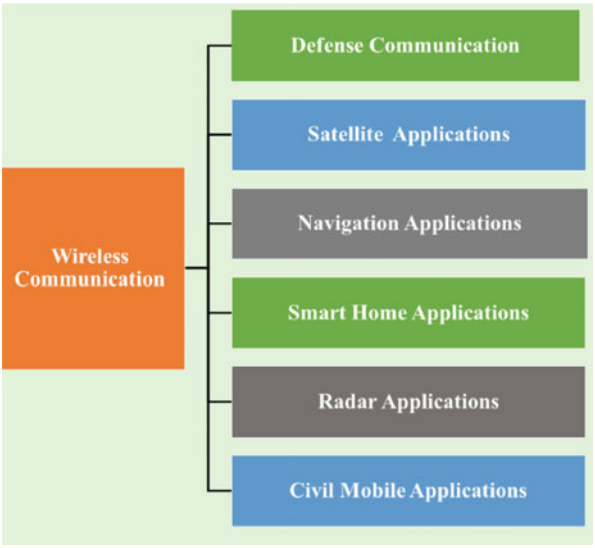

Fig. 1 Different wireless communication applications

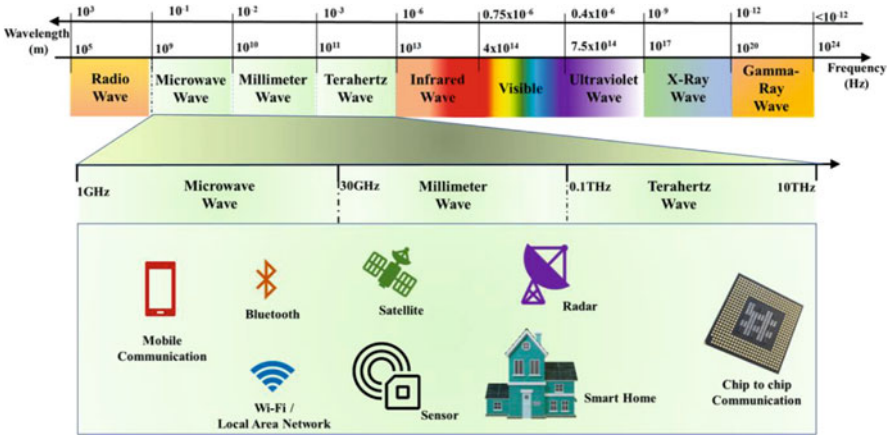

Fig. 2 Different wireless communication applications

little attenuation in the atmosphere, and so it has been widely exploited for wireless communication. However, according to the facts available from statics, the continuous growing demand for wireless communication needs much more bandwidth than microwave frequency regime signals. For example, it is expected that over 24.1 billion devices will be connected to the wireless network by 2030 due to the Internet of things (IoT) and Machine to Machine (M2M) communication paradigm. In addition to the growth of a number of interconnected devices, there will be a huge exploding need for a high data rate in the near future. Following this trend, the demand for a high data rate is doubling every 18 months (Huang and Wang 2011).

The next generation of wireless communication 6G will require a data rate at least 100 times more than the currently provided network, i.e., in several Terabits per second (Tbps) (Sarieddeen et al. 2020) with ultralow latency around 1 ms, reduced energy consumption, and better reliability, security.

To achieve this need of huge data rate, why are researchers looking toward the THz frequency regime? One needs to revisit the Shannon-Hartley theorem. Shanon-Hartley theorem relates the capacity of a channel (C) in bits per second with the signal to noise ratio (SNR = S/N) and the total available bandwidth (BW) as given by relation (1).

$$C = B \log_2(1 + S/N) \text{bits}/s \tag{1}$$

It is abundantly clear from the Eq. (1) that channel capacity can be increased either by enhancing the bandwidth or/and increasing the signal-to-noise ratio (S/N). For human health safety reasons, the signal power (S) cannot be increased beyond an extent. Hence, researchers are continuously exploring higher frequency regimes. This is because of the fact that at higher frequency regimes, higher bandwidths are available. Knowing the fact that at microwave frequency, the available bandwidth for communication systems is limited (a few MHz), the fifth- generation (5G) mobile communication systems are already exploring the millimeter-wave (mm- Wave) frequency band as an alternative along with the sub-six GHz communication technology. Millimeter-wave (mm-wave) frequency has the ability to provide much more data rate as compared to microwave frequency (Huang et al. 2017; Bae and Choi 2015; Zhang et al. 2017a; Rappaport et al. 2011). For example, at 60 GHz (in millimeter wave band), the available bandwidth is 7 GHz. Although the achievable data rate at the mm-wave frequency regime is several Gbps (Agrawal and Sharma 2016), however, the foreseen requirement is much more than this. This could be understood this way: The future wireless local area network (WLAN) and wireless personal area network (WPAN) will require a data rate of at least a couple of ten Gbps. Further, the data rate in virtual and augmented reality devices is expected to be 10 Gbps. In addition to this, the data rate requirement for uncompressed ultra-high-definition videos and 3D videos is expected to be 24 Gbps and 100 Gbps, respectively (Mumtaz et al. 2017). Hence, currently, there is a pressing demand for researchers to explore much higher frequencies than mm-wave.

There are two alternatives at a higher frequency. The first is THz frequency band, and the other is free space optical (FSO) communication. FSO has gained much attention because of its unique advantages like very large bandwidth and high data rate, less power, and license-free spectrum (Tonouchi 2007). However, FSO communication technology has its own limitations because its performance is seriously influenced by natural environmental conditions like clouds, fog, rain, haze, etc. Accordingly, another alternative frequency band is demanded to settle the problems existing in FSO communication systems. Instead of the aforementioned frequency regions, terahertz (THz) electromagnetic radiation (from 0.1 THz to 10 THz) lies in the boundary region between optical frequency. THz frequency provides huge bandwidth and data rate up to Tbps, and it fills the gap between mm-wave and

FSO communication technology. THz frequency communication has been envisioned as one of the most popular spectrum bands to enable ultra-high-speed communications for beyond 5G.

THz frequency spectrum lies in the electromagnetic spectrum between mm-wave and infrared and support signal from 0.1 THz (100 GHz) to 10 THz. It has been envisioned as a great spectrum resource with enormous potential for different applications, including communication (Huang and Wang 2011). In terms of the wavelength, the range varies from 3 mm to 0.03 mm, as shown in Fig. 1. This range of frequency is also called as THz gap because, in this region, technology is under development and in its initial phase (Tonouchi 2007; Federici and Moeller 2010; Song and Nagatsuma 2011; Ostmann and Nagatsuma 2011; Nagatsuma et al. 2013; Akyildiz et al. 2014; Kurner and Priebe 2014; Siegel 2007). Some of its salient features as compared to other frequency parts are as follows:

1. Unlike millimeter-wave communications, THz frequency offers the large available bandwidths and supports several hundred Gbps to Tbps data rates.
2. Compared to free space optical communication, THz communication system is less prone to the atmospheric effects (like cloud, fog, etc.) for outdoor wireless communications.
3. Owing to the shorter wavelengths of THz signals, THz communication systems can support higher link directionality.
4. THz technology components can be realized in much smaller footprints and possess higher resilience to eavesdrop.
5. THz communication systems have additional merit; they can utilize the reflection paths to enhance the link gains.
6. THz band assures to support higher user densities, higher reliability, less latency, and more energy efficiency.

Research Challenges and Potential Solutions for THz Communication

To clearly understand the research challenges of THz communication, it is very important to revisit different components and factors which affect the effective communication link.

Wireless Communication Data Points and Communication Link

Figure 3 shows two wireless data points which are connected through a wireless link. Each data point (for example, node A and node B) comprises a wireless transceiver system and comprises many electronic parts called “circuits” or “components.” These are designed and fabricated to carry out various signal-processing functions, such as signal generation, modulation, and amplification.

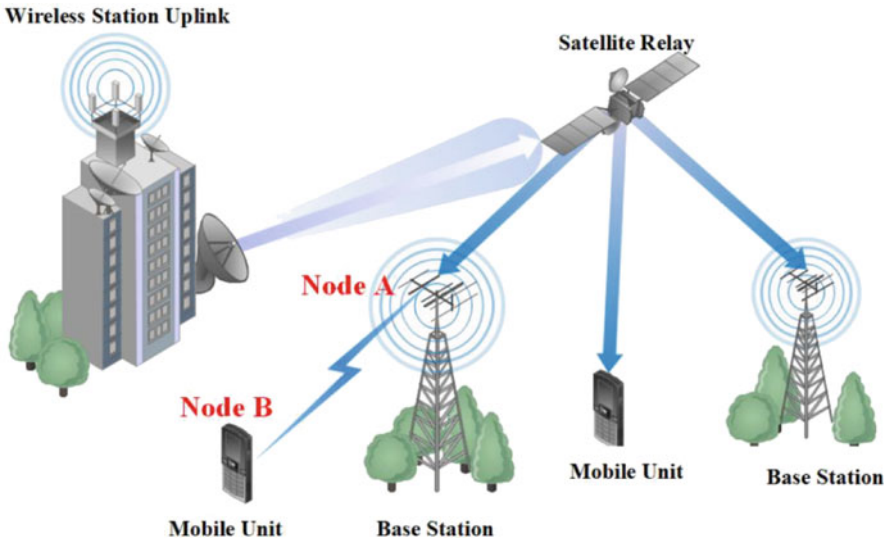

Fig. 3 Wireless communication scenario

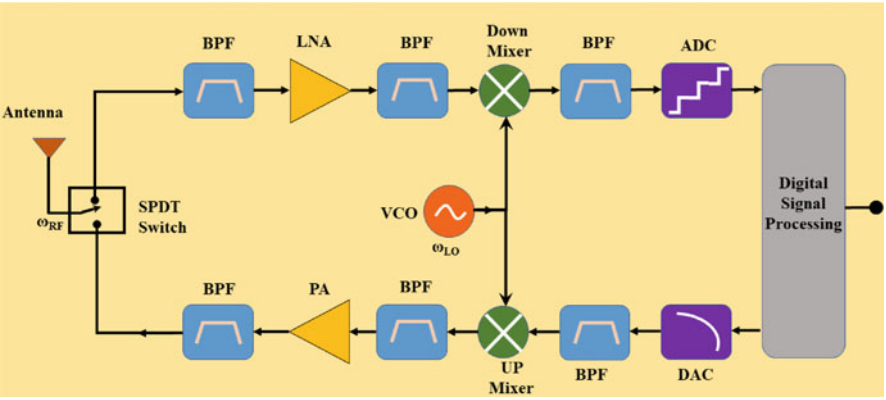

Fig. 4 General block diagram of transceiver architecture

Figure 4 presents a general transceiver architecture of a wireless communication system, which consists of a transmitter and a receiver that is connected to an antenna through a switch. The amplifier and power amplifier, switch, coupler, phase shifter, up- and down-conversion mixers, antenna, oscillator, and filters.

The power loss and noise in the system decide the overall signal-to-noise ratio (SNR) of the wireless system data point. To implement this signal-processing circuit component, many technologies like monolithic semiconductor or hybrid technologies have been used. For example, at terahertz frequency, THz heterojunction

semiconductor technology, which is growing nowadays, has been used. The THz systems are being implemented in heterodyne architecture in monolithic integrated circuits (MICs) environment, which provide low-noise performance, high-power efficiency, and output. The rapid development in semiconductor technology pushed the transit and maximum oscillation frequency of the transistor in THz regime. Today's state-of-the- art SiGe and InP heterojunction bipolar transistor (HBT) has achieved f_{max}/f_t of 350 GHz/300. GHz technology is most promising due to having f_{max} of 1THz and beyond (Lai et al. 2007). This leads to the development of the various THz monolithic integrated circuit (MIC) components, e.g., mixers, switches, and RF amplifiers up to 1 THz and beyond.

In the scenario given in Fig. 3, If Data Point A and B are communicating to each other, then the received power at node B is calculated using the Friis formula (Balanis 2012) given as (2):

$$P_t = \frac{P_t G_t G_r \lambda^2}{(4\pi R)^2} W \tag{2}$$

Where *Pt.* Total transmitted power of node A, *Pr′.* Total received power of receiver at node *B, Gt.* Antenna gain of transmitter node A, G_r: Antenna gain of receiver node B, *R:* distance between node A and B, λ*:* operating wavelength.

In Eq. (2), free space path loss (FSPL) is an important factor that limits the received power at node B and given as (3a) (Balanis 2012):

$$\text{FSPL} = \frac{(4\pi R)^2}{\lambda^2} = \left(\frac{4\pi Rf}{c}\right)^2 \tag{3a}$$

And it can be expressed in dB as follows in (3b) (Balanis 2012):

$$\text{FSPL}(dB) = 20\log R + 20\log f - 147.55 \tag{3b}$$

To successfully recover the information at the node B, its receiver needs minimum power *P*min which is given as (4) (Balanis 2012):

$$P_{\min}(dB) = -174dBm + 10\ \log B + NF\ (dB) + \left(S_0/N_0\right)_{\min} \tag{4}$$

where, *B*: bandwidth, *NF*: noise figure of receiver system, $(S_0/N_0)_{min}$: minimum signal to noise ratio to successfully recover the information.

By observing relation (2)–(4), it is evident that received power is inversely proportioned to free space path loss (FSPL) and directly proportional to transmitted power (P_t); and both the transmitted and received antenna gain (G_t) and (G_r), respectively. Further, to recover the information successfully at the receiver side, total bandwidth (B) and noise figure (NF) play an important role. For successful communication, all of the above parameters play a crucial role.

THz Waves and Challenges for Wireless Communication

As discussed above, THz communication offers many useful features. However, in the direction of THz communication, there are a couple of research challenges. Some of them are listed as follows (Akyildiz et al. 2018):

1. The major challenge for THz communication is the very high propagation loss at THz-band frequencies, which limits the communication link for very short distances. Specifically, the FSPL of a 10-m link at THz frequencies can easily exceed 100 dB, which makes communication above tens of meters extremely challenging.
2. It is evident from the Friis' transmission formula given in Eq. (3a) that doubling the frequency quadruples the free space path loss, which is due to reduced antenna aperture as a function of frequency.
3. THz frequency signals are very much prone to the molecular absorptions, and water vapors in the natural environment result in band-splitting and bandwidth reduction.
4. The field of THz electronics is evolving, so currently THz communication system has substantial hardware challenges that have to be developed in terms of the THz solid state heterodyne receiver.

Different Potential Solution for Improving the THz Communication Link for a Large Coverage Area

1. **Ultra-Massive (UM)-Multiple Input Multiple Output (MIMO) Antenna System:** To combat the atmospheric fading of the THz signal and to increase the aperture of the antenna system, ultramassive multiple input and multiple output antenna technology is being seen as a potential solution. It uses spatial diversity and increased directivity of the antenna array system to enhance the coverage of the antenna in THz transceiver system. Theoretically, this technique is regarded as the most promising approaches for terahertz (THz) band communication (Akyildiz and Jornet 2016). However, practice implementation point of view has many new challenges that hinder their prospect of practical implementation of future deployment. In particular, it requires an extremely large number of high-frequency hardware chains to feed the signal to each element of the UM-MIMO antenna system. This makes their hardware implementation very complex and costly, where not only designers have to work on high-performance RF component designs but also need to work on their isolation of one chain to another. Especially at THz frequency, feature size becomes very small, and the electromagnetic shielding between RF chains and different antenna elements in UM-MIMO system will be very difficult.
2. **Reflective Arrays**: Reflected arrays are two-dimensional surfaces, which utilize the concept of both phased arrays and geometric optics, and signals can be redirected in a particular direction to enhance the coverage. In reflecting array

surface, using the electronic tuning, the phase shift of each element can be controlled and will jointly form an array pattern to receive or transmit the signal to or from desired directions. In environments with dense multipath at mm-Wave and THz bands, reflect arrays can be deployed to serve single and multiple transmitters and receiver pairs to communicate simultaneously and to extend the transmission range (Hum and Perruisseau-Carrier 2014).

3. **Metasurfaces and Frequency Selective Surfaces (FSS)**: Metasurfaces can be defined as a (symmetric or asymmetric) or/and periodic (or aperiodic) structure consisting of an array of subwavelength resonators called meta-atoms which control the electromagnetic response of the surface (Yu and Capasso 2014). The distribution of the meta-atoms (scatters) plays a pivotal role in determining the response of a surface with respect to waveshaping. This property differentiates the metasurfaces from frequency selective surfaces (FSS). A traditional FSS has individual elements (periodicity), which are of the order of the operating wavelength (generally – $\lambda/2$) and usually symmetric and also used to control the electromagnetic performance of the incident EM wave for a particular range of frequency it has been designed for (Vardaxoglou 1997). The FSS perform various operation to control the electromagnetic response of incident electromagnetic wave and are very useful for THz communication due to its planar nature. Also, these surfaces give large aperture area and could be very useful for many THz communication antenna systems. A detailed description about the frequency selective surfaces (FSS) has been given in the following section.

Frequency Selective Surface (FSS): Brief Introduction

Frequency selective surfaces are spatial filtering surfaces composed of an array of elements in a periodic fashion. The FSS has a specific response (i.e., reflect, transmit, or absorb) to the incident field and incident wave angle and/or polarization for a certain band of frequencies. The frequency response of the FSS is dependent on its structural parameters and the amount of electromagnetic coupling of an incident electromagnetic wave to it. Hence, this frequency of selective surfaces depends not only on its physical parameter but also on the polarization of the incident EM wave (Vardaxoglou 1997; Munk 2000).

Frequency Selective Surface (FSS): Working Mechanism

According to a very fundamental definition, FSS is the two-dimensional periodic structure of the element. This element can be different shapes depending upon the requirement. However, for all the FSS structures, the operational mechanism can be explained through the same resonance phenomenon. For example, consider an array of the element on a planar surface. When a plane wave is an incident on this surface, the electromagnetic field will couple to each element. Each element of the FSS structure will be resonant, where their effective electrical length is $\lambda/2$ with respect to

the incident electrical field (Wu 1995). Corresponding to the phase front of the incident wave, these elements have a certain phase delay. As a result, the scattered radiations of individual elements add up coherently and give desired response for the frequency of operation.

Different Types of Frequency Selective Surface

Based on the physical geometrical parameters, FSS can be broadly classified into three classes: metallic, all-dielectric, and hybrid/combined. Further, these structures can be cascaded to achieve higher performance, such as frequency selectivity. Based on the frequency characteristics, it can be categorized as low-pass, high-pass, bandpass, and bandstop.

Metallic Frequency Selective Surfaces

The frequency selective surface lying in this class consists only of metallic elements. There are different kinds of shape geometries; however, the most fundamental geometries are metallic patches and wire-grids. These metallic shapes define the frequency response of the FSS. The metallic patches behave as a capacitive surface and give a lowpass filter response (Munk 2000). The gaps (inter-element spacing) between the square patches in vertical and horizontal direction create capacitive effects. Similarly, a wired grid acts as an inductive array and gives a high-pass response. A conceptual illustration for both cases has been depicted in Fig. 5.

All-Dielectric Frequency Selective Surfaces

To achieve frequency selective operation in an all-dielectric medium, different techniques are used, such as multilayered dielectric materials and guided-mode resonance. As the dielectric has a loss tangent and attenuates the signals, they are not much preferred for transparent surfaces. However, dielectric FSS are mostly used as reflective surfaces or absorption surfaces (Munk 2000).

Hybrid/Combined Frequency Selective Surfaces

In hybrid frequency selective surfaces (FSS), the metallic type of elements can be used on top of the dielectric substrate to gain mechanical strength, reduction in the thickness, and good impedance matching. In these structures, the overall response is the combination of metal structure and dielectric substrate underneath. Additionally, along with the metallic conducting resonating elements, the coupling in between the conducting elements plays an important role in determining the overall response of the complete surface. The frequency response of metallic FSS is also shifted when it

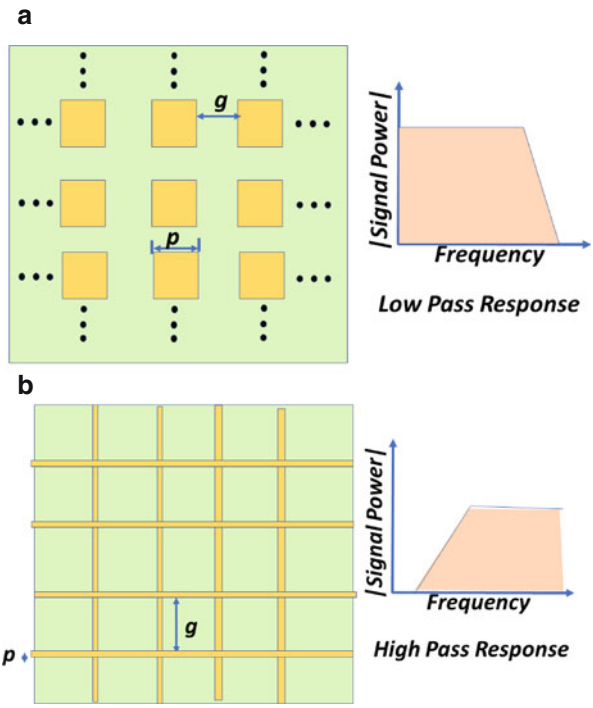

Fig. 5 (**a**) Frequency selective surface consisting periodic metallic patches with its corresponding low pass behavior; (**b**) frequency selective surface consisting periodic metallic wire grids with its corresponding high pass behavior

is backed by the dielectric layer with dielectric constant *Sr* of infinite thickness, and it is given by (5) (Munk 2000).

$$f_r = f_0/\sqrt{\varepsilon_r} \tag{5}$$

However, for a dielectric layer on both sides of the periodic structure, the resonant frequency will be given by

$$f_r = f_0/\sqrt{(\varepsilon_r + 1)2} \tag{6}$$

Here f_o is the resonant frequency of the FSS structure without dielectric loading.

As discussed before, the wired grid structure shows high-pass response, and metallic patch geometries show the low-pass response. Similarly, in order to obtain the bandpass characteristics, a wire-grid frequency selective surface is loaded with additional metallic patches, as shown in Fig. 6. The metallic grid is responsible for the inductance of the structure, and these additional metallic patches will introduce the capacitance. So, a strong resonance can be obtained for a particular range of frequency where reactive impedance offered by the overall inductance and capacitance is canceled out. In the combined/hybrid topology, it should be kept in mind that the total response has been obtained by combining wire-grids, and patches will not

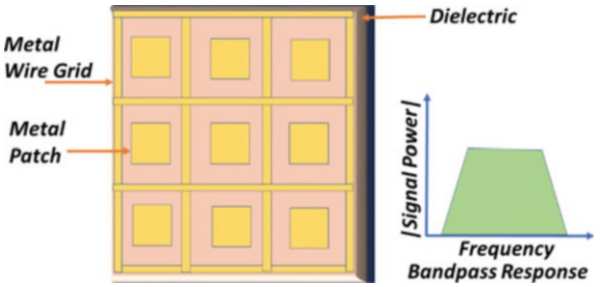

Fig. 6 Frequency selective surface consisting periodic metallic patches and wire grids with its corresponding bandpass behavior

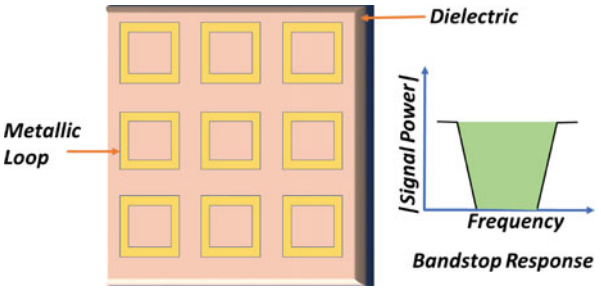

Fig. 7 Frequency selective surface consisting periodic metallic loops with its corresponding bandstop behavior

be a simple addition of the two responses, but the effect of the dielectric layer and the mutual coupling will have to be considered.

Similarly, its complementary structure will provide the bandstop characteristics. A conceptual illustration for the same has been given in Fig. 7.

Cascaded Periodic Structures

In general, the single-layer FSS are narrow-band in nature and show frequency selectivity only at a certain frequency. However, there are many applications where the wideband response is needed. Also, in some cases, a wideband response is desired, along with incident angle stability and sharp edge roll-offs. For achieving these characteristics, cascaded structures are used. There are mainly three to cascade these FSS structures, and their description has been given below (Munk 2000).

Metal-metal cascading: In this arrangement, two or more metallic FSS are cascaded behind each other without any dielectric layer loading in between them, as shown in Fig. 8a along with their response as shown in Fig. 8b. For a double-layer structure, the frequency response roll of sharply near the cut-off frequency and has quite flat amplitude response in the bandpass region in comparison to single layer structure. However, this arrangement is very sensitive to the incident angle and

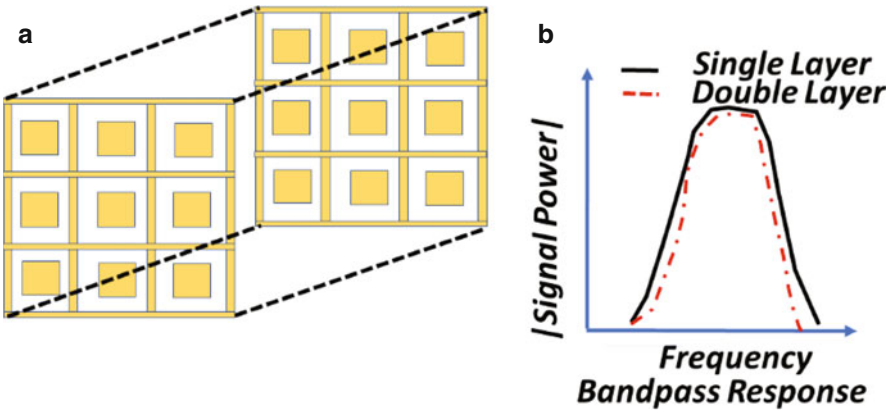

Fig. 8 (**a**) Two-layer frequency selective surface consisting of periodic metallic loops and wire grids; (**b**) conceptual illustration of the frequency response for a single-layer and double layer

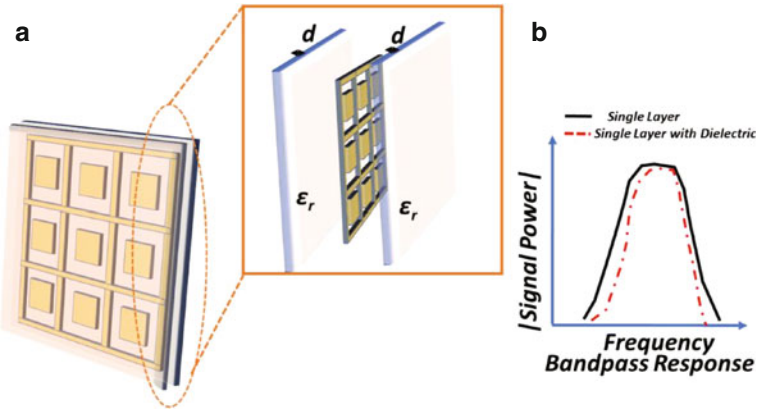

Fig. 9 (**a**) A single-layer FSS sandwiched between two dielectric slabs; (**b**) conceptual illustration of frequency response for a single-layer without dielectric slabs and with dielectric slabs on both sides

shows considerable bandwidth variation for oblique incidence angles and different polarizations.

Dielectric-metal-dielectric: The limitation of angle stability can be overcome by the dielectric-metal-dielectric arrangement as shown in Fig. 9a. However, the thickness and dielectric constant of the loaded dielectric slab on each side of the metallic surface must be the same. It may provide considerable bandwidth, as shown in Fig. 9b. However, a dielectric superstrate contributes to the additional insertion loss. Also, this kind of arrangement is a little bit challenging for conformal applications where FSS covers the structure with curved surfaces.

Design Parameters of Frequency Selective Surfaces

Traditionally, a frequency selective surface is a resonance-based surface. The resonance characteristics of a resonance-length-based FSS usually depend on the way the surface is exposed to the electromagnetic wave. Different design parameters and their corresponding effects are discussed as follows.

Element Geometries: As explained in the above section, frequency selective surfaces are resonance-based structures. It is very much dependent on the aperture shape of each element. It defines many factors such as how much electromagnetic field will be coupled with the element respective to the incident angle, its overall frequency behavior (Low pass, high pass, bandpass, and bandstop), and its packing density which will decide the grating lobes. Based on element geometries, FSS are broadly classified into two types: (a) patch-type elements, and (b) aperture-type of elements. As already mentioned in the above section, FSS structures' two complementary planar arrays, an array of patches and an array of slots, are usually considered. A simple structure consisting of a periodic array of metallic conducting patches shows a low-pass characteristic (Ulrich 1967; Lee et al. 1982; Mittra et al. 1988). The complementary structure, i.e., wire grid, has inductive nature and shows high-pass response. Over the years, several FSS element geometries have been introduced for bandpass and bandstop applications. Some of the most popular shapes are circular (Mittra et al. 1988; Potter 1974); metallic plates such as rectangles and dipoles (Ott et al. 1967; Chen 1967; Montgomery 1975; Schneider and Munk 1994), cross-poles, tripoles, and Jerusalem cross (Pelton and Munk 1979; Hamdy and Parker 1982; Tsao and Mittra 1984; Lagley and Drinkwater 1982; Pelton and Munk 1974); three- or four-legged dipoles (Munk 1974); rings (Parker and Handy 1981; Parker et al. 1981; Huang et al. 1994); square loops (Langley and Parker 1982; Langley and Parker 1983); and gridded square loops (Lee and Langley 1985); etc. as depicted in Fig. 10. Depending on the application requirements, different FSS

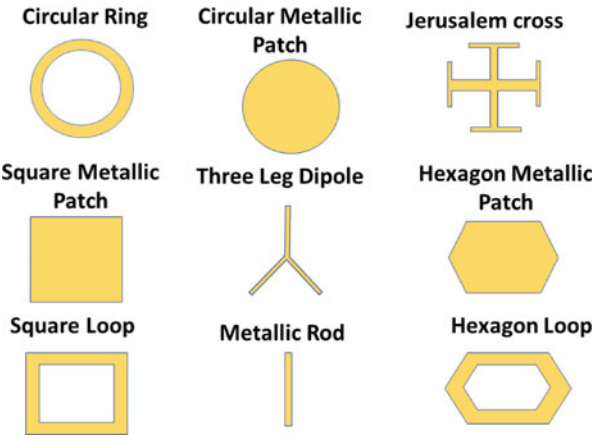

Fig. 10 Traditionally popular FSS elements over past decades

designs can be chosen to fulfill the demands. These requirements usually include frequency behavior (low-pass, high-pass, bandpass, and bandstop), level of dependence on the incidence angle of the incoming wave, the level of cross-polarization, bandwidth, and the level of band separation.

Element Dimensions: An FSS is a traditional design based on resonance elements. The dimensions of the resonating elements play a vital role. For example, FSS containing an array of dipoles creates transmission notches at multiple frequencies where the size of the dipole element is a multiple of half a wavelength. A similar effect can explain the operation of other elements. The square loop element, for example, can be imagined as two dipoles that are connected to one another at each end. Using the same argument as that of the dipole, a loop resonates when the length of the two sides equals the length of a resonant dipole, $\lambda = 2$. The length of each element of the square patch antenna decides the capacitance of the overall FSS structure and provides a low-pass response.

Periodicity and arrangement: Although the shape of the elements has the utmost importance effect in the frequency response, however, the way these elements are arranged in the array format and their periodicity is also part of the design work as we know that FSS is very sensitive to the incident angle of the incoming EM wave. So, the element should be very densely packed to have more angle stabilization which is decided by element shape, their arrangement, and their periodic distance. Also, if the element is not densely packed, it results in a diffraction phenomenon, and so the grating lobes are undesirable. Hence, to delay these grating lobes, it is also required to arrange this element in very near proximity. In addition to this, the periodicity of the elements also defines the inter-element coupling, which decides the overall frequency response. For example, in square patch elements-based FSS, the periodicity defines the overall capacitance between two patches and so the overall capacitance of the FSS structure and the cutoff frequency (Munk 2000).

Substrate: In a hybrid configuration, where both metal elements and dielectric are used, the overall response is greatly affected by the dielectric substrate also. Especially where the miniaturization of the picture comes into the picture. If we have miniaturized elements, then the same area is packed with more elements, and one will get more angle stabilization, delayed grating lobes. The high-dielectric constant material element much shorter than the wavelength can be designed to create the inductive and capacitive traces.

Different Wireless Applications of Frequency Selective Surfaces

When electromagnetic wave radiation is exposed to frequency selective surface (FSS) for a particular range of frequencies, it performs different phenomena such as filtering, absorbing, shielding, polarizing, phase changing, redirecting, etc. All these functions are very crucial for wireless communication and safety purposes. Some of its main applications are described below.

For Radome Applications

As FSS can transmit a certain frequency range signal and reflect other frequency spectrum signal. This property of the FSS can be utilized for hiding purposes. For example, an FSS can be a cover for hiding communication facilities in defense and considered as some potential applications of the FSS. As they are used to cover something, these covers are named radomes. Radomes are bandpass FSS filters that are used to reduce the radar cross-section (RCS) of an antenna system outside its frequency band of operation. Basically, "radar cross-section is the ratio of the scattered power by an object to the power incident on the object." In Balanis (2012)), it is defined as "the area intercepting the amount of power that, when scattered isotopically, produces at the receiver a density that is equal to the density scattered by the actual target." The concept of stealth or being able to operate without the knowledge of the enemy is always one of the main goals of military technology. In order to minimize the detection, FSS layers cover the facilities (Liu et al. 2018; Omar and Shen 2019; Narayan et al., 2018; Zhou et al. 2012), and for a designed range of frequency signal, radomes (FSS) are electromagnetically transparent, which reduces the RCS, and for the frequency other than design frequency, it reflects the signal.

As an Absorber

Absorbers are used to attenuate the electromagnetic signal energy. Absorbs are generally used to reduce the reflections. It is very necessary in communications systems and other applications; otherwise, reflected signals cause interference in system operation. Conventionally, there are different materials which are used for this purpose, such as ferrite tiles, carbon nanotubes, resonant absorbers, pyramidal absorbers, etc. (Baek et al. 2011; Naito and Suetake 1971). Also, the Salisbury screen, Jaumann absorbers (Ke et al. 2011), and pyramidal absorbers are very famous. However, planar FSS-based absorbers are much easy to fabricate and give very promising performance (Zhang et al. 2017b). These FSS are resonant-based absorbers and also sometimes called analog absorbers. One example of FSS-based absorber has been shown in Fig. 11, where FSS is conductor-backed to avoid any kind of transmission. This conductor-backed FSS is now designed in such a way that it matches the free space impedance, and minimum reflection and maximum absorption have been achieved. Single-layer FSS-based absorbers are usually a narrow band. Hence, to further enhance the bandwidth of the structure, multiple stacked FSS with resistive conductive metallic elements are used.

In Filtering Applications

There are many wireless applications where spatial filtering is required. For example, massive MIMO antenna systems are very popular. However, they require an RF chain for each element. This problem becomes more challenging at the THz

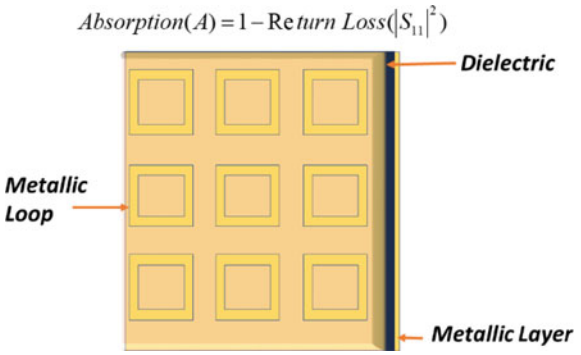

Fig. 11 Conceptual illustration of FSS-based planar absorber

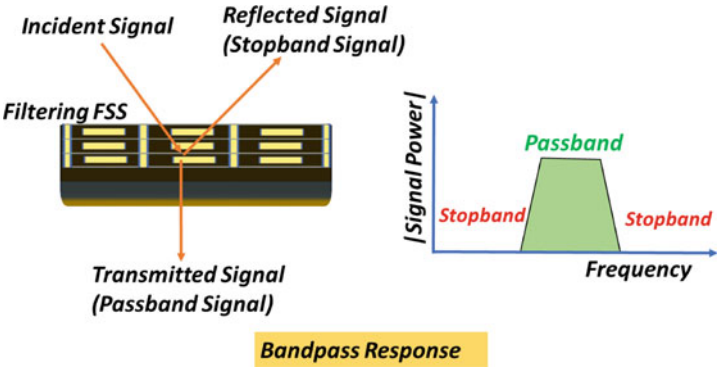

Fig. 12 Conceptual illustration of FSS use as a spatial filter before antenna arrangement

frequency regime because of the very small feature size. To reduce the hardware complexity, spatial filters can be used before the MIMO antenna system as shown in Fig. 12. Using this arrangement, complexity and filtering requirement in each chain can be avoided. Similarly, in waveguide systems to prohibit some frequency signals or to pass certain frequency signals, FSS structures are used (Bashiri et al. 2017; Sampath and Sivasamy 2020; Liu et al. 2020).

In Different Antenna Applications

FSS can be used as subreflector (redirecting element) for an antenna and increase the aperture size of the antenna and also its performance (Wu 1991). Further, frequency-selective surfaces can be used in the design of reactive impedance surfaces (Lee and Fong 1972). Using this concept, where reflective characteristics with a certain phase and amplitude can be achieved (Lam et al. 1988), unlocks a

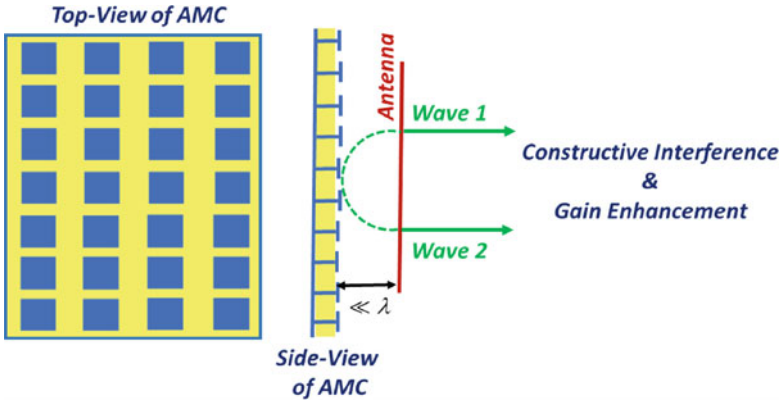

Fig. 13 Conceptual illustration of FSS as AMC for enhancing antenna gain by reducing surface current

whole new exciting area of research, in addition to frequency filtering. FSS structures are utilized to produce a desired reactive response. FSS structures are being used in the design of artificial magnetic conductors (AMC) (or high-impedance surfaces) (Broas et al. 2001), which are playing a crucial role in improving the performance of the antenna. Traditional antenna with metallic ground plane suffers from surface wave, and image current reduces their effective radiation efficiency. Instead of metal, AMC can be used as a ground plane for an antenna in order to suppress the surface waves inside the substrate, which helps to enhance the gain characteristics and reduces losses and grating lobes. A conceptual illustration has been shown in Fig. 13. Another area of interest is the application of FSS as a superstrate over the printed antennas to improve their radiation characteristics. The problem of enhancing the gain of printed circuit antennas has been extensively studied in Broas et al. (2001)). A stack of electric and magnetic superstrate is being used above antenna to implement lens like characters. These layers are chosen and arranged in such a manner that they behave like a lens, and bend the rays emanating from the antenna and incident upon the interface of the substrate and the stack of the superstrates according to Snell's law.

Shielding Applications

Electromagnetic interference (EMI) causes performance degradation of the electronic system, transmission channel by an external electromagnetic disturbance. Similarly, electromagnetic compatibility (EMC) is the ability of electrical equipment or a system is to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbances to anything in that environment. Bandstop behavior of the FSS, hence, provides shielding and is used for EMI and EMC in different wireless applications (Suresh Kumar et al. 2015). A conceptual

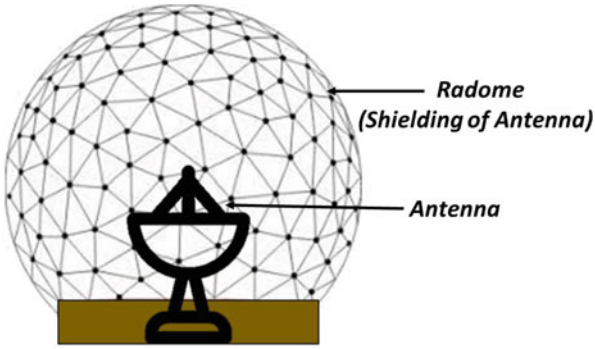

Fig. 14 Conceptual illustration of FSS as shielding application for antenna at operational frequency

illustration of the same has been provided in Fig. 14. Even this approach is now being used in many modern architectures, which are shielding their premises from outside electromagnetic environments (Parker et al. 1997; Hirai and Yokota 1999).

Limitation of Conventional FSS Structures

In traditional FSS designs, the mutual interaction of resonating elements also plays a vital role. Hence, to achieve desired sharp transition frequency response, a large number of the resonating unit cells must be present. Consequently, by arranging a large number of cells, the overall surface becomes very large electrically. Also, for a large surface, maintaining the same incident angle is very difficult and results into grating lobes which are undesirable. In addition to that, FSS structures, where resonating element size is comparable to incident wave, experience higher-order harmonics problem which further degrade the performance. In many antennas-related applications, where these FSS are used as a AMC and reflector, the minimum distance between FSS and antenna is kept to λ/2. Also, the ordinary superstrates are placed over the antennas through a half-wave (λ/2) spacer. This distance is large sometimes, and the whole setup takes much more space.

Metamaterial FSS: Introduction and How It Combats the Limitation of Traditional FSS Structure

In modern days, the word "metamaterial" is being used in designing of many components like antenna, filters, feeding lines, etc. It will be fair to mention that metamaterials are not technology itself, but a technology-enabling design philosophy. The word metamaterial is a combination of the Greek word meta, meaning "beyond," and the Latin word material, meaning "material or matter." So metamaterial would mean beyond naturally occurring materials (Caloz and Itoh 2005; Cui et al. 2010). According to the definition available in the literature:

"Electromagnetic metamaterials are artificial effectively homogeneous electromagnetic structures with unusual properties not readily available in nature" (Caloz and Itoh 2005).

Material Constitutive Parameters

A material is mainly characterized by macroscopic parameters, i.e., electric permittivity (ε) and magnetic permeability (μ). This classification of material at the macroscopic properties level decides the wave-matter interaction. The electric permittivity of a material (ε) defines its ability to store the electrical energy when an electric field is applied. Similarly, the magnetic permeability of any material depicts its ability to get magnetized or store magnetic energy in response to an applied magnetic field. For free space, the permittivity and the permeability are denoted by *so* and uo, respectively, where $\varepsilon o = 8{:}858 \times 10^{-12}$ F/m. and $\mu_o = 4\ \mu \times 10^{-7}$ H/m. The permittivity, and permeability of all other materials, is determined relative to the free space permittivity and permeability, respectively. The relative permittivity (ε_r) and the relative permeability (ur) of any material are defined as $\varepsilon_r = \varepsilon/\varepsilon_0$ and $\mu_r = \mu/\mu o$, respectively. Moreover, the refractive index of a material is given by $n = \sqrt{\varepsilon_r \mu_r}$ which defines by which factor the wave velocity in any medium slows down with respect to free space.

Resonant and Nonresonant Metamaterial

In general terms, any material is identified by its cellular architecture (size) and its constituting material. Similarly, for metamaterials also, its cellular architecture and constituting parameters decide its extraordinary properties. An element (unit cell) whose structural average cell size (p) is much smaller than the guided wavelength of operation (λ_g) is called an effectively homogeneous electromagnetic structure in terms of effective permittivity and permeability. To have a medium, in which phenomena of refraction dominate over diffraction or scattering, it is recommended that the medium obeys the $p = \lambda_g/4$ condition, and waves see a uniform medium in the incident direction. Thus, the medium is electromagnetically uniform along the propagation direction (Caloz and Itoh 2005). However, the properties are also dependent on the type of substrate used for metallization, and therefore, it must be chosen carefully.

As mentioned before, the constituent materials have an important role in determining the behavior of the metamaterial structure; more specifically, the unit cell must be carefully chosen to achieve a desired effective permittivity and permeability. These unit cell elements can either be resonant or nonresonant, and both have their own advantages and disadvantages.

Due to the high-quality factor of resonant structures, the frequency response is narrowband. In these narrowband regions, the impedance of the structure is matched

to the impedance of the incident medium. Changes in the material parameters cause a shift in the resonance frequency very sharply because the surface impedance is a function of frequency (f), permittivity (ε), and permeability (μ) (Caloz and Itoh 2005). This means that any change in the geometric size of the resonating element will shift the resonance frequency, and this, in turn, will cause changes in the ε and μ values. The benefit of such a behavior is that large dynamic-range material parameters can be realized. However, they are very narrowband in nature and provide high attenuation beyond the narrow band.

In nonresonant metamaterials, the effective permittivity and permeability change slowly over a wide band, and so they provide wide band response. The fact is that it is not true that the nonresonant metamaterials do not have a resonant frequency, but it is much higher than the frequencies a particular structure is designed for. This way, nonresonant metamaterials operate at low losses and can be very useful in wave shaping, reducing the wavelength, and miniaturization.

Metamaterial Frequency Selective Surfaces

Similar to metamaterials, metamaterial FSS are not only a technology, but also a technology-enabling design philosophy according to which some of the limitations of conventional FSS are trying to be overcome by cellular architecture and constitutive parameters of metamaterials structures. One of the main virtues of metamaterial structure is that it enables wave-engineering. Using the metamaterials concept, the elements of the FSS are designed much smaller than the operating wavelength ($<\lambda/4$) so that the wave does not have prominent scattering and diffraction. However, it sees an effective medium along the direction of propagation. This way, refraction is a more effective phenomenon rather than scattering and diffraction. Violation of traditional rule of thumb in FSS design, i.e., using resonant length elements ($\lambda/2$), by metamaterial concept where element size is kept designed much smaller than the operating wavelength ($<\lambda/4$), features very interesting behaviors. This unique feature of the new design is its localized frequency-selective property. It enables miniaturizing of the size of the resonating element and inter-element spacing and so provides large-angle stabilization, reduces the grating lobes, and eliminates the problem of harmonics. Wideband behavior can be obtained by using multiple elements with overlapping individual resonance frequencies when combined. It suppresses harmonics of the intended frequency response. Moreover, by designing the unit cell appropriately, very thin FSS layers, with high-order frequency responses and very low sensitivity with respect to the incidence angle of the wave, are achieved. Different metamaterial FSS structures have been developed to exploit its characteristics (Xiong et al. 2013; Gu et al. 2013; Landy et al. 2008; Bayatpur and Sarabandi 2009; Narayan et al., 2018; Ghosh and Srivastava 2015; Cheng et al. 2020; Pirhadi et al. 2012; Narayan and Jha 2005).

Terahertz Metamaterial FSS for Future Wireless Communication Systems

For achieving an ultrahigh data rate, THz communication is crucial, so significant research contributions exist in THz communication. However, the major challenge in the direction of THz wireless communication is that its high propagation loss significantly reduces the communication link coverage. To successfully recover the information, we require high antenna gain, a simple THz chain hardware system with simple architecture with low loss, cross-coupling, and energy-efficient components. In this direction, the role of metamaterial FSS is crucial (Chen et al. 2019; Tang et al. 2020; Sabah et al. 2014; Rikkinen et al. 2020; Rashid and Zhang 2019; Sanphuang et al. 2014). For example, ultramassive multiple-inputs and multiple-outputs antenna systems are being sought as a potential solution to combat the free space loss and enhance the coverage area by enhancing the antenna gain and spatial diversity. However, it requires a complex and costly hardware structure with n number of THz hardware chain. In this direction, using compact THz metamaterial FSS structure before the UM-MIMO system, filter component in all hardware chain can be completely eliminated which significantly reduces the hardware cost, loss, and complexity similar to as discussed in (Tang et al. 2020). It has been illustrated in Fig. 15. Also, to avoid the cross-coupling between different antenna elements in the UM-MIMO system, shielding can be provided by the THz metamaterial FSS.

In addition to this, the metamaterial concept enables the use of frequency selective surfaces on conformal surfaces with a stable angular response, which will be very useful for THz beamsteering antenna design, redirecting the EM wave for smart indoor home applications. In ordinary superstrates, the conventional FSS designs are placed over the antennas through a half-wave spacer. The advantage of using the new metamaterial FSS-based superstrate, as opposed to conventional FSS, is that this metamaterial FSS can be positioned at very close distance ($\lambda_g/10$) to the antenna for providing a uniform wave front if the unit cell dimension is $\lambda_g/10$. By placing this THz metamaterial FSS very close to the antenna, the miniaturized FSS structures have a minor effect on the radiation performance of the antenna. This way, the new achievement is the practical demonstration of thin-layer FSS-antenna structures. As a concluding remark, the metamaterial concept in FSS allows to miniaturize the element size and the inter-element spacing and to design thin FSS structures. So, metamaterial FSS has many advantages and applications in THz communication, such as radome, filtering, absorber, and antenna with a wide range of crucial characteristics such as wideband excellent stable angular response, delayed grating lobes, and suppressed harmonics.

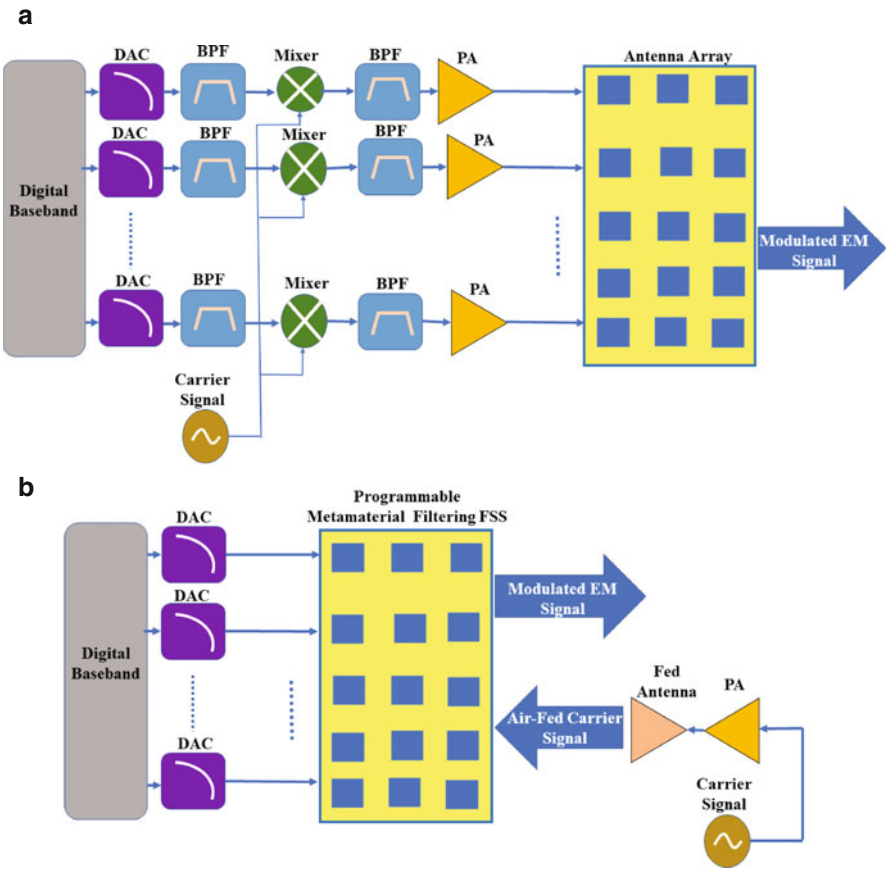

Fig. 15 Comparison of the (**a**) conventional wireless transmitter and the (**b**) programmable filtering FSS chain-free wireless transmitter

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Metamaterial Enabled FSS for Beam-Tilting MM-Wave Antenna Applications

15

Mehri Borhani-Kakhki and Tayeb A. Denidni

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Abstract

In the fifth generation (5G) of wireless communications, millimeter-wave frequency band is an attractive technology due to the ability of offering multi-Gb/s transmission rates and supporting higher number of users due to its wide bandwidth. In fact, 5G is going to commercially support the growing demands for high capacity, low latency, and widespread mobile access, which has an important role in connecting and enabling services. In mm-wave wireless communications, beam-tilting antennas as a smart antenna have recently received extensive attention due to their ability to restrict the radiated power to a predefined zone rather than whole directions. Therefore, the system coverage is greatly improved and the rate of interference is reduced. Different beam-tilting methods with various

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strategies have been introduced in the literature. However, considering advantages and disadvantages of the conventional methods, an alternative solution is to achieve pattern reconfigurability using metamaterial which offers less loss, more antenna functionality, and less cost beside a considerable save in terms of space and size. Thus nowadays, this field is becoming one of the most popular subjects of study. Among the metamaterials, Frequency Selective Surfaces (FSSs) are the most famous one for beam-tilting applications. In fact, Frequency Selective Surfaces work as a filter and are able to transmit or reflect electromagnetic (EM) waves in the frequency band of operation. In this chapter, we will go over introducing available beam-tilting techniques with the latest advancements in beam-tilting antenna system designs based on Frequency Selective Surfaces for wireless communications and will highlight various challenges that can be encountered.

Introduction

Wireless communication systems, such as mobile communications, have evolved dramatically from their humble beginnings. In wireless technologies, antennas are a key component which have been used for various applications to point-to-point and mobile cellular communications and also for enabling high-data rate transmission. In the spectral domain, in order to increase the data rates and also to boost the available capacity, millimeter-wave frequencies which are in the range of 30–300 GHz, so having 1–10 mm wavelengths, are attracting attentions. In mm-wave frequencies, in addition to the generous bandwidth, much smaller RF components including antennas are offered compared to lower frequencies. Therefore, due to the compact size, many of the components can be packed together into small form factors, in which passive, active components and antennas can be integrated in the same package (Borhani Kakhki et al. 2018; Pucci 2013). However, compared to the signals in lower-frequency which have longer wavelengths, wireless systems in the mm-wave frequency suffer from a significant loss owing to atmospheric absorption. In fact, due to the short wavelength of the signals at 24 GHz and above, they can be easily absorbed by the located objects in their propagating path, for example, trees, rain, buildings, etc., which can highly reduce the communication link quality. To solve this issue, high-gain antennas have been offered; however, high-gain antennas suffer from particular difficulties in beam alignment due to their narrow beam-width, especially when there is a signal blockage between the receiver and transmitter. Therefore, beam-tilting and beam-switching antennas have been introduced as an attractive solution, in which the antenna's beam can be bounded in an appropriate zone rather than transmitting in whole areas. This technology reduces the effect of interference, increases the signal to noise ratio (SNR), and improves the system coverage. Antenna's beam-tilting can be accomplished with different controlling methods such as mechanical, electrical, using metamaterial, etc. In electrically beam-tilting method, active components such as PIN-diodes, RF-MEMS switches, and Varactor diodes are implemented inside the antenna's structure for pattern

reconfigurability. Although using this technique the radiated power can be guided toward the desired angle, the gain drops when the beam is tilted (Kim et al. 2006; Romanofsky 2007; Preston et al. 1997; Sharifi et al. 2017; Borhani Kakhki and Rezaei 2017; Borhani Kakhki et al. 2015, 2016). The limited scan angle is another drawback of electronic beam steering.

Steering antenna's beam by mechanical approach is pretty easy, and all we need is just to install the antenna on a rotating structure or physically changing the position of the antenna element to reach the desired characteristics. It causes a fixed angle beam with no reduction in the antenna gain as well as advantage of having a wider scan angle. However, beside the low speed of this technique, implementation and installation of the mechanical approach in base-stations is very complex and costly compared to the electrical approach.

Integration of phase shifter is another method used to steer the main beam direction electronically, to change progressively the phase of each element. In this technique, to adjust the phase of each element, a linear or multidimensional array of antennas are controlled using a feed network. Therefore, direction of the radiated EM waves from the antenna can be controlled toward the predefined angle. Although the phased array antennas are not facing some of the mentioned difficulties compared to other electrical methods, with this approach the antenna's gain is extremely reduced. Moreover, the complex feeding network of the phased antenna array and phase shifters results to complex and high costs design processes and involves lots of loss.

To deal with the electrical and mechanical beam-steering challenges, an alternative method is using metamaterial which recently has drawn researcher's attention. In the metamaterial technology, the user is able to produce custom dielectrics that cannot be found in the nature or they are difficult to be reached using traditional dielectrics. Implementing a beam-tilting structure using metamaterial technology improves the performance of antenna in terms of power consumption, loss, speed, and freedom degree for design at mm-wave frequencies.

Metamaterial are essentially consisting of a dense array of periodic metallic structures embedded in a host medium which show exotic properties. The basic parameters of metamaterial are relative permeability known with μ_r and relative permittivity known with ε_r. Metamaterials are categorized in three groups (based on these two parameters): 1- Double Positive (DPS), where relative permittivity and permeability both are positive; 2-Single Negative (SNG), in which either μ_r or ε_r is negative but not simultaneously; and 3-Double Negative (DNG), where both μ_r and ε_r are negative. Characterization method of metamaterials involves calculation of the reflection and transmission coefficients.

In fact, metamaterials can be realized with different structures depending on the way that they behave with the incident EM waves. Artificial Magnetic Conductor (AMC), Electromagnetic Bandgap structures (EBG), High Impedance Surface (HIS), Split Ring Resonators (SRR), Frequency Selective Surfaces (FSS), etc. are the most famous kinds of metamaterials (Borhani Kakhki 2020).

FSSs are planar metamaterials made of periodic array of metallic apertures or patches with various shapes for frequency selectiveness with transmission/reflection characteristics. Generally speaking, the element shape, size, and periodicity of an

FSS are important parameters which affect the resonance (Munk 2000). Frequency Selective Surfaces have been studied extensively over the decades and were frequently used in reflector antennas, radio frequency identification (RFID), dichroic subreflectors, lenses antennas, and protection from electromagnetic interference. Antenna radomes and controlling radar cross-section (RCS) are one of the most famous applications of FSSs (Munk 2000).

With the development of metamaterial as engineered EM materials, FSS has received considerable attention for the radiation pattern-sweeping applications. Moreover, in Frequency Selective Surfaces, the resonance properties can be varied when implemented by active components such as PIN diodes (Selvam et al. 2017), varactor diodes, (Luther et al. 2012) and MEMS (Microelectromechanical Systems) switches (Cheng et al. 2009). Low profile, dual polarization, reduced periodicity, angular stability, and easy manufacturability are some of the FSSs' desirable properties.

In this chapter, we provide an overview on Frequency Selective Surfaces and its applications in beam-tilting antenna systems and highlight some of their challenges particularly for mm-wave frequency bands.

Frequency Selective Surfaces

In the 1960s, Frequency Selective Surfaces have been broadly used in the military applications. Even though, Marconi and Franklin in 1919 have formulated the initial sample of the FSS, due to its high interest for military applications, the further intense experimental and theoretical investigations of FSS have been reported after the 1960s.

Frequency Selective Surfaces (FSSs) are fixed frequency metamaterial where spacing and static geometries of the unit cell specify the frequency response of a metamaterial. FSSs are known as a two- or three-dimensional periodic structure, composed of an infinite array of metal apertures or patches in a metal sheet printed on a dielectric substrate. FSSs have basically been introduced as a spatial filter in which an incoming EM wave will be either reflected or transmitted, partially or completely over a specific frequency band depending on the array element design. To do so, the frequency of the plane wave should be matched with the resonance frequency of the FSS unit cell.

Basically, the FSS can be excited by an incident plane wave of E^i, which will be partly reflected as E^r in the specular direction, and partly transmitted in the forward direction as E^t, as shown in Fig. 1 (Li 2015).

The reflection and transmission coefficients can be defined as $\Gamma=\frac{E^r}{E^i}$ and $\tau=\frac{E^t}{E^i}$, respectively. Energy conservation for a lossless network simply needs for the transmission and reflection coefficient; thus, $\Gamma^2 + \tau^2 = 1$. Therefore, it is clear that for a high amount of transmission coefficient, the reflection coefficient reduces automatically (Li 2015). A good explanation of FSS unit cell design and boundary condition assignment has been introduced at (Frequency Selective Surfaces n.d.; Anusha et al. 2019).

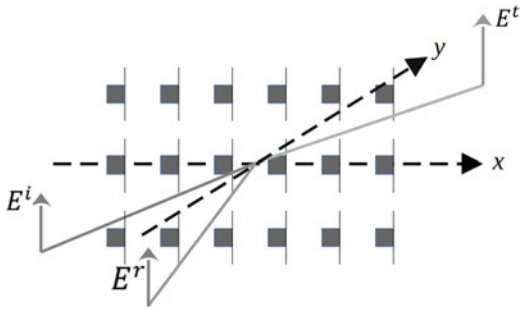


Fig. 1 FSS excited by plane wave

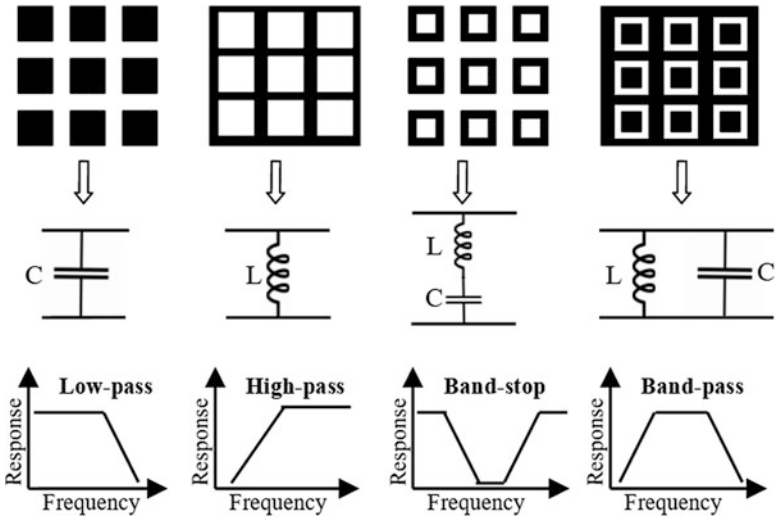


Fig. 2 FSS periodic structures, their equivalent circuits, and the frequency responses (top to bottom)

According to the filter characteristics, FSSs are categorized in four types as low-pass, high-pass, band-stop, and band-pass where their basic structures, equivalent circuits, and their filters responses are demonstrated in Fig. 2 (Volakis 2008). In this picture, the black color demonstrates the metal material. From Fig. 2, the patches provide low-pass filter characteristics in which lower frequency range is allowed to pass through the structure, and higher frequency range is bypassed. High-pass filter characteristics on the EM field can be formed by metal grid while its operation is a counterpart of low-pass filter function by applying Babinet principle. The metal patch has band-stop-filtering characteristics and blocks undesired frequencies, while the apertures have band-pass characteristics and allow only specific frequency range. In summary, the patch element group exhibits a capacitive response and consists of both low-pass FSSs and band-stop, whereas high-pass and band-pass FSSs belong to the aperture (grid) or slot group element that shows an inductive response.

Physically, FSS patches generate inductance (L) and resistance (R), while capacitance (C) is constructed by the gaps among the FSS elements. Therefore, a filter response is created using these inductive and capacitive elements combination. Moreover, any variation in the dimensional parameters of FSS results in an equivalent change in the C or L values. For example, elements such as dipole, loop, and cross which are known as single resonant elements can be modeled by a series combination of inductor and capacitor whereas just a capacitance can represent nonresonant elements such as wire grid and patch. It should be noted that the number of resonances is proportional to the number of lumped elements. In fact, when an FSS unit cell is incited by the electromagnetic wave, the structure can be modeled by an equivalent resonance circuit with the frequency resonance of $f_r = \frac{1}{2\pi\sqrt{LC}}$ in which C and L are the equivalent capacitance and inductance of an FSS cell, respectively (Anwar et al. 2018).

Low profile, dual polarization, reduced periodicity, easy manufacturability, and angular stability are some of the desirable features of Frequency Selective Surfaces. Nevertheless, for the FSSs designers, it has been challenging to achieve all the abovementioned characteristics. In design of an FSS unit cell, several metrics should be considered to be able to obtain the anticipated performances. Although the details of such features have been discussed in the previously published works such as (Anwar et al. 2018; Li 2018), in the following section we tried to list them briefly:

Standard FSS Geometries

Generally, different shapes of FSS elements have been broadly classified into four groups as shown in Fig. 3. The first group consists of centrally connected structures such as dipole and Jerusalem crosses. The size of the FSS element for this category equals to half of the wavelength. Looped elements such as circular, square, hexagonal loops come under the second group, which show better resistance for the oblique incidence of the waves. Moreover, the broad range of bandwidth variation and compact size are the benefits of these group elements. In the loop elements, FSS

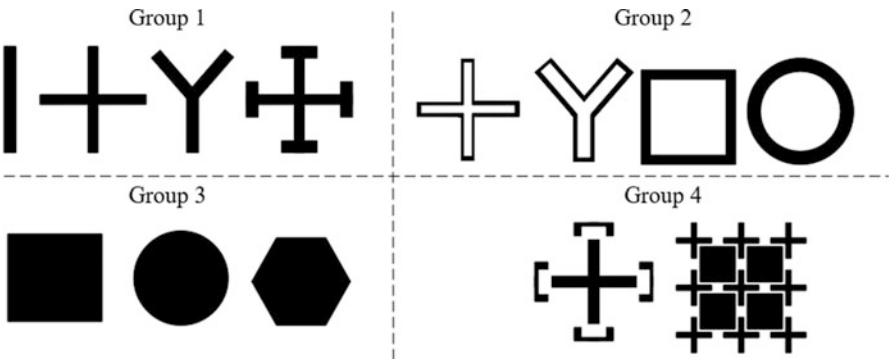


Fig. 3 Typical shapes of FSS elements

reaches the resonance when the loop circumference is equal to a wavelength. Moreover, by changing the shape of the loops, a different range of bandwidth can be achieved. For example, narrow band structures can be reached by three or four-legged elements, while ultrawideband structures are modeled using hexagon loops. The third group consists of solid interior elements or patch shapes of different shapes, which are desired for sensitive applications to the oblique incidence and grating lobes onset. In the fourth group, in fact the other three group elements are combined, so specific characteristic of each one can be used for a desired response (Li 2018; Niroo-Jazi 2012). It should be noted that the angular stability and polarization stability of any unit cell improves with the symmetry of the unit cell.

Effect of Inter-element Spacing

The inter-element spacing is an important factor in the FSS bandwidth determination. The inter-element spacing can be shortened by selecting proper FSS elements, and therefore the FSS bandwidth can be enhanced. It should be mentioned that early onset of the grating lobes can be caused if the inter-element spacing is more than half of the wavelength. The bandwidth increases when the inter-element spacing is reduced. As a rule of thumb, there is a smaller inter-element spacing in the loop element compared to the center-connected element which delays the onset of grating lobes and increases the bandwidth. Therefore, the loop element is preferred in comparison with the center-connected element (Xua et al. 2015).

In fact, the periodicity of FSS elements can change the frequency of resonant and the phase of an FSS array. In most of the elements, varying the inter-element spacing does not significantly change the frequency of resonant. Elements such as slot FSS with high-end capacitance demonstrate large shift in frequency of resonance for any change in the inter-element spacing. In these elements, by increasing the periodicity of the elements while the slot size is kept constant, the resonant frequency can be moved to higher frequencies. On the other side, with increasing the inter-element spacings, the mutual coupling decreases. But it should be noted that the inter-element spacing cannot be increased to any extent because it will be resulted in the grating lobes onset which could be as strong to reduce the original frequency of resonance. The effect of the inter-element spacing reduction on the impedance bandwidth enhancement can be explained considering the fact that the impedance of an element in an array is lower than the identical element isolated in free space. This can be also justified by the fact that, although a single element can store charge between its edges to infinity, the same element located in an array can only store charge between the edge of the element to half the distance to its closest neighbor. This decreases the capacity in which charge can be stored and thus reduces the amount of the stored charge. The magnitude of reflection amplitude does not rise when the inter-element spacing is reduced, but the real part of mutual impedance increases, resulting in higher bandwidth. The imaginary part is decreased which leads to the resonance frequency stability with angle of incidence. In order to get a stable resonance with angle of incident, a smaller inter-element spacing can be used which simply needs use of small

elements. Although increasing the periodicity is not so advantageous, however, the small variation in periodicity might result in different phase value which is desirable in some applications such as millimeter wave diffuser system (Islam et al. 2010).

In (Islam et al. 2010), the authors showed the influence of variation of inter-element spacing on the antenna arrays reflection characteristics by considering grounded FSS arrays as the basic element. They have studied different grounded FSS arrays with various inter-element spacings of shorter or larger than half a wavelength and presented good understanding on magnitudes variation both in phase and amplitude when the inter-element spacing varies. As a result, the arrays show higher bandwidth when the inter-element spacing is less than half a wavelength compared to the array of inter-element spacing of λ/2. Moreover, it is showed that to realize phase delay, the inter-element spacing can be employed as a parameter for the phase optimization in the first-order mode as long as they are less than a wavelength (Anwar et al. 2018; Islam et al. 2010).

FSS-Dielectric Substrates

Due to the mechanical reasons, Frequency Selective Surfaces are usually printed on dielectric slabs since the dielectric supports the FSS and contributes in the roll-off rate, incident angle performance, and flatness of bandwidth. The frequency response of an FSS unit cell etched on dielectric substrate is reduced by a rate of $\frac{1}{\sqrt{\varepsilon_r}}$, in which ε_r stands for dielectric substrate-relative permittivity. Furthermore, the thickness of the dielectric substrate affects the polarization and the angular stability of the resonant FSS. In the articles, multiple FSSs layers with dielectric cladding are used for better wideband performance (Kesavan 2019). Therefore, embedding the FSS with dielectric on both sides of dielectric or bonding the FSS in one side extremely changes its transmission responses and can improve the angle stability.

Beam-Tilting Antenna Applications

There is a wide collection of beam-tilting antenna systems implemented by different types of antennas for a broad range of applications and large range of frequencies. It has found numerous applications in sonar, radar, seismology, radio astronomy, wireless communications, biomedicine, and acoustics. In this section, we aimed to preset some demonstrative examples of the beam-tilting antenna applications.

Base Station Antenna

In the base station antenna, the system capacity and interference immunity can be improved by tilting the antenna's radiation pattern down and toward a preferred angle. In the base station, antennas mostly having a fixed main beam toward the down-tilt angle are preferred which results in the receive signal improvement in a mobile device as shown in Fig. 4.

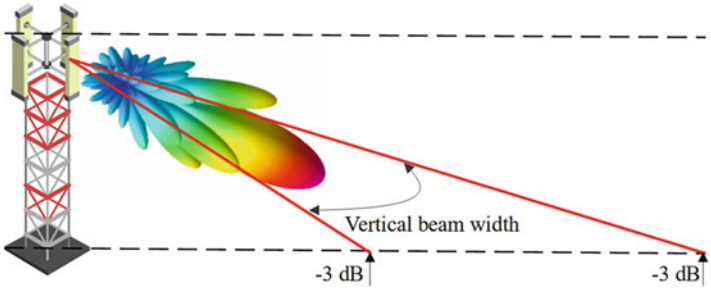


Fig. 4 Down-tilt radiation pattern of a base station antenna

Tilting the antenna's beam in the base stations can be reached using antenna array in which the position of the array is mechanically changed or phase of each element is electronically controlled. Another alternative to achieve higher performance is based on metamaterial structures since it is suitable for multiband, low-profile applications and is able to steer radiation pattern toward a preferred direction. Moreover, to serve different users, various frequency beams can also be steered toward different directions. To effectively track the user, the base station dynamically calculates the direction that a signal is sent in, as the end point moves. Switching between the various fixed beams takes place based on the user location. In fact, the end point will switch to a different beam if a beam cannot track a user. The application of adaptive beam-tilting at the base stations promises a significant increase in the network capacity of the cellular networks (Kim and Rahmat-Samii 2015).

Routers

In very simplified terms, beam-tilting is all about focusing a signal in a specific direction. Traditionally, when the router broadcasts a signal, it broadcasts the data in all directions. With beam-tilting, the router projects a stronger signal in a specific direction where the device, e.g., laptop, smartphone, tablet, or whatever else, is located as illustrated in Fig. 5. Beam-tilting promises a faster, stronger signal with longer range and more secure data transfer for each device. Instead of simply broadcasting the signal in whole zones, the router tries to broadcast wireless data in a way which is optimal for the device.

Military Applications

In the 1960s, adaptive beam-tilting was initially developed for the military applications of radar and sonar to mitigate the signal jamming effect in the military domain. Beam-tilting can be used in both the receiving and transmitting ends to reach spatial selectivity. The radar applications make use of either static or dynamic beam-tilting (Fig. 6). Directional beam-tilting antennas transmit a signal toward a specific receiving device, instead of spreading the signal in whole directions, as they

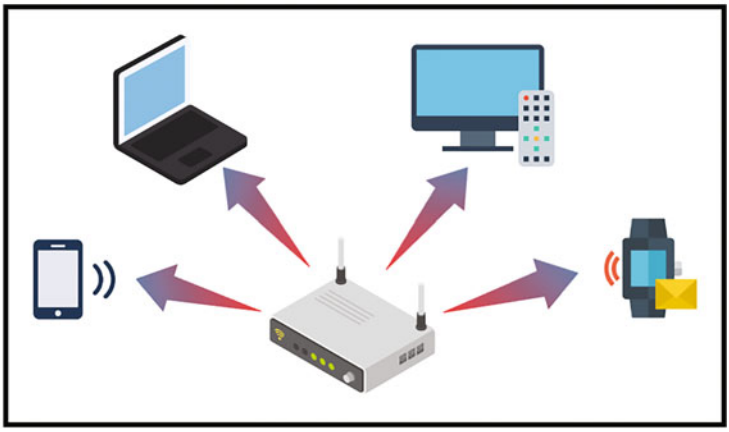


Fig. 5 Schematic of a beam-tilting router

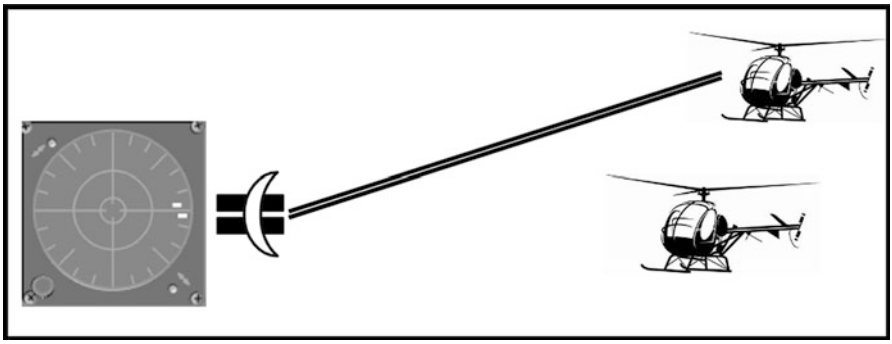


Fig. 6 Schematic of a beam-tilting radar

normally would. In directional beam-tilting systems, due to the need for multiple large antenna elements, the cost is high. However, millimeter wave 5G technology is attracting much more attention to reduce the weight, size, and cost of directional beam-tilting systems, due to its smaller size and high-frequency output, which permits the systems to be integrated easier to army platforms (Fig. 6).

FSS Solutions for Beam-Tilting Antennas

FSSs have extensively been investigated over the years and were frequently used in a variety of applications. The shielding on the door of a microwave oven is an example of FSS applications which permits us to see the food inside the oven without being affected by the radiated EM waves. In fact, the printed Frequency Selective Surfaces on the door of oven work as a filter and reflect microwave energy. Furthermore, FSSs

have been broadly used in civil applications such as robotic navigation systems, electromagnetic shielding structures in airports and hospitals, anticollision systems for autonomous vehicles, etc. FSSs usage in the military domain as radome is another important application which is able to reduce the RCS or radar cross section of antennas and keep them hidden from the enemy.

Recently, the growing needs for multifunctionality in the same design platform resulted in significant development on reconfigurable FSSs in wireless systems for beam-tilting applications. Implementing reconfigurable FSSs can be done using tuning and tunable elements like PIN diodes, microelectromechanical system (MEMS) switches, varactor diodes, or MEMS capacitors (Li 2015). Moreover, due to the unique properties of FSSs, they have been also widely used in beam-tilting applications without active components. In this regard, some examples of FSS-based beam-tilting antennas techniques that have been recently reported in the literature are introduced in this section along with their advantages and disadvantages.

Beam-Tilting Antennas Based on Passive FSS

Frequency Selective Surfaces (FSSs) have been used broadly to tilt the antenna beam toward the desired directions when it works to transmit (pass-band) or reflect (stop-band) electromagnetic wave. For example, Borhani et al. in (Borhani Kakhki et al. 2018) introduced a design of down tilted beam antenna based on multilayer FSSs with gain enhancement for mm-wave 5G applications. In the proposed FSS unit cell design, two C-shaped strips are printed on the top layer while a slotted circular patch is placed on the bottom layer of a dual-sided element with a total size of $0.46\lambda \times 0.46\lambda$ at 28 GHz, as shown in Fig. 7.

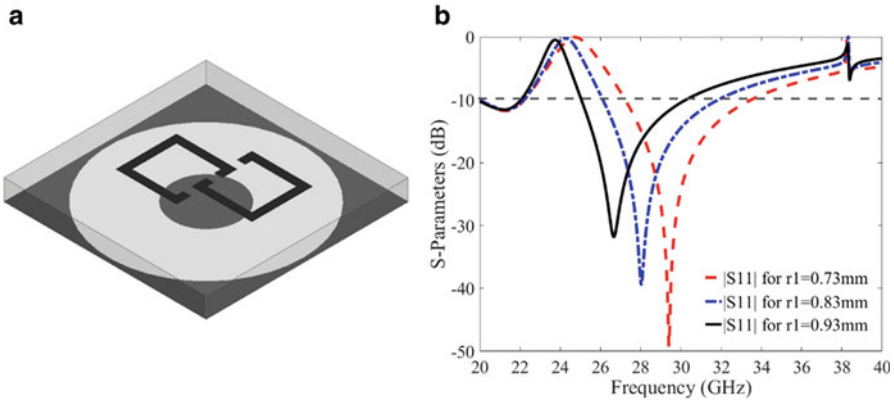


Fig. 7 (**a**) 3D schematic of the FSS unit cell and (**b**) the effect of radius variation of small circular patch located in the back plane of the FSS on the frequency response (© 2018 IEEE, Reprinted with Permission)

The scattering parameters of the element are optimized in which at 28 GHz the surface operates on its transmission mode when an incident wave spatially excites the FSS unit cell, while it shows reflection characteristics out of the band. As shown in Fig. 7b, increasing the radius of the smaller circular patch of radius "r1" in the back plane of the unit cell shifts the frequency resonance of the unit cell toward lower frequencies. In fact, the conductors inside and surrounding the loop-shaped slot, in the back plane of FSS, behave as a capacitor. When the radius of the smaller circular patch is increased, these conductors come closer and thus the equivalent capacitance is increased. Therefore, with the capacitance enhancement, the frequency of resonance goes down. In this work, r1 is chosen to be 0.83 mm; thus, the FSS shows a resonance frequency at 28 GHz.

To excite the FSS layers, a wideband Vivaldi antenna having an end-fire radiation pattern is used. To tilt the main antenna's beam, bellow the horizon, two unequal-size FSS layers as transmitter with 45° rotation have been located under the Vivaldi antenna as demonstrated in Fig. 8. It resulted in a maximum beam tilt angle of 38° at 28 GHz as shown in Fig. 9. Moreover, locating FSS in the near fields of the Vivaldi antenna not only does not affect the reflection coefficient, but also up to 3 dBi gain enhancement is seen due to the FSS loading. The electrical field distribution of the Vivaldi antenna with and without FSS loading is shown in Fig. 10a and b, respectively. It can be seen that when the radiator is loaded with FSS layers, the direction of electric field is deflected toward the FSS superstrates and induces current in the FSS unit cells. Based on this phenomenon, the FSS acts as a lens and attracts the radiation beam of the antenna and improves the gain due to constructive interference of the reflected field.

To validate the beam-tilting mechanism, Snell's law can be also used and applied on the boundary of the air interface and the layers as shown in Fig. 11a. According to Snell's law:

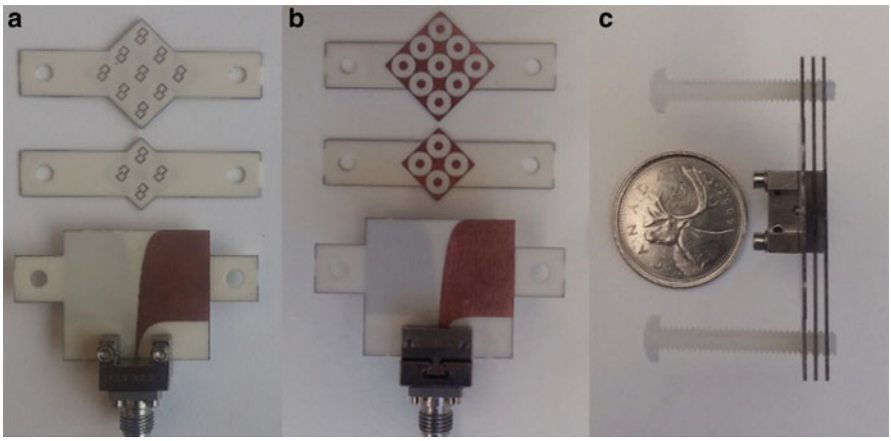


Fig. 8 Antenna's photograph from (**a**) top, (**b**) bottom, and (**c**) side views (© 2018 IEEE, Reprinted with Permission)

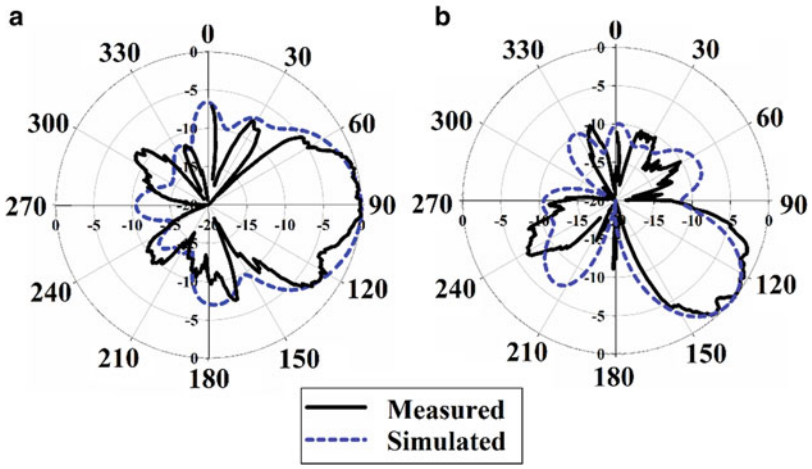


Fig. 9 Vivaldi antenna's normalized radiation patterns in the H-plane (**a**) without and (**b**) with FSS layers at 28 GHz (© 2018 IEEE, Reprinted with Permission)

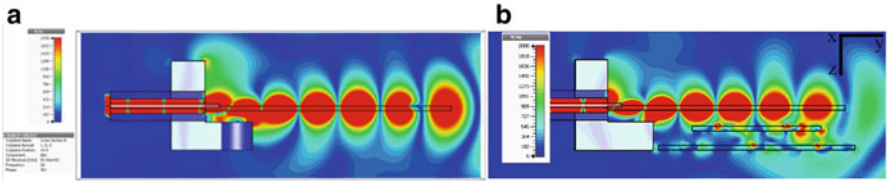


Fig. 10 E-field distribution over the antenna (**a**) without and (**b**) with FSS layers at 28 GHz (© 2018 IEEE, Reprinted with Permission)

$$n_{air}.sin\theta_1 = n_{FSS}.sin\theta_2 \tag{1}$$

where n_{air} is the refractive indexes of the air, n_{FSS} is the refractive indexes of the FSS unit cell, and θ_1, θ_2, and θ_3 are the waves' incident angles. According to Fig. 11b, the designed FSS unit cell has a refractive index of 1.9 at 28 GHz.

It is shown in Fig. 11a that the emitted wave from the feed antenna element goes along the length $R = \sqrt{(L^2 + d^2)}$, while θ_1 can be calculated by d and L. Based on the n_{FSS} value presented in Fig. 11b, the final optimized parameters of d and L, and the refractive indexes average of multilayer FSSs, the calculated reflected angle by Eq. (1) is found to be 38.3° at 28 GHz, which agrees with the simulated beam tilt angles.

The same author in (Borhani Kakhki et al. 2020) introduced a technique of beam deflection on Dielectric Resonator Antennas (DRA) fed through Printed Ridge Gap Waveguide (PRGW) based on multilayer FSS at 28 GHz. The introduced FSS unit cell in this work operates on its reflection mode at 28 GHz when an incident wave spatially excites the element as shown in Fig. 12. The loop FSS is a structure with

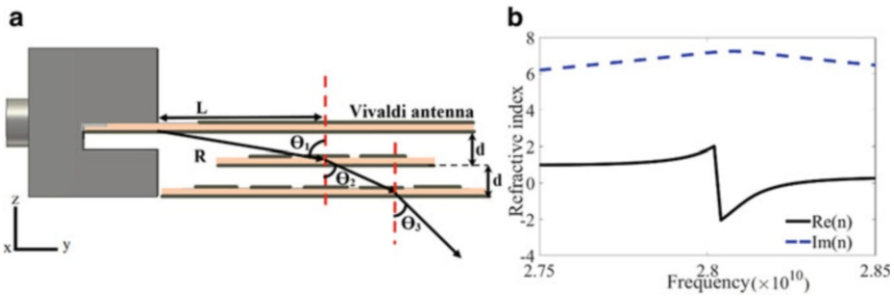


Fig. 11 (**a**) The beam-tilting mechanism of the antenna and (**b**) refractive-index of the FSS unit cell (© 2018 IEEE, Reprinted with Permission)

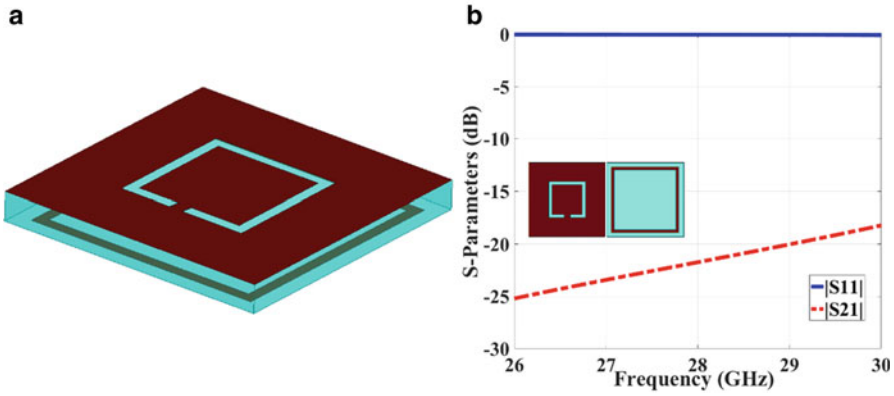


Fig. 12 (**a**) 3D view and (**b**) simulated S-parameters of the FSS unit cell

polarization independency which supports vertical, horizontal, and circular polarizations. In fact, in this unit cell, for the FSS to be opaque at 28 GHz, a perfectly conducting thin strip is located in the square slotted patch in the top plane where the E-field is maximum, to eliminate the capacitive effect between the pertinent conductors. Hence, as shown in Fig. 12a by shorting the adjacent conductors using a strip in the y-direction, the capacitance is canceled out. The thin conducting strips are named "polarization cancelers" as by removing the capacitance between the vertical conductors, they cancel out the vertical polarization component of the incident electric field. Thus, the square slotted loop on the top plane of the designed FSS is changed to resonate just for horizontally polarized waves and reject vertically polarized waves. Figure 12b shows the S-parameters which demonstrate reflection behavior in the Ka-band.

In this work, beam-tilting is obtained by fixing two FSS layers, and each consists of 3 × 3 FSS inclusion, over and offset from the centers of two DRA antennas as

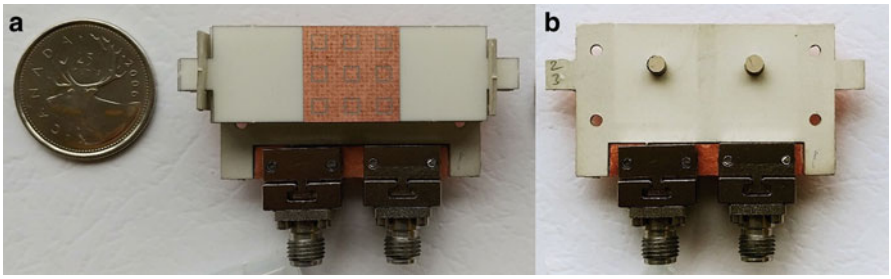


Fig. 13 (**a**) Photograph of the fabricated double fed DRA antenna (**a**) with two FSS layers and (**b**) without FSS layers

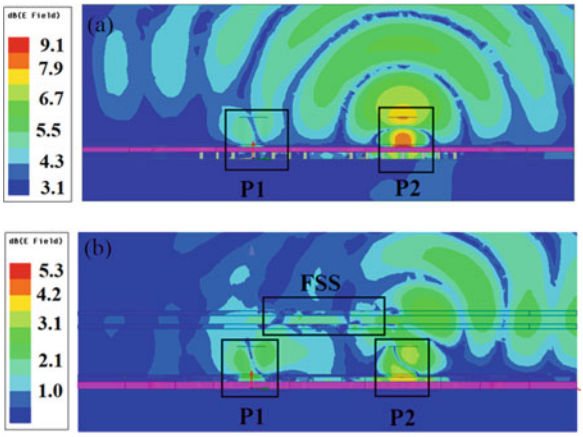


Fig. 14 E-field distribution when port 2 is ON (**a**) without and (**b**) with FSS loading at 28 GHz

illustrated in Fig. 13. The FSS unit cells are designed to work on their reflection mode. The DRA antenna is excited with a slot at the center, hence the fundamental mode HEM11δ is excited. Turning ON each port while the other port is kept OFF resulted in ±77° beam-switching at 28 GHz while a 3.16 dBi gain enhancement is also achieved due to the FSS loading compared to the unloaded case. In fact, the designed FSS can redirect the main beam of the DRA toward the opposite direction of the FSSs. The E-field distribution when port 2 is ON at 28 GHz is showed in Fig. 14. It is clear that the proposed FSS structure is able to deflect the DRA main beam to the opposite direction of the FSSs.

Using the proposed techniques, not only the antenna beam can be tilted but also low-profile, compact, and low-cost structures are reached. Moreover, due to FSS loading the gain is increased which is opposite to the phased array antenna in which the gain will be dropped when the beam is tilted. However, once the passive FSS is fabricated, its EM characteristics such as the working bandwidth and the resonance frequency cannot be modified while the active FSSs can solve this issue.

Beam-Tilting Antennas Based on Active FSS

Active frequency selective surface is also called reconfigurable FSS, controllable FSS, and tunable FSS. According to a variety of principles, active FSSs are realized broadly using electronically controlled materials and elements in the microwave field. Tunability permits the FSS unit cell to compensate for the best performance in terms of bandwidth, directivity, and gain in order to adapt to spectral changes. Numerous Frequency Selective Surfaces designs have been reported in the literature which are incorporated with the active components including PIN diode switches, RF-MEMS switches, and Varactor diodes, as well as cantilever switches to be able to electronically control the FSS characteristics. However, the huge use of active components in FSS for beam-tilting applications will increase the cost, complexity, and the potential for failure (Li 2015). In this section, some of the reconfigurable FSS structures presented in the literature are briefly introduced.

Reconfigurable FSS-Incorporated RF-MEMS Switches

Microelectromechanical switches (MEMS) are a new type of RF devices. For the first time, E. R. Brown in 1998 introduced design of a reconfigurable antenna based on RF-MEMS switches (Brown 1998). MEMS switches offer good frequency characteristics, such as small size, low loss, and harmonics beside high integration. MEMS switches have been commonly used in the microstrip antennas structure for radiation pattern reconfigurability (Huff and Bernhard 2006; Jung et al. 2006; Russo et al. 2010; Coutts et al. 2008).

For example, in (Huff and Bernhard 2006), two MEMS switches are implemented inside a resonant square spiral microstrip antenna to reconfigure the radiation patterns between the end-fire and broadside modes over a common impedance bandwidth. In (Jung et al. 2006), a set of RF-MEMS switches are planted inside a spiral antenna, which is fed in the center by a coaxial cable, to change the radiation beam direction of the antenna by controlling the status of the integrated RF-MEMS switches.

The current path of Frequency Selective Surfaces can also be controlled by RF-MEMS switches. For example, in (Russo et al. 2010) beam-steering capability of active FSSs working in transmission mode is demonstrated. A slot grid FSS, which is the complementary structure of a dipole grid, is designed, in which MEMS switches are imbedded through the FSS to locally control the frequency response of the FSS. Therefore, to obtain large phase agility a pass-band FSS loaded with varactor has been designed. In this work, a multilayer arrangement is presented to allow $\pm 90^\circ$ steering angle for the transmitted beam due to the phase agility of about 360° for the unit cell. Furthermore, the introduced technique permits a simultaneous control of amplitude and phase.

The disadvantage of MEMS switches is that they have not been broadly marketed for fabrication as it is costly, and hence design of active FSSs based on MEMS switches is limited. Moreover, for the reconfigurability of the FSS the MEMS switches have mostly been used as a capacitive load on an aperture. However, the capacitance variation of the MEMS switches is very low while the cost of fabrication is very high. Moreover, MEMS switches offer slower response

and higher bias voltage compared to the pin diodes which are another disadvantage of MEMS switches.

Reconfigurable FSS-Incorporated Pin Diode Switches

Pin diode switches offer two different states of ON or OFF and are broadly used in FSS for beam-tilting antenna applications. The PIN diode can be modeled with an equivalent RC circuit with OFF-state parallel RC circuit and ON-state resistor as shown in Fig. 15.

Edalati et al. in (Edalati and Denidni 2013) have proposed a beam-tilting antenna based on reconfigurable FSSs incorporated with PIN diodes at 2.5 GHz. In this work, metallic discontinuous strips are considered as FSS unit cell design in which PIN diodes are located on its discontinuities to modify the transmission and reflection coefficients. The PIN diode is modeled with its equivalent RC circuit with OFF-state parallel RC circuit and ON-state resistor as shown in Fig. 16a and b. The gap dimension is considered to be almost equal to the diode dimension. In the low frequencies, the FSS shows a high transmission coefficient when the PIN diode is on its OFF-state and a high reflection coefficient in the ON-state.

The beam-tilting antenna presented in this work consists of a reconfigurable FSS in a cylindrical shape which is surrounding an omnidirectional dipole antenna as shown in Fig. 16c. The cylindrical shaped FSS is made of 18 columns, each having six strips while five PIN diodes are embedded in the gaps of the strips and five resistors are located at the top and bottom of each line. Using metallic sheets, the cylindrical FSS is divided into six sectors, and to create a directive radiation pattern, two metallic cones are placed at the top and bottom of the structure. In this technique, to switch the antenna's radiation pattern between the sectors, each time the embedded diodes in one FSS-sector are set to OFF to make the sector transparent to the incident electromagnetic waves; while, the diodes on the other sectors are set to ON to be opaque and reflect the incident wave. In summary, it means each time one sector is open and all the other sectors are kept to be close to the EM waves propagation. Thus, the main beam

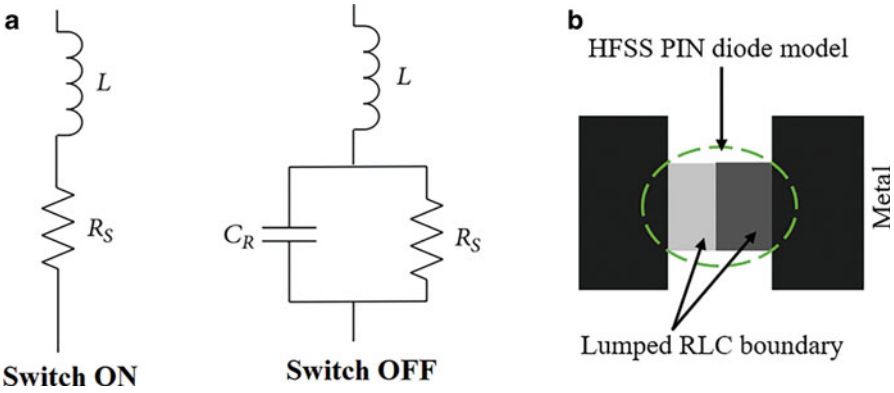


Fig. 15 (**a**) Equivalent circuit model and (**b**) HFSS model of RF PIN diode

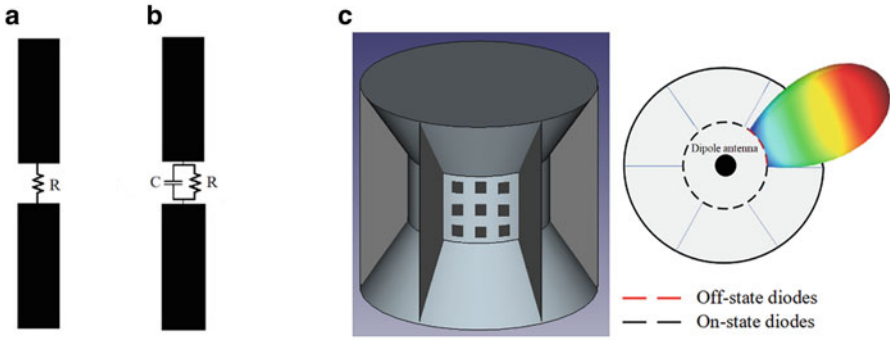


Fig. 16 Schematic of the FSS unit cell presented in (Edalati and Denidni 2013) for (**a**) ON-state equivalent circuit; (**b**) OFF-state equivalent circuit; and (**c**) schematic of the beam-tilting antenna

will be directed toward the sector with OFF-state PIN diodes and provides the antenna beam-tilting toward desired directions with the steering ability of the antenna's radiation pattern in six steps over 360° azimuth plane.

The disadvantage of the PIN diode switches is that the biasing network reflection may decrease the overall antenna's performance. In addition, using a large number of PIN diodes per antenna makes the structure high cost and lossy.

Reconfigurable FSS-Incorporated Varactor Diodes

In terms of functionality, Varactor diodes and PIN diodes are very similar, which capacitances are controlled by adjusting the reverse bias voltage. By switching the characteristics of embedded varactor diodes on FSSs, the tunable frequency selective surface can be realized. For example, Pan et al. in (Pan et al. 2013) have integrated multilayers of active FSS as a radome in front of a horn antenna to show the beam-tilting ability of FSS. As shown in Fig. 17a, the designed FSS unit cell is composed of a metallic square ring and a patch while two Varactor diodes are placed between them. A wire with an embedded via is also placed on the opposite side of the square slot on the unit cell. The varactor diodes are able to tune the emitted wave transmission phase by changing their bias voltage. Five FSS layers, each consisting of 6×6 unit cell, are located in front of the horn antenna as illustrated in Fig. 17b. Through different configurations of the bias voltages, beam-tilting in a range of $\pm 30^\circ$ in both H- and E-plane at 5.3 GHz is realized with a bandwidth of 180 MHz. However, the bias network that is used to control Varactor diodes in the unit cells has negative effect on the antenna parameters.

Reconfigurable FSS-Incorporated Cantilever Switches

In active FSS based on cantilever switches, loss, capacitive variation, and need of biasing network are less compared to the PIN diode and MEMS switches. In fact, the cantilever arms located on the top layer of the unit cell, as shown in Fig. 18, work as an active switch. The resonance frequency of the unit cell is tuned by the capacitive loading since the dielectric substance at the cantilever arms is isolating the two metallic

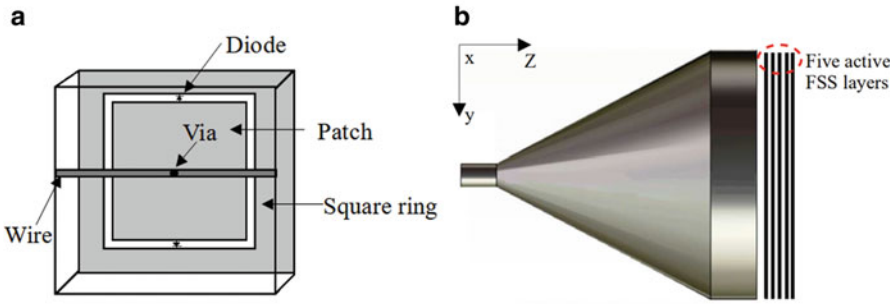
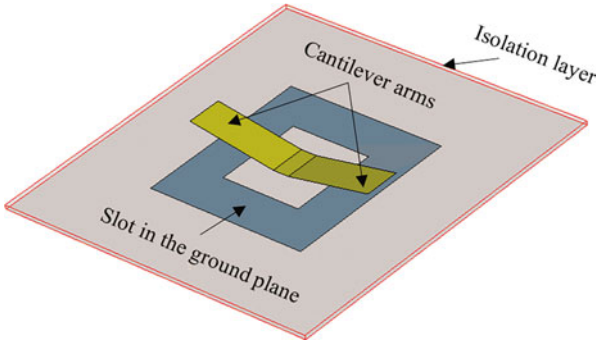


Fig. 17 Schematic model of (**a**) the FSS unit cell incorporated Varactor diodes and (**b**) horn antenna loaded with multilayers of active FSS presented in (Pan et al. 2013)

Fig. 18 Schematic of a cantilever-enabled FSS



layers. The slot's equivalent capacitance varies by the downward and upward movements of the cantilever arms, and therefore providing a frequency tunability for the FSS unit cell. In cantilever-enabled FSS, the design fundamentally consists of two DC-isolated metal layers, the lower ground plane with slot which is static and the upper cantilever arms which are the movable parts. The cantilever arms are attracted toward the ground plane and pulled down to the dielectric layer when a sufficient DC voltage is applied between the two layers. Changing the cantilever arms' positions changes the capacitance and hence shifts the resonant frequency of the FSS.

In (Kesavan et al. 2018), a reconfigurable cantilever-based FSS is proposed for beam-tilting applications at 30 GHz. An omnidirectional dielectric resonator antenna (DRA) consists of a cylindrical DR which is considered an exciting source which is located on the ground plane and is fed through coaxial line as showed in Fig. 19a. The dielectric resonator antenna is located in the middle of six side-panels with hexagonal geometry; each consists of 1×5 array of the cantilever-enabled FSS unit cell (see Fig. 19b). To convert the omnidirectional radiation pattern of the DRA to directive beam in the azimuth plane, each time, four panels are set to be in the DOWN state while two consecutive panels are fixed in UP state. In this situation, the UP-state panels act as bandpass FSS and therefore are transparent to the radiation

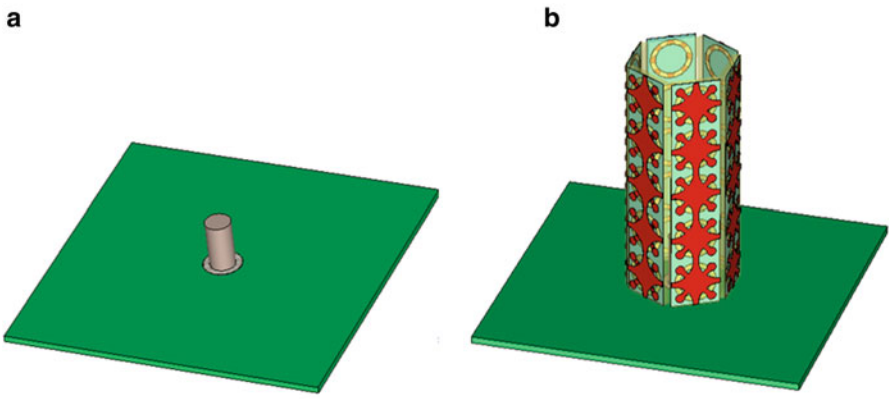


Fig. 19 Configuration of (**a**) DRA antenna and (**b**) DRA antenna located in the middle of an FSS hexagonal cylinder based on cantilever switches presented in (Kesavan et al. 2018)

from the DRA antenna. However, the DOWN-state panels blocked the radiation and reflected them. Therefore, by changing the state of the six panels with different orders, beam-switching in six different directions in the DRA azimuth plane is reached. Moreover, up to 3 dBi gain enhancement is obtained due to using FSS compared to the single DRA antenna.

The disadvantage of cantilever switches is their difficult fabrication process and low speed of switching compared to PIN diode.

Conclusion

In the fifth generation of wireless communications, beam-tilting antennas are becoming very popular due to their ability to reduce the rate of interference of communication systems and improve the system coverage by focusing the radiated power toward desired direction rather than whole areas. Frequency Selective Surface (FSS) as a metamaterial is broadly used for beam-tilting applications and has drawn a lot of attention in the past decades and also will continue to be a driving field of research in the future. This chapter has introduced the conceptions of FSS and beam-tilting and showed various advancements in FSS-based beam-tilting antenna designs introduced in recently published works in the literature. In 5G, millimeter-wave frequency spectrum is an essential feature, and the ability of tilting the antenna's beam within that band is essential.

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Wideband MM FSS Reflector for CP Millimeter-Wave Antennas

16

Amit Kumar, Binod Kumar Kanaujia, and Abdul Quaiyum Ansari

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Abstract

In this chapter, we have discussed the importance of Circularly Polarized (CP) Millimeter-wave (MMW) Antennas (30–300 GHz) as a need for high data rate transmission up to 10 Gbp/s, real-time video streaming, and higher transmission

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bandwidth like in 5G at 28, 38, and 60 GHz MMW frequencies. We have shifted to MMW frequencies to have wider bandwidth and less congestion on the link. But losses increase as the wavelength decrease causing point-to-point communication and getting interference from trees, buildings, and many more. A high-gain antenna is required to encounter the losses at the MMW frequencies. We have focused on CP MMW antennas as these are immune to Faraday rotation, mitigation of multipath propagation, and polarization mismatch loss and thus enhance the performance of the MMW antennas in a lossy environment. To achieve high gain, linear to circular polarizer, wider CP Bandwidth, spatially coupling reduction in MIMO Antennas and Broadband Radar Cross Section (RCS) Reduction in MMW antennas, MM Frequency Selective Surface (FSS) reflector importance is comprehensively discussed, examined, and encouraged.

Keywords

Circularly Polarized (CP) · Frequency Selective Surface (FSS) · Millimeter-wave (MMW) Antennas · Fabry-Perot cavity (FPC)

Introduction

The demand for high-speed wireless data transmission like up to 10 Gbits/s especially in 5G has encouraged us to move to millimeter-wave (MMW) frequencies to have wider bandwidth and less congestion on the link. But losses increase as the wavelength decrease causing point-to-point communication and getting interference from trees, buildings, and many more (Glushko et al. 2013; Sun et al. 2013). Also, during satellite communication, the electromagnetic waves (EM) have to encounter the ionosphere where the problem of Faraday rotation arises which can be resolved by using a circularly polarized (CP) antenna. To overcome the losses high-gain antenna is preferred in wireless communication. So, to have a linearly to circularly polarized high-gain antenna we need a Frequency Selective Surface (FSS) which acts as a reflector for a particular range of frequency and also acts as a polarizer convertor (Akbari et al. 2017b). Let's have a look at the importance of MMW frequencies, CP antennas, and FSS.

Millimeter-Wave

In today's era of high-speed wireless data transmission of almost $\geq$100 Gbps for the trending Fifth-generation (5G) and even more for Sixth-generation (6G) wireless communications and Internet-of-Things (IoT) technology, we need to shift to Millimeter-wave (MMW) antennas having an operating frequency ranging from 30 to 300 GHz. The larger available bandwidth (BW) in this MMW electromagnetic spectrum will support up to 10 Gbits/s which will be supporting "real-time" video streaming, Machine-to-Machine (M2M) communications, and IoT. The major

disadvantage of the MMW systems is high attenuation in the ionosphere region of the atmosphere which weakens the strength and range of the traveling waves. Due to MMW frequency, the wavelength is in few mm due and so travels in a line of sight and get obstructed by physical objects like trees and buildings (Glushko et al. 2013; Sun et al. 2013). To overcome the signal fading and physical obstruction high-gain antenna and CP polarized antenna are preferred for MMW communications. To increase the gain we use MM Frequency Selective Surface (FSS) reflector. CP MMW antennas help to overcome the challenges of polarization mismatch, signal drop-out, adverse weather conditions, phasing issues, multipath fading, and non-line-of-sight applications (Huang et al. 1999; Kalhor et al. 2012; Akbari et al. 2017c).

Advantages of CP

Immunity to Faraday Rotation

Linearly polarized (LP) signal faces the problem of Faraday rotation as it passes through the ionosphere. The polarization plane changes as the EM wave comes in contact with the Earth's magnetic field and the charged particles which become more problematic as we move higher in the ionosphere where the magnetic flux intensity increases due to highly ionized plasma (Davies and Smith 2002). Due to two orthogonal components in the CP antennas, any Faraday rotation will be on both components, and hence the CP characteristics will maintain in every sense and so immune to Faraday rotation (Choubar 2018).

Mitigation of Multipath Propagation

This phenomenon occurs when at the same time both the reflected signal and the original signal will reach the receiver's end due to multi-path which creates a problem of "out of phase." Now, it becomes difficult for the receiver to distinguish and process the proper signal. LP antennas have severe multi-path distortion due to the higher probability of rotation and reflection (Choubar 2018).

Polarization Mismatch Loss

The problem arises due to the misalignment of the transmitting and receiving antennas. While in the case of the CP antenna, there is always a component aligned to the horizontal or vertical plane, unlike the LP antenna.

FSS

The periodic structure has special importance in the field of microwave and antenna applications. The special advantage of structure periodicity helps to achieve some desired requirements in the proposed design. FSS is one of the most famous periodic structures on which many applications are based, such as its characteristics as a frequency filter. FSS enhances the gain of the antenna, CP bandwidth, reduces spatial coupling among elements of a MIMO antenna, and suppresses the radar

cross-section (RCS) of different devices, such as random, reflector antennas, microwave circuits, etc. We will discuss the FSS advantages in detail in section "Overview of FSS."

In this book chapter, we will be discussing the importance of CP MMW antennas. The significance of FSS for achieving Gain enhancement, Circular polarization, Spatial coupling reduction, and RCS reduction are discussed in this book chapter. We are going to discuss the FSS background and applications in section "FSS Background and Applications," the overview of FSS along with its advantages in section "Overview of FSS," FSS Layer: Enhancing antenna gain in section "FSS Layer: Enhancing Antenna Gain," Advantages of FSS-type polarizer in section "Advantages of FSS-Type Polarizer," and lastly Benefits of RCS reduction in section "Benefits of RCS Reduction" followed by conclusions and future Work.

FSS Background and Applications

When exposed to EM radiation, FSS works as a filter that allows some EM waves of particular frequency bands while some are reflected. Unwanted interference and mitigating noise can be handled by this FSS-based spatial filter. Due to this filtering characteristic which helps to reduce the RCS, FSS structures help in the design of radomes for the radar antenna. In the early 1960s, the structures based on FSS have been an area of intensive research due to their potential military applications (Wu 1995, 2005; Vardaxoglou 1997; Munk 2000). Marconi and Franklin were the first (Munk 2000) to design a parabolic reflector in 1919 (Marconi and Franklin 1914). FSS is now extensively used in radomes, missiles, stealth technology, and EM shielding applications. FSSs are most commonly found in two-dimensional space in the form of planar, periodic metal-dielectric arrays. FSS frequency's behavior is determined by the geometry of the surface in one unit cell provided the surface size is infinite. Mainly, the FSSs operational mechanism can be explained by the resonance phenomenon.

In a research article (Wu 1995), it has been observed that an array of elements resonates at the frequencies corresponding to the effective length of the elements generally is in the multiple of $\lambda/2$. The resonance characteristics defer the way the EM wave approached the surface based on the FSS structure. A practical example to secure buildings (Parker et al. 2004; Parker and Savia 2004; Turitsyna and Webb 2005; Batchelor et al. 2009; Samadi et al. 2019) has been shown in Fig. 1a, b. A collaborative project to control wireless propagation inside the building based on FSS has been jointly carried out between the universities of the UK and New Zealand. The project aimed at finding solutions to spectral overcrowding, security, and interference while allowing radio systems to carry out their emergency services (TETRA in Europe) while entering the building (Batchelor et al. 2009) as shown in Fig. 1b. For better selectivity and performance three-dimensional FSS building structures are preferred over two-dimensional wallboard structures (Parker et al. 2004). One small isolation capability based on FSS screened has been described in Parker and Savia (2004).

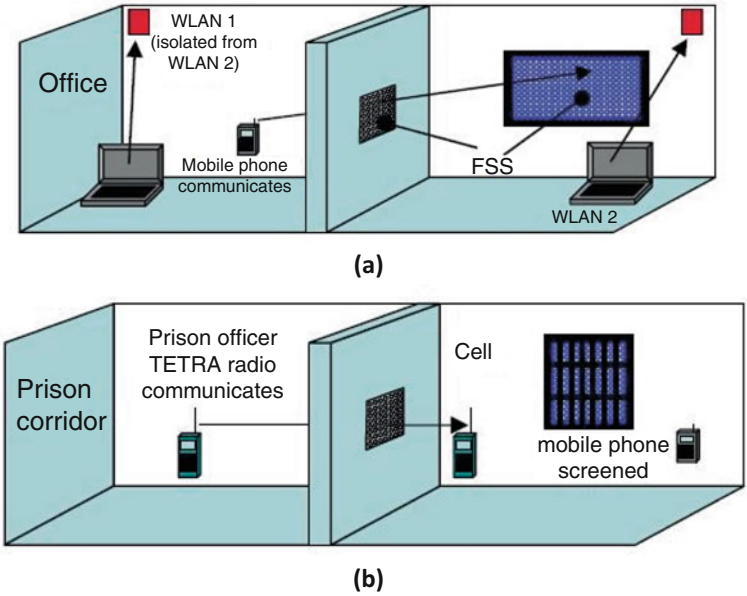

Fig. 1 Demonstration of finite FSS panel usage in (**a**) a civil office environment (Batchelor et al. 2009) (**b**) a prison building environment (Batchelor et al. 2009)

Some more practical examples for conference and study rooms have been discussed in Qasem (2012) as shown in Fig. 2a, b.

FSS structures can also be used to act as low emittance windows (Gustafsson et al. 2005) as depicted in Fig. 3, it will block infrared radiation from the sun while it will allow visible light and microwave frequencies to pass through it.

Overview of FSS

FSSs are planar periodic structures that characterize various features of transmission and/or refection based on operating frequencies. FSS surface is periodic strictures in either one direction or both, made up of thin metallic materials of few micro-mm thicknesses printed on a dielectric substrate of thickness in the range of mm. These structures are further categorized into aperture type or patch type. The structural orientation decides its frequency characterization: “low-pass,” “high-pass,” “band-pass,” and “band-stop” (Gustafsson et al. 2005) as shown in Fig. 4. We can see in Fig. 4 that the low and high-pass, similarly band-pass and band-stop filter response structures are complementary to each other and as a whole cover the entire metallic area. The same concept has been strengthened according to Babinet’s principle that the transmission and reflection coefficients are complementary to each other based on the complementary structure characteristics of the surface.

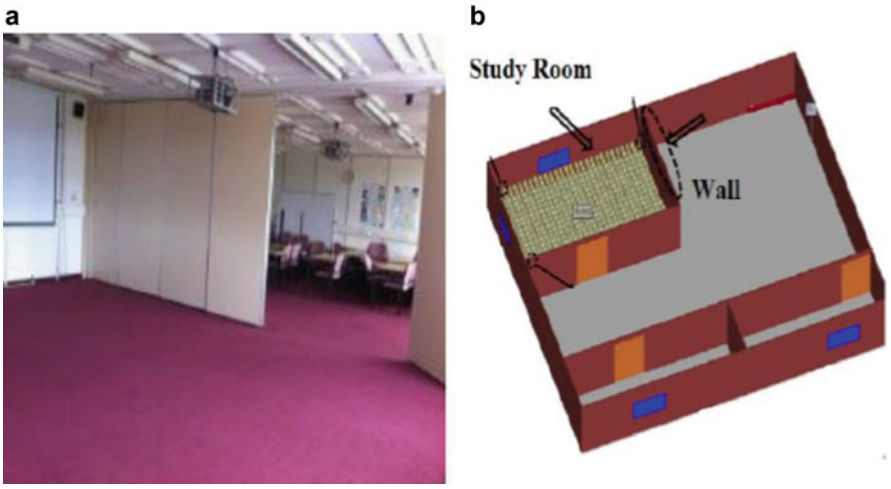

Fig. 2 A prototype of FSS walls (**a**) for a conference room (Qasem 2012) (**b**) a study room (Qasem 2012)

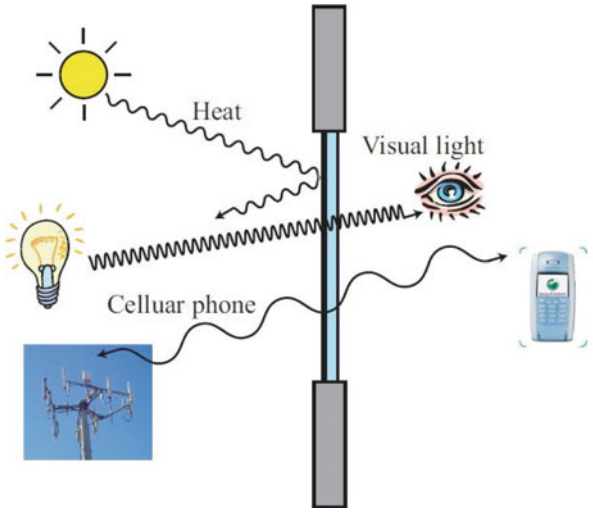

Fig. 3 A view of frequency selective window (Gustafsson et al. 2005)

Types of FSS

Turki (2015) has mentioned several designs of FSS shapes with each design/shape having its unique characterization of reflection/transmission characteristics, polarization type, bandwidth, and sensitivity to the oblique angle of incidence. While in Munk (2000) FSS arrays are classified into four categories as shown in Fig. 5a–d. The dipole elements are shown in Fig. 5a to help create single-polarization, while

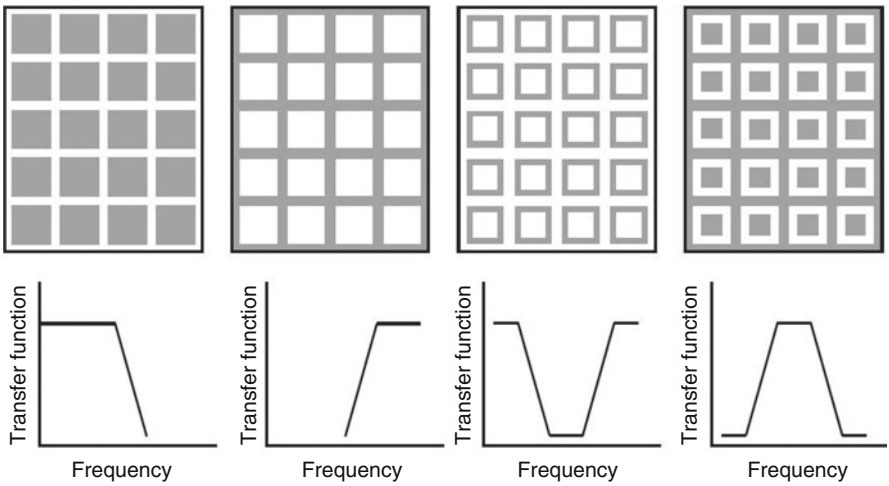

Fig. 4 FSS structures and their frequency responses (Gustafsson et al. 2005)

cross dipole elements are shown in Fig. 5b can create dual polarization. The unloaded tripole arrays can be tightly packed and helpful in achieving dual polarization. Wider bandwidth and out-of-band grating response can be achieved by anchor elements due to the end capacitance (Wu 1995; Munk 2000; Turki 2015).

Advantages of FSS

Gain Enhancement

A considerable enhancement in gain is achieved by illuminating a Fabry-Perot cavity (FPC) by a source antenna (Ip and Jackson 1990; Feresidis and Vardaxoglou 2001; Lee et al. 2004; Liu 2010; Jackson et al. 2011). In FPC approach-based antenna, an air-filled cavity is generated between the ground plane and the partially reflective surface (PRS), or an FSS superstrate placed at a distance of half-wavelength. FSS superstrate helps in enhancement of gain of circularly polarized dielectric resonator antenna at 30-GHz. It also enhances the CP bandwidth (Akbari et al. 2016a).

Spatially Coupling Reduction

Spatial coupling is reduced in MIMO Antennas Using FSS (Akbari et al. 2017a; Farahani et al. 2017). In Akbari et al. (2017a) the FSS layers reflect a large portion of the incident EM wave thus concentrating the space wave upwards and hence minimizing the space wave coupling in MIMO antennas. While in Farahani et al. (2017) an H-plane-oriented metamaterial polarization-rotator wall is placed between two DRAs which helps in minimizing spatial coupling as the wall makes the TE modes of the antenna orthogonal.

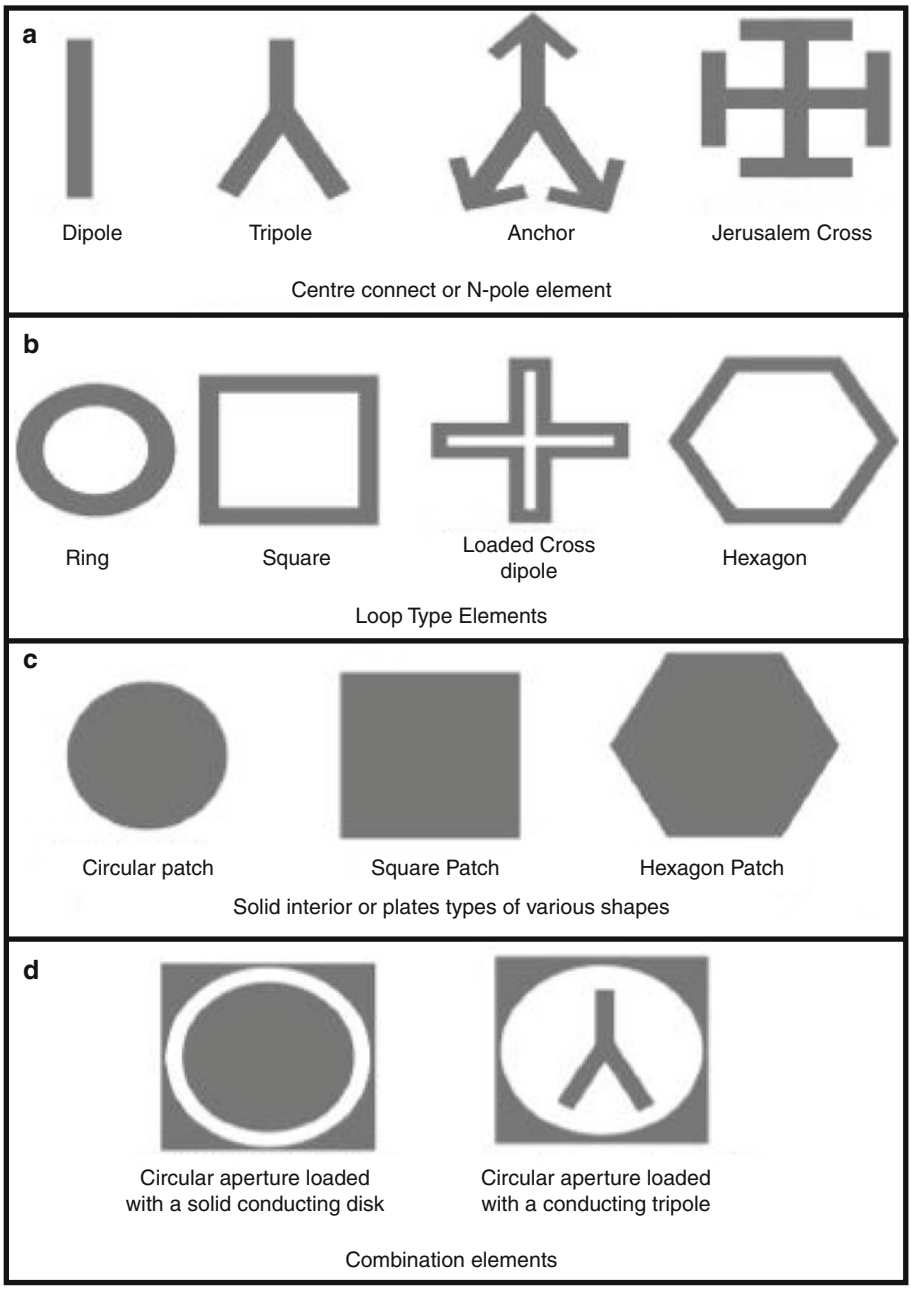

Fig. 5 Common FSS element types (Turki 2015)

Polarizer

The polarizer is placed on the top of the radiating structure of the planar array to achieve circular polarization of the transmitted waves. Basically, the polarizer decomposes the Linearly polarized incident plane wave into nearly two orthogonal components with a phase difference of 90° of equal amplitude which helps in achieving CP transmitted plane wave. Several polarizers have been investigated in the existing literature (Joyal and Laurin 2012; Zhu et al. 2013, 2014; Martinez-Lopez et al. 2014; Orr et al. 2014; Alonso et al. 2015; Fei et al. 2015; Li et al. 2015; Abadi and Behdad 2016; Cheng et al. 2016). FSS is generally used to design planar polarizers. These planar polarizers decompose a wave into two orthogonal components by providing a reactive surface to one and capacitive to another component of the field or vice versa. When the two decomposed components have equal magnitudes, the incident LP wave is converted to the transmitted CP wave (Abadi and Behdad 2016). Many polarizers were designed using FSS structures (Martinez-Lopez et al. 2014; Orr et al. 2014; Li et al. 2015; Abadi and Behdad 2016), while others are based on traditional line and loop configuration (Joyal and Laurin 2012; Alonso et al. 2015; Fei et al. 2015). In Li et al. (2015) truncated square patches were used to design a multilayer FSS to have a phase difference of 90° between components orthogonal to each other. The polarizer described in Abadi and Behdad (2016) has arrays of capacitive patches of subwavelength and wire grids of subwavelength depicting inductive characteristics well insulated by dielectric substrates to achieve orthogonality. A double-sided PRS has been used in the polarizer (Orr et al. 2014). A four-layer FSS based polarizer is proposed in Martinez-Lopez et al. (2014) has been bisected by a metal strip that consists of split-rings.

Broadband RCS Reduction

Broadband RCS Reduction is done using FSS Metasurface (Akbari et al. 2019; Zarbakhsh et al. 2019). In Zarbakhsh et al. (2019) the proposed antenna is based on sequential feeding technique and the Fabry-Perot cavity (FPC). The proposed structure produced highly directive wideband CP maintaining low RCS. All the above-mentioned benefits were achieved by an appropriate FSS metasurface having tuned feeding lengths to meet the required phase variation. A reflection phase variation of $180^{\circ} \pm 37^{\circ}$ has been realized on the metasurface between adjacent FSS unit cells resulting in RCS suppression.

There has been a comprehensive improvement of gain of at least 8 dB and a peak gain of 20 dB at 28.5 GHz with metasurface as compared to the primary antenna. A 3-dB axial-ratio from 27.5 to 33.5 GHz (19.7%) was achieved. For a normal incidence, monostatic RCS reduction has been achieved from 28 to 48 GHz (52%). So, FPC can be enormously helpful to gain enhancement, achieve low RCS, and achieve CP features (Choubar 2018).

FSS Layer: Enhancing Antenna Gain

Due to the various losses at MMW frequencies, highly efficient and high-gain antennas have been the center of research and development. The demand can be fulfilled by the reflector and parabolic antennas, but they are bulky and need mechanical motors for beam steering. Planar structures based on array antennas have the capability of steering of beams by a complex feeding network on the circuitry by controlling the phase distribution. But due to its complex feeding network or circuitry, the antenna suffers from structural losses. On the other hand, planar FPC antennas can produce directional radiation patterns using a single excitation point and be creating cavities by placing a PRS on top of a ground plane, as shown in previous works (Ip and Jackson 1990; Feresidis and Vardaxoglou 2001; Lee et al. 2004; Zhao et al. 2005a, b; Gardelli et al. 2006; Lovat et al. 2006; Lovat and Burghignoli 2006; Liu 2010; Jackson et al. 2011; Orr et al. 2014; Akbari et al. 2016a, b, 2018; Asaadi et al. 2018; Zarbakhsh et al. 2019; Goudarzi et al. 2020). We can see in Fig. 6a, b that how an FPC antenna based on the FSS superstrate layer has been designed (Akbari et al. 2016a). All layers of photographs have been shown in Fig. 7a–d.

An FSS superstrate has been placed above a basic aperture coupled Dielectric Resonator Antenna (DRA) maintaining some air gap to improve the antenna gain as shown in Fig. 8a, b. The use of more FSS superstrates will result in various resonant frequencies, but the size of the antenna will be an issue. One drawback of using the FSS superstrate has been the reduction in the impedance bandwidth. The layout and the dimensions of the unit FSS cell has been shown in Fig. 8a. Figure 8b depicts a sketch of multiple reflections inside the cavity created between the aperture coupled DRA and the FSS superstrate and the EM wave leakage through the superstrate which got concentrated in the broadside direction of antenna radiation pattern when all the wave phases are precisely tuned to each other. So, a considerable directed beam has been created in the antenna broadside direction which results in increased antenna gain. The equivalent circuit model of the FSS superstrate coupled with the aperture antenna has been depicted in Fig. 9 (Pirhadi et al. 2012). PRS superstrates

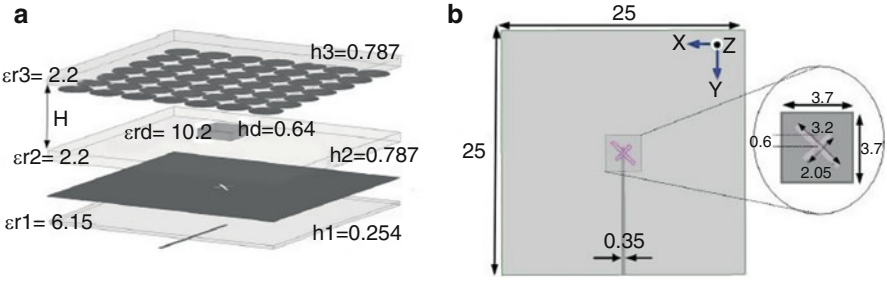

Fig. 6 Outline of the proposed FPC along with dielectric resonator (DR) and superstrate layer based on FSS (**a**) 3D view (Akbari et al. 2016a) and (**b**) top view (Akbari et al. 2016a)

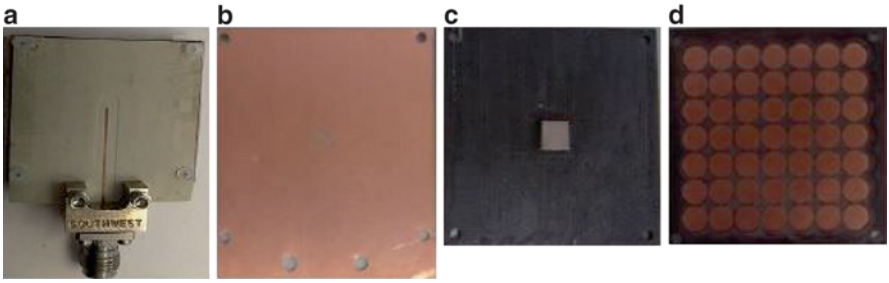

Fig. 7 View of the fabricated antenna (**a**) First layer: Bottom view (feed line), (**b**) First layer: Top view (ground plane), (**c**) Second layer: Top view (DR), and (**d**) Third layer: Bottom view (FSS elements) (Akbari et al. 2016a)

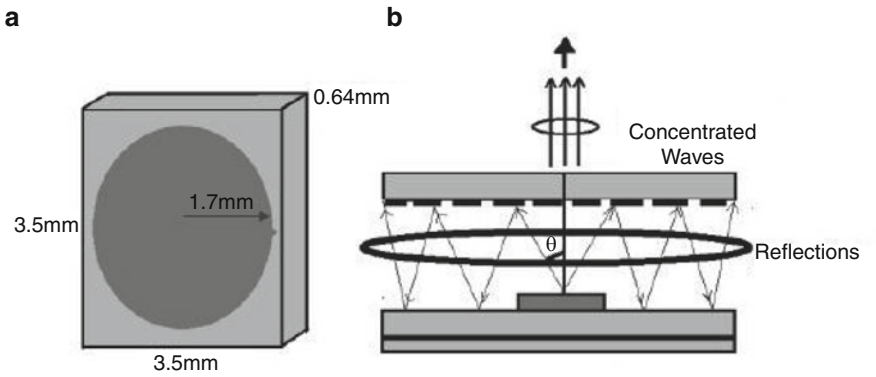

Fig. 8 (**a**) FSS unit cell (Akbari et al. 2016a) and (**b**) depiction of multiple reflections and leaky-waves through FSS superstrate (Akbari et al. 2016a)

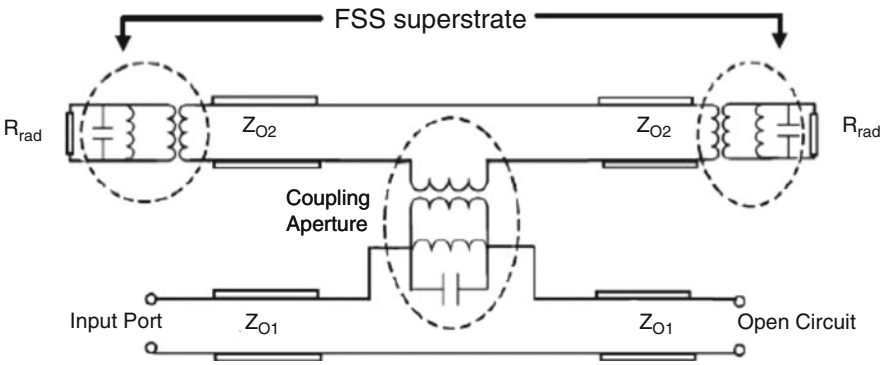

Fig. 9 Equivalent circuit model of the FSS superstrate layer coupled with the aperture antenna (Pirhadi et al. 2012)

made of periodic metallic patches or slots have been used to design FPC antennas (Zhao et al. 2005a, b, c) or multilayer dielectric slab superstrates (Jackson and Oliner 1993; Gardelli et al. 2006). In Lovat et al. (2006) and Lovat and Burghignoli (2006), the radiation behavior of these FPC antennas was studied through excited leaky-waves (LW) inside these cavities.

Based on leaky-waves (LW)'s explanation (Lovat et al. 2006; Lovat and Burghignoli 2006), a broadside directive radiation pattern like a pencil beam has been produced perpendicular to the FSS ground plane around its resonating frequency. In the case of frequencies higher than the resonating frequency of the cavity, the beam starts shifting to the sides of the antenna. The problem can be settled by using FSS based on highly reflective inductive or capacitive surfaces or changing the height of the cavity resonance that can control the main beam of the antenna around its center frequency. But 3-dB gain/power bandwidth of such cavities limits the bandwidth of the FPC antennas.

Advantages of FSS-Type Polarizer

We have already discussed the CP antenna's benefits over the LP antennas. Now, we will see how we can convert an LP antenna to a CP antenna and also enhance CP bandwidth using an FSS-type polarizer. A simple way of creating a CP wave is to make the LP wave pass through the polarizer plane.

A schematic model of LP to CP polarizer has been shown in Fig. 10. The polarizer has come into action when the LP EM wave strikes its surface while propagation in the k direction and gets tilted by 45° from the x-axis. Thus the decomposed orthogonal electric field will have one component along the x-axis and the other along the y-axis (Akbari et al. 2017b).

The method in which the polarizer is placed on the top of the antenna to transmit CP waves becomes more attractive and productive when the radiating structure contains a planar array unlike many complex designs (Martinez-Lopez et al. 2014; Orr et al. 2014; Li et al. 2015; Abadi and Behdad 2016) which consist of multilayer structures. In Fei et al. (2015) single layer polarizer having meandered line loop geometry has been proposed as shown in Fig. 11a, b. The 3-dB axial ratio BW has been 46.8% ranging from 18 to 29 GHz. Such kinds of single-layer polarizers are best suitable for conformal antenna applications where we need to tilt or bend the antenna as per the requirements which are not possible for multilayer or bulky structures.

In Alonso et al. (2015) CP has been achieved by implementing a meander line polarizer with the substrate integrated waveguide (SIW) antenna at Ku-Band as depicted in Fig. 12a. Alonso et al. has used an eight standing-wave slotted SIW array with a meander line polarizer on the top to obtain the CP.

Some tapering among slots has been done to reduce the sidelobe level. On the lower level, they have introduced a distribution network to reduce antenna size along with a coupling H-slot was to couple the power. In another approach, rotating FSS structure (Zhu et al. 2014) was implemented using SIW technology as represented in

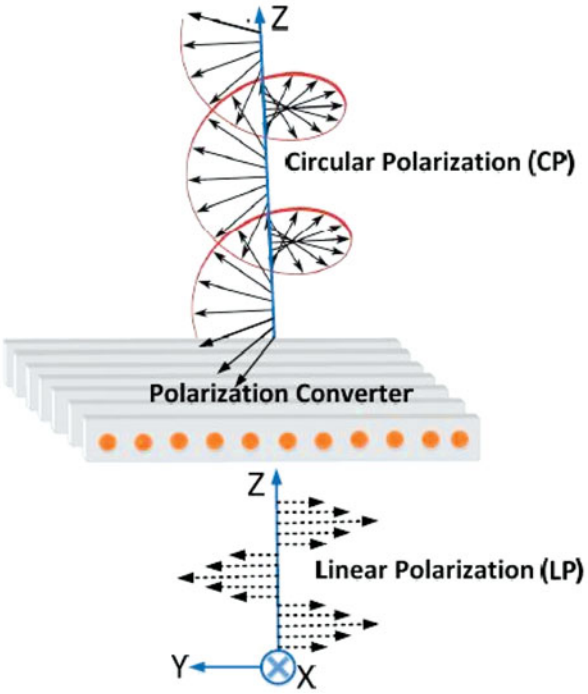

Fig. 10 Linear-to-circular polarizer (Akbari et al. 2017b)

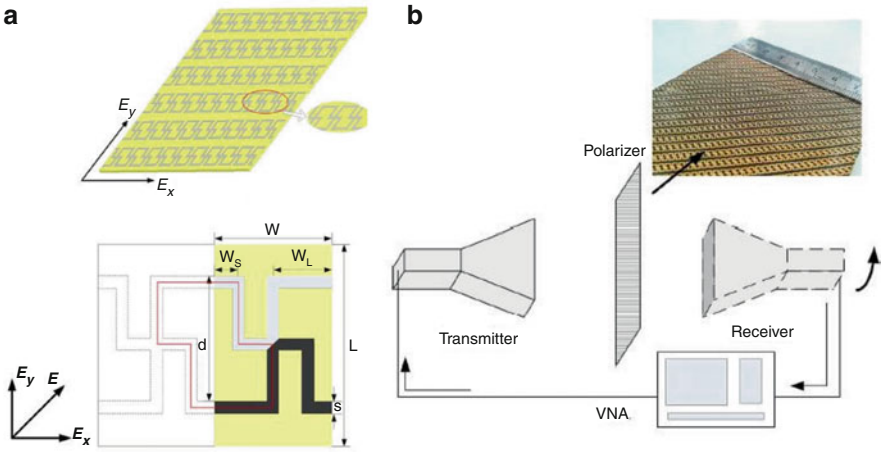

Fig. 11 Single layer polarizer (**a**) polarizer layer and single-cell view (Fei et al. 2015) (**b**) test setup of the polarizer (Fei et al. 2015)

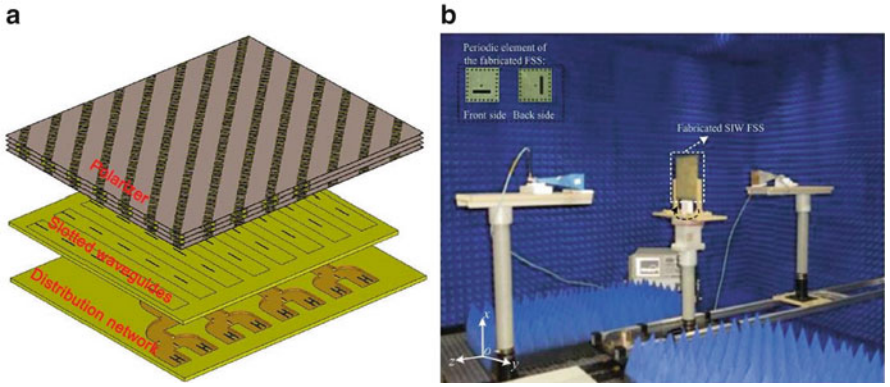

Fig. 12 (**a**) Meander line polarizer based on SIW antenna (Alonso et al. 2015) (**b**) measurement setup of SIW FSS (Zhu et al. 2014)

Fig. 12b. The proposed FSS was made up of a periodic array of the novel triple-mode SIW cavity resonators with two orthogonal slots on the front and back surfaces for polarization rotation. The triple-mode SIW cavity resonator was achieved through a metallic via placed at the center of the dual-mode SIW cavity resonator. Properly merging these three modes of the SIW cavity resonator helps in CP BW enhancement. Sometimes, metamaterial and meta-surface have been used to enhance the CP BW (Zhu et al. 2013; Cheng et al. 2016). In Zhu et al. (2013) the transmission BW has been improved using a SIW cavity resonator based on triple-mode. The meta-surface proposed in Zhu et al. (2013) has 16 rectangular loop unit cells. An ultra-compact chiral metamaterial (CMM) was suggested in Cheng et al. (2016) having twisted split-ring resonators (TSRRs) triple-layer structure for CP bandwidth enhancement.

Benefits of RCS Reduction

A polarization conversion-based RCS reduction using wideband FSS metasurface has been proposed (Choubar 2018). The group search optimization (GSO) algorithm has been used to generate a binary coded matrix on which basis the elements are arranged to suppress the maximum bistatic RCS by redistributing the scattered EM waves on the metasurface. To achieve a phase difference of 180° between the two elements "0" and "1" of the binary coding matrix on a broad frequency band, the designing of reflective two-layer metasurface has been kept on priority. Later on, Least Square Error (LSE) method has been used on the ratio of the "0" and "1" elements to find the best ratio value.

RCS reduction for wider BW and controlling the scattering of EM waves have been always challenging for our designers (Chen et al. 2015). This in turn helps in increasing antenna gain helpful at MMW frequency. Simultaneously, easy integration of FSSs with objects is even more challenging (Iriarte Galarregui et al. 2013). Previously many traditional methods like loading of the object with the radar absorbing materials

(RAM) and/or passive or active cancellation have been done to prevent the scattered waves from going in the backscatter direction (Salisbury 1952). But the implementation of RAM suffers from some disadvantages like its overall thickness, weight, air drag, and also its angle and frequency sensitivity that restricts these to a narrow operating frequency band (Salisbury 1952; Iriarte Galarregui et al. 2013; Edalati and Sarabandi 2014; Chen et al. 2015). In Paquay et al. (2007) artificial magnetic conductors (AMC) as well as perfect electric conductors (PEC) have been used together for achieving RCS reduction. AMC and PEC unit cell together produce an overall phase difference of 180° which results in destructive interference and hence creating a null in the specular angle (Salisbury 1952; Knott et al. 1993; Paquay et al. 2007; Iriarte Galarregui et al. 2013; Edalati and Sarabandi 2014). The main drawback in the AMC cell-based approach is the narrowband behavior as low as 5% only as described in Knott et al. (1993). But in Iriarte Galarregui et al. (2013) [97], two Jerusalem Cross AMCs with different sizes were implemented to achieve RCS reduction BW of nearly 40%. Simulated RCS BW of approximately 58% was reported in Zhang et al. (2009) while by modifying the blended checkerboard design (Zhang et al. 2009) almost 91% 10-dB RCS reduction bandwidth was achieved showing a considerable improvement in the RCS reduction fractional BW. However, it is difficult to manipulate the polarization of EM waves and at the same time maintain a good RCS reduction fractional BW. Recently some efforts have been made for RCS reduction which mainly depends on the principle of polarization conversion (Feng et al. 2013; Jia et al. 2016; Liu et al. 2016), but the practical application has been limited due to narrow bandwidth. For example, in Liu et al. (2016) fishbone-shaped arrays having a polarization conversion metasurface (PCM) creates narrow bands. A polarization rotation reflective surface (PRRS) has been used in Jia et al. (2016) in which a square patch is connected through vias to the ground which helps in achieving RCS reduction BW of 49%. However, the bandwidth enhancement of these kinds of designs still requires some rigorous research.

A high gain and low RCS CP antenna based on the FPC cavity have been proposed in Akbari et al. (2018) as illustrated in Fig. 13a, b. We can see the combination of the FPC antenna with the polarization conversion surface (PCS) in Fig. 13b. The proposed

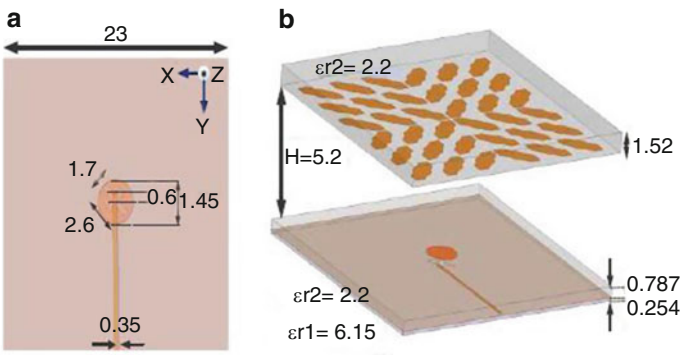

Fig. 13 Geometry of the (**a**) source antenna (Akbari et al. 2018) (**b**) along with PCS (Akbari et al. 2018)

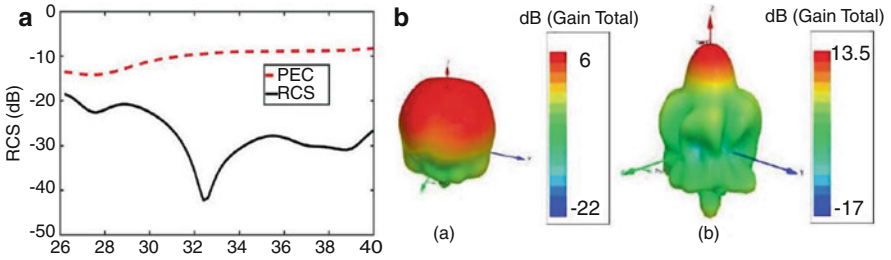

Fig. 14 (**a**) RCS with normal incidence (Akbari et al. 2018) (**b**) 3-D gain without and with PCS at 29 GHz (Akbari et al. 2018)

structure helps in RCS reduction BW by almost 35%. The maximum RCS reduction has been observed around 32 GHz as described in Fig. 14a and an enhancement of gain of 7.5 dB with the introduction of PCS as illustrated in Fig. 14b.

Conclusion and Future Work

Frequency Selective Surface (FSS) plays an important role especially for circularly polarized (CP) millimeter-wave (MMW) frequency antenna. FSS helps in enhancing the antenna gain at MMW frequency where losses are high by concentrating the electromagnetic waves in the broadside direction. It also acts as a linear to circular polarizer when placed at the top of the ground plane thus helping in achieving CP antennas when LP waves are passed through it. Further RCS reduction using a Fabry-Perot cavity (FPC) antenna based on the FSS superstrate helps in maximizing the gain as well. Nevertheless, the FSS based polarizer wall helps in spatial mutual coupling reduction which is an added advantage when dealing with MMW CP MIMO antenna.

In the future, the work can be extended by exploring surfaces based on the FSS technology such as coding diffuse surfaces, metasurface-based LP conversion using grounded vias, and cross-polarization rotation using reflective anisotropic surfaces.

Polarizers based on FSS having different characteristics along with active components create a huge opportunity for future research.

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Metamaterial-FSS Antenna for Microwave Image Sensing

17

Athira Raveendran and Sujith Raman

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Abstract

This chapter discusses the use of metamaterials and frequency selective surfaces for improving the performance of microwave or millimeter wave image-sensing antennas. The selection of a particular image sensing antenna system mainly depends on the application and required quality of the reconstructed images. Different metamaterial-frequency-selective surfaces-based antenna-designing approaches for image sensing are presented here. Conventional microwave-imaging schemes employ mechanically scanning antennas or electronic phased arrays to raster scan a given scene of interest. The use of moving parts or mechanical actuators increases the image data acquisition time in mechanically steerable antenna arrays, whereas the presence of phase shift control circuitry and power amplifiers in electronic phased arrays increases system power consumption. To circumvent these problems, metasurface antennas which are capable of illuminating spatially diverse radiation patterns are widely studied in recent years

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due to its multiplexing nature, less power consumption, planar and compact architecture, and fast image acquisition time. The illumination of a target with diverse radiation patterns reduces the number of measurement modes, while it increases the complexity of image reconstruction algorithm. But the advanced signal-processing capabilities of the modern high-performance computers make the image processing practical in real-time applications. Therefore, these metasurface antennas provide a cost-effective imaging solution. The various metamaterial-frequency-selective surfaces-based antennas described in this chapter have promising applications in security screening at airports, medical image screening, through wall imaging, nondestructive evaluation of construction structures, etc.

Keywords

Antennas · Metamaterial · Microwave image sensing

Introduction

Microwave or millimeter imaging has a plethora of applications ranging from security screening to biomedical imaging. Microwaves are relatively safer for human exposure due to their nonionizing nature, whereas often, exposure to high-energy X-rays during medical imaging can damage living tissues. In addition, microwaves are transparent to many optically opaque materials such as clothing, plastic, ceramics, concrete, etc. (Pastorino 2010). Therefore, microwave imaging can be used to detect concealed weapons and explosives at security screening, through wall imaging, to detect the presence of terrorists in military operations and hidden crevasse in glacier terrain (Gopikrishnan et al. 2020). Metamaterials exhibit anomalous electromagnetic properties which are not readily available in nature. The capabilities of metamaterials or frequency selective surfaces to perform wavefront transformation gave new perspectives for many emerging fields of research such as imaging beyond diffraction limit, enhancement of antenna performance, reduction of radar cross section, electromagnetic cloaking, etc. (Raveendran et al. 2019). A large body of literature is dedicated to the enhancement of radiation characteristics of antennas using artificially engineered materials such as metamaterials, frequency selective surfaces, electronic band gap structures, etc. The advanced signal-processing capabilities of modern computers have led to the realization of compuational microwave imaging. This imaging scheme uses low-profile simple metasurface antennas which illuminate an entire imaging scene with spatially diverse complex patterns. This chapter discusses the recent developments in metamaterial-frequency selective surfaces (FSS)-based antennas for imaging-sensing applications, and its advantages over antennas used in conventional microwave-imaging modalities such as mechanically scanning antennas, electronic phased arrays, and synthetic aperture radar (SAR).

This chapter is structured as follows: First, we give an overview of microwave image sensing and its various applications in different fields of science and technology. Next, we describe the role and significance of antennas in a microwave-imaging system. Then, advantages of metamaterial-frequency-selective surfaces-based image-sensing antennas in comparison to antennas used in conventional imaging modalities are discussed. Different types of metamaterial-FSS antenna designs for image-sensing application are described in the later sections. In addition, there is a brief overview of the capability of metamaterials in subdiffraction imaging or free-space imaging beyond classical diffraction limit. Finally, we conclude the chapter by giving a brief summary.

Overview of Microwave Image Sensing

The aim of microwave image sensing is to sense the physical characteristics of a given scene and reconstruct images using microwave and millimeter waves. The frequency range of microwaves is from 300 MHz to 30 GHz (with corresponding wavelength 1000 mm–10 mm) in the electromagnetic spectrum while millimeter waves span from 30 GHz to 300 GHz (with corresponding wavelength 10 mm–1 mm). One of the main advantages of using microwave-imaging technique is that microwaves are transparent to many optically opaque materials such as textile materials, ceramics, plastics, rubber, concrete, wood, etc. The energy associated with electromagnetic waves is given by $E = h\nu$, where h is the Planck's constant and ν is the frequency of radiation. In the microwave spectrum, radiation energy is approximately from 1.24×10^{-6} eV to 1.24×10^{-3} eV, which is lower than the energy required for breaking molecular bonds (Ida and Meyendorf 2019). Therefore, microwaves are nonionizing radiations and are relatively safe for human exposure in comparison to high-energy X-rays. These features of microwave-imaging technique are responsible for its wide attention and growing demands in the field of science, biomedical engineering, defense, and industries. Microwave imaging is a promising diagnostic tool to detect tumors in medical screening. This is mainly carried out based on effect of electromagnetic wave propagation in different dielectric media. The dielectric properties of healthy living tissues vary from those of cancerous tissues. The medical image shows high contrast between normal and malignant tissues due to the variation in their dielectric properties. In civil engineering and industrial applications, microwave imaging is mainly used for nondestructive testing (NDT) such as evaluation of structural integrity in construction structures, crack detection, surface corrosion, flaws detection, porosity of ceramics, etc. (Pastorino 2010). Since microwaves can penetrate through optically opaque materials such as clothing, microwave imaging is used in security screening at airports to detect concealed weapons and to identify hidden crevasse in glacier terrain (Gopikrishnan et al. 2020). Through wall radar imaging aids in finding survivors who are trapped underground during rescue operation and to detect explosives and terrorists inside a building.

There are mainly two types of microwave-imaging methods: passive imaging and active imaging. Every object emits and reflects radiations based on its characteristic emissivity (Yujiri et al. 2003). In passive microwave imaging, the characteristic emissions and reflected background radiations of a given scene are collected by a receiver to form images as shown in Fig. 1a. This method does not require illumination of the scene by additional electromagnetic source. Thus, passive microwave image is a function of the emissivity and reflectivity of an imaged object with respect to the frequency of radiation (Yujiri et al. 2003). This method provides better quality images in outdoor environment due to low-background radiation temperature of sky. Whereas in indoor applications, the image contrast degrades due to the high-background radiation temperature. Hence, high-performance detectors with cooling system are necessary to achieve better signal to noise ratio which in turn increases the overall cost of the system (Yujiri et al. 2003). Passive microwave-imaging systems detect incoherent radiations from the object, therefore optical elements such as lenses are required to map all parts of the object to a two-dimensional image plane (Hunt et al. 2014). Moreover, it fails to provide depth information about the imaged object. In active microwave imaging, targeted object is illuminated by a microwave source and the reflected/scattered or transmitted fields are detected using receiving antenna as illustrated in Fig. 1b. The image reconstruction is accomplished by applying appropriate image-retrieval algorithm on the collected data. Active imaging systems can detect the radiation coherently which allows the measurement of both the amplitude and phase of the reflected or scattered signal (Hunt et al. 2014). These systems are based on reflection or attenuation of interrogating signal by the object under test. The reflectivity of targeted object under illumination is mapped in reflection-based imaging scheme, while absorption and attenuation of interrogated signal through the target is characterized in transmission imaging setup. Active imaging systems provide high-radiometric contrast images with high-dynamic range in comparison to passive microwave-imaging systems (Yujiri et al. 2003).

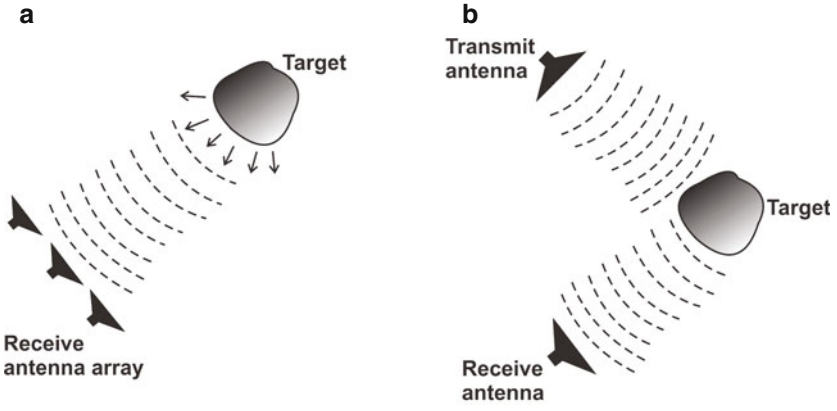

Fig. 1 Microwave-imaging methods: (**a**) passive imaging, (**b**) active imaging

The imaging performance of a system relies on the system resolution. Resolution may be defined as the ability to distinguish two closely spaced objects. In general, aperture resolution limits of a microwave-imaging system are described in terms of range, azimuth cross-range, and elevation cross-range resolutions. The azimuth and elevation cross range-resolution which is also known as lateral resolution is a measure of two-dimensional resolution ability of an imaging system. Range resolution represents the smallest separation between two objects that the imaging system can distinguish in the range direction. It is a function of the transmitted waveform and is inversely proportional to the signal bandwidth (Ahmed et al. 2012). To achieve three-dimensional (3D) image reconstruction of a target, a two-dimensional (2D) aperture needs to be sampled using a broadband measurement signal of bandwidth (B), at each selected transmit-receive combination. The lateral resolution ($\delta_{x,y}$) depends on the spatial extension of the aperture and the distance between object and aperture (L), and it is given by (Ahmed et al. 2012)

$$\delta_{x,y} \approx \frac{\lambda}{D_{xy}} L \tag{1}$$

where λ represents the wavelength of radiation and $D_{x,y}$ the aperture dimension in the corresponding direction. Since the size of the aperture is inversely proportional to lateral resolution, a very large aperture is required to achieve desired resolution. The range resolution (δ_z) is given by (Ahmed et al. 2012)

$$\delta_z \approx \frac{c}{2B} \tag{2}$$

where c denotes the speed of electromagnetic wave propagation. Larger the signal bandwidth is, higher the resolution in range direction. But in practical applications, it is challenging to obtain wider signal bandwidth due to the limitations on oscillators, mixers, and amplifiers used in the system.

To study the penetration of electromagnetic waves inside the object under investigation, consider the propagation of plane waves in a homogeneous medium of permittivity $\varepsilon = \varepsilon' - j\varepsilon''$ and permeability μ. Then, the attenuation constant (α) of medium can be expressed as (Pastorino 2010),

$$\alpha = \omega \left\{ \frac{\mu \varepsilon'}{2} \left[\sqrt{1 + \left(\frac{\varepsilon^n}{\varepsilon'} \right)^2} - 1 \right] \right\}^{1/2} \tag{3}$$

$$\delta = \frac{1}{\alpha} \tag{4}$$

where ω is the angular frequency of the signal and δ denotes the penetration depth or skin depth which is a measure of distance that electromagnetic waves can propagate inside the medium. For a lossless medium ($\alpha = 0$), fields are not attenuated inside the medium. But in case of perfect conductors (conductivity, $\sigma >> \omega\varepsilon$), penetration depth is nearly zero. The fields attenuate exponentially inside a medium with $\alpha > 0$.

Role of Antenna in Microwave Image Sensing

Antenna is an integral part of a microwave-imaging system, and its optimal performance is a crucial factor in determining the quality of generated images. Microwave or millimeter imaging is performed by scanning an antenna or antenna array over a target under investigation. As the lateral resolution of the image is mainly dependent on the size of the aperture, a large aperture is formed by mechanically scanning an antenna or using a huge multistatic array or by synthesizing large imaging aperture. The spatial sampling of a given scene can be done either by mechanical raster-scanning approaches or by switching transmit and receive antenna elements of a multistatic array electronically (Ahmed et al. 2012). In active imaging approach, a two-dimensional image aperture can be sampled using two array architectures: monostatic and multistatic array arrangement. In monostatic array architecture, a target is illuminated sequentially by every transceiver or by a single transceiver moving around the target (Ahmed et al. 2012). Each antenna element in the 2D aperture illuminates the target and receives the reflected/scattered signal at the same observation point.

Figure 2 shows a monostatic imaging scheme, in which black dotted lines depict the mechanical scanning path of the transmitting/receiving antenna. Even though the use of a single transmitting/receiving antenna results in a single transmit/receive channel with reduced complexity, this method slows down the image data acquisition speed. The data acquisition speed can be improved by electronic switching of a large number of transmitting and receiving antennas. But it results in a large number of channels and increases technical complexity of the system. For a monostatic system array, if we neglect the effect of free-space attenuation, the desired reflectivity distribution $o(\vec{r})$ of the object under test is given by (Ahmed et al. 2012)

$$o\left(\vec{r}\right) = \sum_{N_\omega}\sum_{N_A} s\left(\vec{r}_A, \omega\right) e^{j2\left(\frac{\omega}{c_0}\right)R} \tag{5}$$

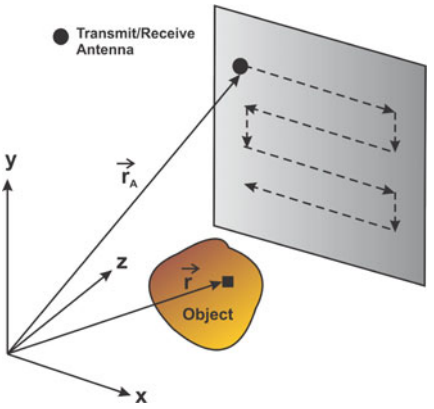

Fig. 2 Monostatic array arrangement for microwave imaging. (Modified and redrawn using the data from (Ahmed et al. 2012))

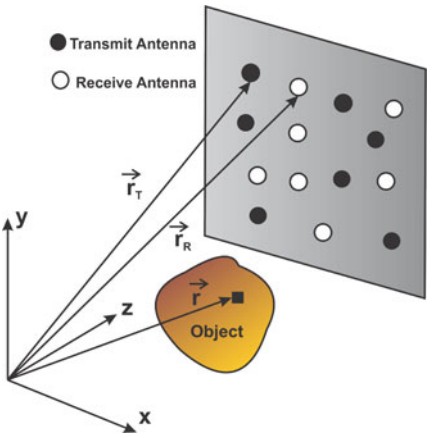

Fig. 3 Multistatic array arrangement for microwave imaging. (Modified and redrawn using the data from (Ahmed et al. 2012))

where $s\left(\vec{r}_A, \omega\right)$ represents the complex signal received at location $\vec{r}_A$ and ω is the angular frequency. The distance between desired voxel position $\vec{r}$ and the respective position of the antenna $\vec{r}_A$ is denoted by R which is given by $R = \mid \vec{r}_A - \vec{r} \mid$.

Figure 3 illustrates a multistatic array arrangement in which spatially distributed multiple antenna elements sample the image aperture. The object to be inspected is illuminated by the transmitting antennas, and the reflected waves are detected coherently by all the receiving antenna elements. Therefore, a less number of transmit/receive channels are required for the same number of measurements in comparison to an equivalent monostatic array. The focusing in a multistatic array arrangement for image reconstruction can be expressed as (Ahmed et al. 2012)

$$o\left(\vec{r}\right) = \sum_{N_\omega} \sum_{N_T} \sum_{N_R} s\left(\vec{r}_T, \vec{r}_R, \omega\right) e^{j2\left(\frac{\omega}{c_0}\right) R_T R_R} \tag{6}$$

where $R_T = \mid \vec{r}_T - \vec{r} \mid$ is the distance between transmitting antenna and the position of the desired voxel and $R_R = \mid \vec{r}_R - \vec{r} \mid$ denotes the distance between the position of receiving antenna and the position of desired voxel. The image reconstruction can be implemented with methods based on Fast Fourier Transformation (FFT). In multistatic imaging, a minimum half wavelength distance is maintained between individual elements of array in order to avoid ambiguities in the image (Ahmed et al. 2012). Furthermore, multistatic imaging method supports parallel data acquisition which helps in real-time imaging. If the object under test is located in the far field of antenna array, transmitted and reflected waves can be approximated as plane waves. This simplifies the reconstruction formulas (5) and (6). On contrary, if the object is in the near field of antenna array, the transmitted or reflected fields are considered as spherical waves (Ahmed et al. 2012).

The scanning of a given scene at Nyquist rate can be performed either mechanically or electronically. Mechanically scanning systems have high-image data acquisition time due to the limitations in the achievable scanning speed and accurate positioning of the antenna. Beamforming technology is a powerful technique to enhance the performance of antennas in an imaging system. The scanning of a directive radiation beam across a given scene can be achieved by controlling the amplitude and phase of every individual elements of the antenna array (Ahmed et al. 2012). The flexible steering of electromagnetic waves in the desired direction can be performed in two ways: analog beamforming and digital beamforming.

The architectures of analog beamforming imager and digital beamforming imager in receiving mode are shown in Figs. 4 and 5, respectively. In an analog beamforming imager, analog devices such as phase shifters, amplifiers are connected to each antenna elements in the array as shown in Fig. 4. The received signals at each antenna elements are combined, and a weighted sum of these signals is then converted into digital format for further digital processing of the data for image formation. This causes a reduction in signal dimensionality from N (number of antenna elements in the array) to one (Hans 1987). On the other hand, in a digital beamforming system shown in Fig. 5, the reflected signals detected coherently at each antenna element are digitalized and stored. Then, a proper calibration is performed to compensate for the transmit/receive transfer functions, and complex correction factors are applied to the data to nullify the effect of free space transfer function. Finally, all the received signals are summed coherently to form the image. Thus, this approach preserves all the information available at the array aperture, and

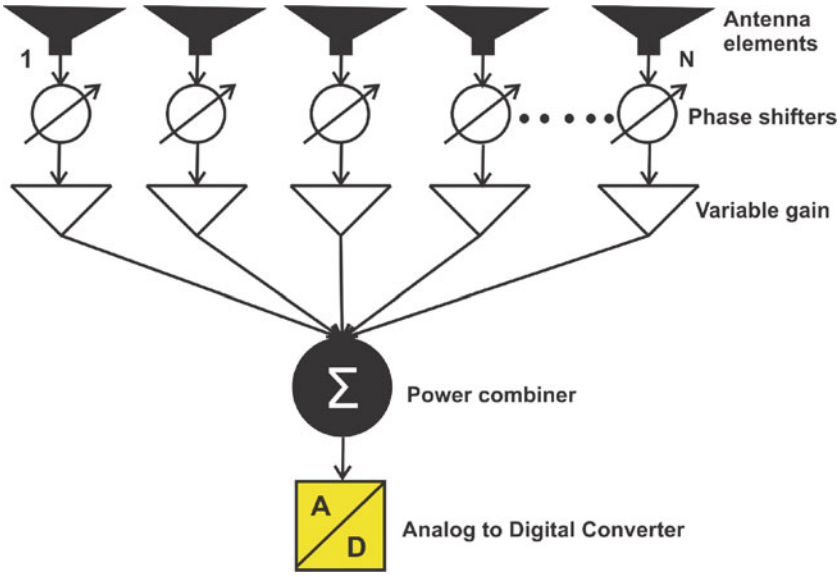

Fig. 4 Architecture of analog beamforming imager in receiving mode. (Modified and redrawn using the data from (Ahmed et al. 2012))

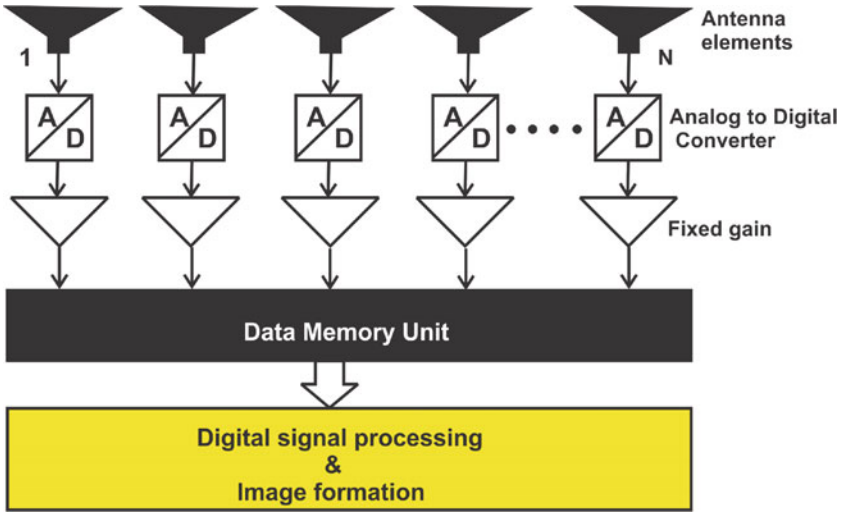

Fig. 5 Architecture of digital beamforming imager in receiving mode. (Modified and redrawn using the data from (Ahmed et al. 2012))

this high level of information can be used to generate multiple images by applying different weighted amplitude to the same raw data (Ahmed et al. 2012).

The image reconstruction speed strongly depends on the measurement time which is related to the selected beamforming antenna array architecture and the complexity of the image reconstruction algorithm. In analog beamforming imager, the measurement time required for image data acquisition relies on the number of scanned samples or voxels and the scanning time per voxel which is determined by intermediate frequency bandwidth and switching speed of array elements (Ahmed et al. 2012). Whereas in digital beamforming system, measurement time depends on the number of sequential measurements, transmitter-switching speed, intermediate frequency bandwidth, RF bandwidth, and the required image resolution and dynamic range (Ahmed et al. 2012). Even though digital beamforming system requires high-digital signal processing capability and the use of complex imaging algorithm in order to handle enormous recorded data, high-performance modern computers can perform these tasks to generate images.

The selection of antennas used for a microwave-imaging sensing system is associated with the following five principle requirements of the imaging system:

- High spatial and dynamic range resolution
- Desired penetration level of the electromagnetic waves inside the object under investigation
- Less system complexity
- Compactness of system and portability
- Cost of manufacture

The transmitting and receiving antennas are required to achieve perfect impedance matching with the medium of wave propagation and other RF circuit elements connected to the antenna elements in the array to minimize the losses in the system. The spatial resolution of an image is highly dependent on the effective illumination of target by the proper positioning of transmitting and receiving antennas in the aperture. Broadband measurement signals are necessary to accomplish excellent range resolution, and therefore the antennas should support wider bandwidth of operation. Another important criterion in the image-sensing antenna designing is that the antenna should exhibit stable phase centers over the beamwidth and bandwidth of operation (Ahmed et al. 2012). This is because image reconstruction algorithm applies spherical wave front approximation in the processing of received signal, and any variations in the approximated values will result in poor quality images. Phase center can be defined as an imaginary point which acts the source center of a radiated spherical wave front, and the phase of the signal is equal at any point on the sphere. Polarization purity of the antenna is of center importance in case of polarimetric imaging. Polarimetric imaging is used to collect additional information of a targeted scene by extracting polarization information from the scattered/reflected signals. It has many potentials in military and surveillance applications. The operating frequency of an imaging antenna varies from application to application. For example, antennas operating in lower microwave frequency bands are employed in the detection of deeply buried objects (Yektakhah et al. 2019) whereas millimeter and terahertz waves are preferred to collect finer physical details of a scene of interest (Gollub et al. 2017). Dipoles (Ebihara et al. 2020), horn antennas (Amineh et al. 2010), open waveguide antennas (Mirzaee et al. 2021), microstrip patch (Aguilar et al. 2013), and slotted array antennas (Kharkovsky et al. 2013) are the common antenna types used in microwave image sensing. In medical screening based on microwave tomography, printed directional ultrawideband antennas are widely employed (Islam et al. 2019). Intense research on the use of frequency-selective surfaces (FSS) and metamaterials for radiation characteristics enhancement has led to the development of metasurface-FSS-based antennas for microwave image-sensing applications. In the next section, recent developments in the field of metasurfaces and FSS antennas and their advantages in comparison to conventional mechanically or electronically scanning antenna arrays are discussed.

Metamaterial-FSS Antennas for Microwave Image Sensing

Metamaterials are widely known for their exceptional electromagnetic properties such as negative refractive index, reverse Doppler effect, and backward propagation of phase front (Raveendran et al. 2019). They are composed of artificial structural units of dimensions which are considerably smaller than the wavelength of incident electromagnetic waves. They exhibit exotic electromagnetic properties which are not readily found in their constituent elements as well as in natural materials. The precise

shape, size, geometry, orientation, and arrangement of the constituent elements give them anomalous properties which are capable of manipulating electromagnetic waves by absorbing, reflecting, guiding, or bending waves.

The macroscopic response of a metamaterial is described by effective medium theory and characterized in terms of material properties such as permittivity, permeability, and refractive index (Raveendran et al. 2019). A periodic, aperiodic, or quasi periodic arrangement of metamaterial elements on a planar surface is called metasurface. Frequency selective surfaces (FSS) are another class of periodic structures which are widely investigated for their spatial filtering properties. They consist of subwavelength constituent structures, and their electromagnetic response depends on the frequency, polarization, and angle of incidence of the impinging electromagnetic wave (Munk 2005). The applicability of these artificially engineered surfaces in improving the performance of antennas is a highly investigated field of research, and a plethora of works has been reported in this domain. Based on the design and mode of operation, FSS-metamaterial antennas used for image sensing can be classified as follows:

1. Image-sensing metasurface antennas
2. Digital or programmable metasurface antennas
3. FSS- or metamaterial-loaded image-sensing antennas

In general, the ability of metasurfaces or frequency selective surfaces in modulating the amplitude and phase of electromagnetic waves is utilized in metasurface- or FSS-based antenna design to obtain a desired radiation pattern. Wideband antennas with well-defined radiating beams are used in conventional microwave imaging. In the first antenna design approach, there is a drastic difference in the antenna radiation characteristics and imaging methodology in comparison with the conventional imaging method. A metasurface is excited by guided/surface modes to generate spatially diverse complex radiation patterns. The information content of a given scene is multiplexed into a few measurements (Imani et al. 2020). A digital metasurface capable for modulating the impinging electromagnetic waves is the vital component in a programmable metasurface antenna. Various sets of metamaterial elements are turned ON sequentially to generate diverse radiation patterns or test modes Li et al. (2019). In the context of FSS or metamaterial loaded image-sensing antennas, a conventional antenna (printed vivaldi or antipodal vivaldi antenna, printed monopole antenna, etc.) is loaded with metamaterial or FSS elements to improve its radiation characteristics such as gain, directivity, bandwidth, front to back ratio, etc. (Islam et al. 2019). The metamaterial- or FSS-loaded high-performance antenna with directional beam is then used to illuminate a target under test. Traditional mechanical/electronic raster-scanning method or multistatic array configuration is used to illuminate a scene of interest, and back scattered signals are processed to reconstruct images. A detail study of FSS-metamaterial-based antennas for microwave image sensing is given in the following section.

Image-Sensing Metasurface Antennas

A metasurface antenna is composed of an electrically large (as compared to the wavelength of operation) metasurface decorated with metamaterials elements with no inherent periodicity (Imani et al. 2020). It is fed by a guided mode within a feeding structure such as microstrip line, printed cavity, or waveguide cavity. The guided wave interacts with the metamaterial elements and couples energy from the feed into free space forming complex radiation pattern. The coupling of guided wave energy into the metasurface is governed by characteristic resonance of each metamaterial element which in turn depends on its geometrical properties. As the driving frequency is swept over measurement bandwidth, different metamaterial elements on the aperture get excited and generate spatially diverse complex radiation patterns (Hunt et al. 2014). The principle of metasurface microwave imaging is that a metasurface antenna sequentially illuminates the entire scene of interest with spatially diverse radiation patterns instead of electronically or mechanically scanning a scene of interest. And the information content of a given scene is encoded into a few simple measurement modes in the frequency domain (Hunt et al. 2014). Hence, metasurface antenna-imaging approach does not require any mechanical moving parts or actuators as compared to traditional mechanically scanning antennas or complex feeding structures and phase shifters as in case of a phased antenna array system.

Based on the feeding system, an image-sensing metasurface antenna is divided into two types:

(i) Waveguide-based metasurface antennas
(ii) Cavity-backed metasurface antennas

A waveguide-based metasurface antenna is fed by a guided mode propagating along a waveguide structure such as microstrip lines (Hunt et al. 2013) or parallel plate waveguides (Lipworth et al. 2013). On the other hand, a cavity-backed metasurface antenna is excited by a guided mode inside a hollow metallic cavity (Marks et al. 2017) or printed circuit cavity formed by closely spaced metallic vias around the perimeter of the cavity (F. Imani et al. 2016). Figure 6 illustrates a waveguide-based metasurface antenna and a cavity-backed metasurface antenna. In general, metasurfaces are designed based on a periodic arrangement of identical structural elements or elements with a gradual variation in geometrical dimensions. Even though this approach simplifies the design and electromagnetic modeling of metasurfaces, it restricts the number of possible diverse patterns in case of a metasurface antenna. Therefore, inherent periodicity of metamaterial elements is not maintained in metasurface image-sensing antenna design in order to accomplish a maximum number of spatially diverse patterns (Imani et al. 2020). In other words, the number of measurement modes or spatially diverse radiation patterns is proportional to the number of distinct metamaterials on the aperture. Hence, the geometrical properties of each metamaterial element are selected to be different from its neighboring elements.

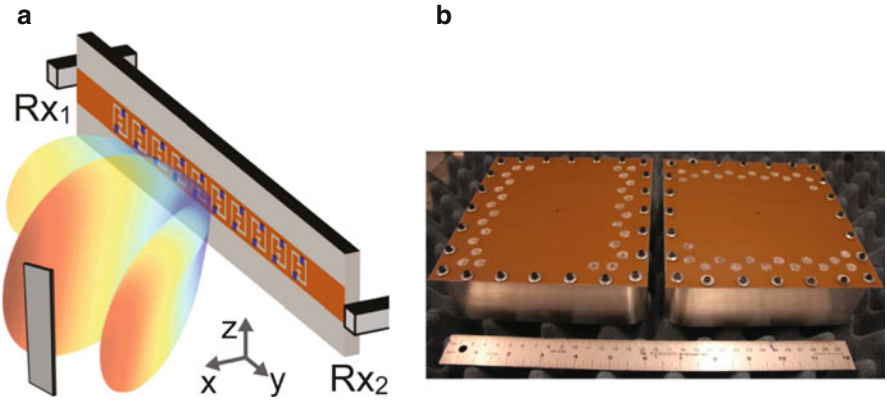

Fig. 6 (**a**) Waveguide-based metasurface antenna. (Adapted with permission from (Sleasman et al. 2015) © The AIP Publishing); (**b**) cavity-backed metasurface antenna. (Adapted with permission from (Marks et al. 2017) © The Optical Society)

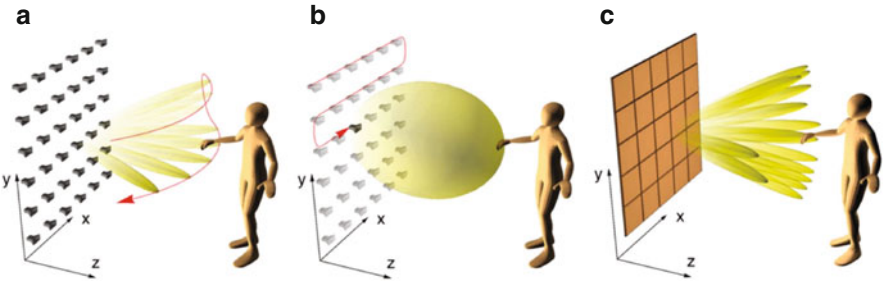

Fig. 7 Different coherent microwave-imaging modalities: (**a**) Phased array ESA technology; (**b**) synthetic aperture radar (SAR); and (**c**) metamaterial holographic imaging. (Adapted with permission from (Hunt et al. 2014) © The Optical Society)

Figure 7 illustrates different coherent microwave-imaging schemes. Phased array technology and synthetic aperture radar (SAR) are the commonly used microwave-imaging modalities (Hunt et al. 2014). Electronically scanned array (ESA) technology realizes image sensing by raster scanning a highly directive beam over scene of interest at Nyquist rate as shown in Fig. 7a. The phase shifters connected to each element of the antenna array contribute insertion loss in the system which are compensated by the use of power amplifiers (Yurduseven et al. 2018). This results in high-power consumption, complex system architecture, and high cost. In SAR, a single stand-alone transceiver illuminates a target with a comparatively broad radiation beam, and mechanically raster scans the entire scene at various transceiver locations as depicted in Fig. 7b. The resolution of the reconstructed image improves with the increase in size of the synthesized aperture (Hunt et al. 2014). For a SAR system, mechanical scanning speed and the necessity of accurate placement of antennas pose limitations in the image data acquisition time. These factors restrict the widespread access and usage of

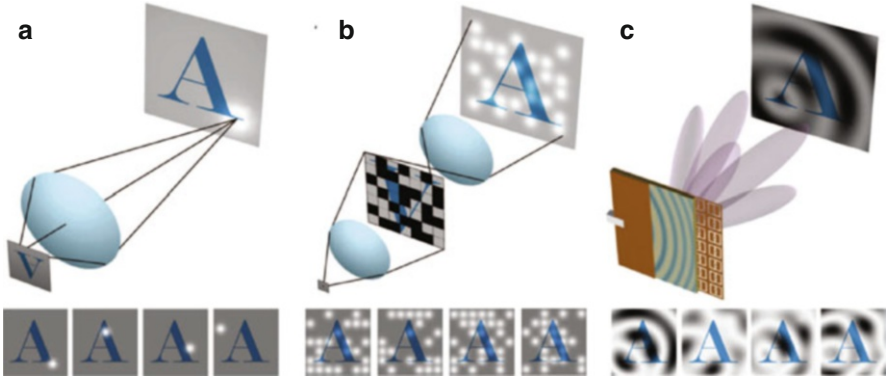

Fig. 8 Comparison of conventional lens-imaging schemes and metasurface antenna-based imaging scheme: (**a**) conventional lens-based imaging; (**b**) lens-based compressive imaging; and (**c**) microwave metamaterial-based imaging. "Adapted with permission from (Ref.) © The American Association for the Advancement of Science"

traditional microwave-imaging techniques and triggered intense research to find a more cost-effective system with improved capabilities.

Figure 8 shows a comparison of optical lens-based imaging schemes and metasurface antenna-based computational imaging. In conventional imaging, measurement modes are formed by optical lens, and each mode contains highly specific information about a point in a given scene (Hunt et al. 2013). Each measurement mode maps one-to-one relationship between each point in the scene of interest and image plane as shown in Fig. 8a. All modes can be collected simultaneously using an array of detectors or can be recorded sequentially in case of single pixel scheme. Compressive imaging scheme uses two lenses and a random mask to form incoherent measurement modes as illustrated in Fig. 8b. These measurement modes sample the image plane and capture the respective region information (Hunt et al. 2013). On the contrary, microwave metamaterial imager eliminates the need of expensive lenses to form measurement modes. The metaimager fed by a microwave source produces random spatially diverse illumination patterns to sample the entire scene of interest as depicted in Fig. 8c. The image data collected over the operating frequency bandwidth is multiplexed into a small set of measurements. The growing interest behind the use of metamaterials-based radiation pattern synthesis for imaging applications is mainly driven by the following facts.

- Cost-effective due to the absence of phase shifters and complex feeding networks (as used in electronically scanned arrays) and expensive optical elements (as required in conventional imaging technique)
- Less system complexity
- Low-power consumption in comparison to phase array technology
- Low-data acquisition time due to the avoidance of mechanical raster-scanning methods
- Planar and compact system architecture

For a given aperture, diffraction-limited contents of a scene can be sampled into a finite number of pixels or voxels. The information content associated with each voxel is accessed by either scanning (mechanical or electronic) or using multistatic array of detectors over the image plane (Lipworth et al. 2013). In the context of metasurface antennas, a measurement mode represents a distinct radiation pattern illuminating the entire scene of interest and the corresponding fields radiated from the transmission aperture. The quantity of measurements for a diffraction-limited resolution imaging is determined by space bandwidth product (SBP), M of the system (Lipworth et al. 2013). The number of distinct illumination patterns or measurement modes required to perform diffraction-limited imaging of a given scene is approximately (Hunt et al. 2014),

$$N_m = \left(\frac{D_{\text{eff}}}{\lambda}\right)^2 \tag{7}$$

where D_{eff} denotes the effective aperture length.

To perform N measurements, N modes or distinct illumination patterns are generated by sweeping through N field distribution across transmission aperture. Assuming sampling of a scene and aperture fields at discrete intervals, the imaging equation commonly used in computational imaging is given by (Lipworth et al. 2013),

$$g = \mathrm{Hf} \tag{8}$$

or

$$\begin{bmatrix} g1 \\ g2 \\ \vdots \\ gN \end{bmatrix} = \begin{bmatrix} h_{11}\, h_{12} \cdots h_{1M} \\ h_{21}\, h_{22} \cdots h_{2M} \\ \vdots \quad \vdots \quad \ddots \quad \vdots \\ h_{N1} \cdots \cdots h_{NM} \end{bmatrix} \begin{bmatrix} f_1 \\ f_2 \\ \vdots \\ f_M \end{bmatrix} \tag{9}$$

where **H** represents an $N \times M$ measurement matrix or sensing matrix, **g** is the received signal, and **f** is the reflectivity of the scene. When we neglect the effect of noise, maximum diffraction-limited reflectivity (**f**) of a given scene can be retrieved by simple matrix inversion (Lipworth et al. 2013). But in practical imaging systems, the presence of system noise, and the overlapping of measurement modes or correlated radiation patterns, significantly reduces the number of possible distinct measurement modes. Thus, the number of maximum achievable measurements N is less than the total number of voxels in the scene. As a result, in practice, **H** is not a square matrix and the direct matrix inversion is not possible. Then, we can rewrite the imaging equation as (Hunt et al. 2014),

$$\mathbf{g} = \mathbf{Hf} + \mathbf{n} \tag{10}$$

where **n** denotes noise of the system. Under first Born approximation, the radiation pattern of an antenna is related to the measurement matrix as (Lipworth et al. 2015),

$$H_{ij} = k_0 \vec{E}_i^R\left(\vec{r}_j\right)\vec{E}_i^T\left(\vec{r}_j\right) \tag{11}$$

where k_0 denotes a normalizing constant, $\vec{E}_i^R\left(\vec{r}_j\right)$ and $\vec{E}_i^T\left(\vec{r}_j\right)$ are the fields radiated by receiving and transmission aperture at scene location $\vec{r}_j$, respectively. The index i counts measurements (i.e., changes in radiation pattern, frequency, antenna position, etc.), while j refers to voxels that sample the scene of interest. Accurate characterization of radiated fields of a metasurface antenna is necessary to determine the measurement matrix **H**. Any errors in the estimation of **H** will degrade the quality of the reconstructed image (Hunt et al. 2014). The mutual coupling between adjacent metamaterial elements and the coupling of fields between guided wave and a large number of metamaterial radiators on the aperture have to be taken into account for the accurate characterization of radiated fields (Hunt et al. 2014). Thus, a full wave simulation of the metamaterial aperture is tedious and highly time-consuming. Hence, near-field scanning of the metamaterial aperture is recommended to measure the radiated field components in real-time applications. It is performed by scanning an open-ended waveguide probe across a plane which is located in the radiated near field and parallel to the metasurface.

Akin to all the other imaging modalities, aperture size and signal bandwidth are the key parameters determining the performance of metasurface antenna-imaging systems. These parameters have already been discussed in the beginning of this chapter. Other system performance parameters which are of primary importance in the context of designing an efficient metasurface antenna with spatially diverse patterns for image sensing applications are the following:

(i) **Signal Fidelity**

Signal fidelity is a measure of accurate preservation of relative amplitude of different signal components in the display output. Metasurface antennas synthesize pseudorandom and complex radiation patterns which necessitate accurate characterization of the antenna in determining sensing matrix **H** (Imani et al. 2020). This could be accomplished by near-field scanning of the metasurface antenna. Moreover, metasurfaces used for pattern synthesis are electrically large in order to reduce the effect of edge diffraction. Therefore, the system is susceptible to misalignment of these metasurface beamformers during imaging and near-field scanning. Hence, the precise calibration of the RF system is a prerequisite, and it can be done by performing imaging of a standard known object or by placing backscatters in the scene of interest.

(ii) **Signal to Noise Ratio**

High signal to noise ratio is a key requirement to obtain high-quality images in any imaging system. A metasurface image-sensing antenna can multiplex image data of an entire scene of interest into a small measurement set. This multiplexing feature of diverse radiation pattern averages the effect of noise in image reconstruction which results in the improved signal to noise ratio. The improvement in signal to noise ratio due to the multiplexing nature is referred to as multiplexing advantage or Fellgett's advantage (Imani et al. 2020).

(iii) **Spatial Sampling**

A precise and uniform sampling of the spatial content of a given scene is crucial in the reconstruction of high-quality images. On the other hand, a nonuniform spatial sampling leads to aliasing in the reconstructed image. Thus, the placement of metamaterial unit cells and their overall density across the imaging aperture have to be optimized in order to ensure a uniform sampling of spatial frequencies to mitigate aliasing (Imani et al. 2020).

(iv) **Quantity of Measurements**

Space bandwidth product of an imaging system describes the information capacity of the system. It is the product of field of view that an imaging system can cover and the spatial frequency bandwidth (Claus et al. 2011). Theoretically, the maximum number of possible measurements of an imaging system is determined by the space bandwidth product of the system. In metasurface holographic imaging systems, the number of possible measurements is proportional to the number of metamaterial elements in the aperture due to the linear superposition of radiated field by them (Imani et al. 2020). Therefore, the performance of an imaging system can be improved by using a metasurface with large number of radiating elements. But the quantity of measurements cannot be increased beyond the number of measurements governed by space bandwidth product.

(v) **Quality of Measurements**

A metasurface antenna produces different pseudorandom and complex radiation patterns with the variation in excitation frequency alone or in combination with the tuning of metamaterials elements. Higher correlation among the illuminated radiation patterns degrades the image quality (Imani et al. 2020). Hence, metamaterial antenna has to be designed with minimum correlation among the radiation patterns.

In the context of designing high-performance metasurface image-sensing antennas, the number of metamaterial radiators is chosen as a trade-off between quantity and quality of measurements, whereas the distribution of radiating metaelements in the aperture is optimized to accomplish a uniform sampling of spatial contents. There are three types metasurface antenna design approaches to synthesize spatially diverse radiation patterns for image-sensing applications.

(i) Frequency diverse metasurface antennas
(ii) Dynamic metasurface antennas
(iii) Hybrid metasurface antennas

Frequency-Diverse Metasurface Antennas

In this configuration, a metasurface containing metamaterial elements, whose resonance frequencies are distributed randomly over a broader frequency band of operation, is fed by a guided mode (Lipworth et al. 2013). The guided waves interact with the metamaterial elements, and meta-aperture generates abundant spatially diverse radiation patterns as the driven frequency is swept over the measurement

bandwidth. The number of possible measurement modes is equal to the number of spatially diverse patterns generated by the metasurface as a function of frequency (Lipworth et al. 2013). Basically, there are two different approaches to design frequency-diverse metasurface antennas:

(i) Frequency diversity via variation in geometrical properties of metamaterial elements
(ii) Frequency diversity via feed diversity

In the first approach, the geometric properties of the metamaterial radiators are optimized to achieve sharp resonances which are randomly distributed over the frequency band of operation. As the excitation signal frequency varies, different sets of metamaterial elements couple energy into free space depending on the geometrical dimensions of respective metamaterial elements. The frequency-diverse metamaterial aperture can be designed in one-dimensional (1D) or two-dimensional (2D) fashion. The first implementation of frequency-diverse metasurface for computation imaging was reported by Hunt et al. in 2013. Figure 9a shows the schematic diagram of the reported one-dimensional metamaterial aperture. It consists of a standard microstrip line loaded with complementary electric-inductor-capacitors (cELCs) metamaterial elements (Hunt et al. 2013). The frequency diversity is achieved by varying the geometrical properties of adjacent metamaterial elements. As the wave propagates through the microstrip line, each of the complementary metamaterial elements couples the waveguide mode to free space radiation.

Thus, the operation of a frequency-diverse metasurface antenna has a close resemblance with conventional leaky wave antennas. But there are two fundamental differences. The first difference in the periodicity of metamaterial elements used in metasurface antennas is very small compared to the wavelength of operation, but such small elements are inefficient in coupling energy in case of leaky wave antenna (Lipworth et al. 2013). Another difference is the use of subwavelength resonators in metaimager in order to achieve transcendent electromagnetic properties and the control of radiation pattern as a function of frequency, whereas the resonating elements in a leaky wave antenna need not exhibit exotic electromagnetic properties. The radiation patterns of the proposed one-dimensional metasurface antenna at 18.5 GHz and 21.8 GHz are depicted in Fig. 9b. The spatial diversity in radiation patterns of frequency-diverse metasurface can be improved by using 2D frequency-diverse metaimagers instead of 1D metasurface antenna. The degree of diverse spatial patterns of a 2D metaimager can be further enhanced by using multiple metasurface antenna panels as shown in Fig. 9c. This bistatic system contains a number of transmitting frequency-diverse metasurface antennas and receiving open-ended waveguide probes (Lipworth et al. 2015). A probe introduced inside the parallel plate waveguide launches cylindrical guided waves, and the complementary resonant metamaterial elements on the plate top radiate energy into free space by simple frequency sweeping (Lipworth et al. 2015). Each metamaterial imager illuminates a target sequentially, and the reflected signals received through low-gain probes are processed to reconstruct an image.

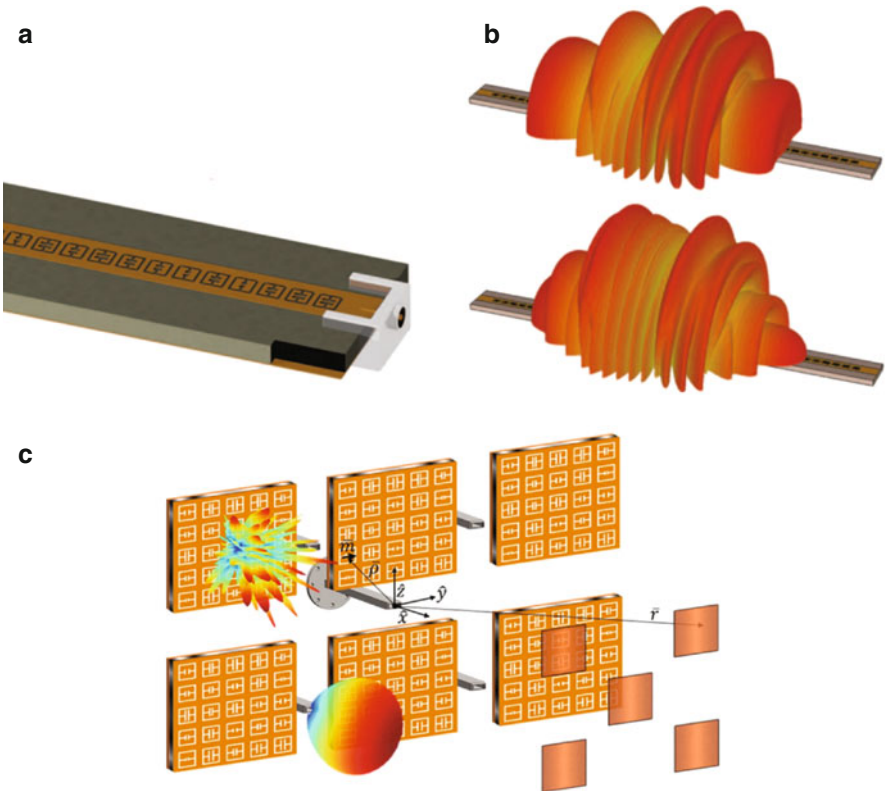

Fig. 9 Frequency-diverse metasurface antenna based on element diversity: (**a**) schematic diagram of 1D frequency-diverse metasurface antenna. "Adapted with permission from (Ref.) © The American Association for the Advancement of Science"; (**b**) radiation pattern at 18.5 GHz (top) and 21.8 GHz (bottom). "Adapted with permission from (Ref.) © The American Association for the Advancement of Science"; and (**c**) bistatic imaging setup using 2D frequency-diverse metasurface antennas. (Adapted with permission from (Lipworth et al. 2015) © The Optical Society)

In the second approach, frequency diversity is accomplished by using spatially diverse guided modes within a feeding structure to excite metamaterial elements (Fromenteze et al. 2015). For example, Fig. 10a shows a cavity-backed frequency-diverse metasurface antenna. It consists of a rectangular waveguide cavity with spherical deformation at one of its corners, and the top side of the cavity is patterned with periodically arranged small holes (Fromenteze et al. 2015). The periodicity of holes should be less than the operating wavelength in order to couple all resonant modes inside the cavity. In addition, the number of modes supported by a cavity is governed by the volume of the cavity. Thus, the size of the cavity is selected as large as possible in all directions with respect to the wavelength of operation. The perturbation of the cavity modes by the incorporation of the spherical deformation results in the generation of distinct frequency-dependent modes (Fromenteze et al. 2015). The energy corresponding to these modes is coupled to the free space via

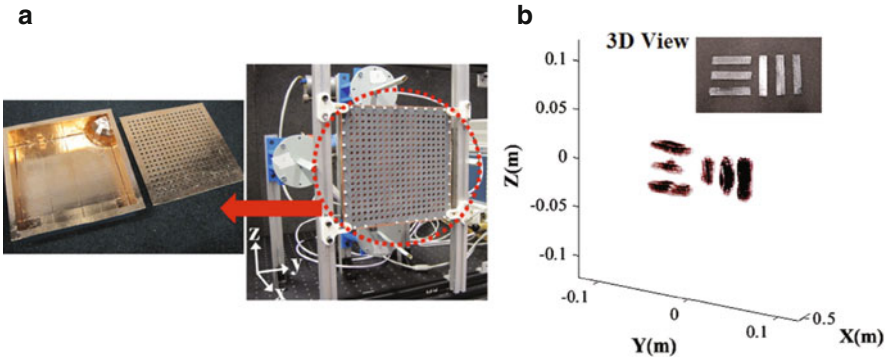

Fig. 10 Frequency-diverse metasurface antenna based on feed diversity: (**a**) cavity-backed metasurface antenna (on left) and measurement setup (on right); (**b**) reconstructed 3D image of a target. (Adapted with permission from (Fromenteze et al. 2015) © The AIP Publishing)

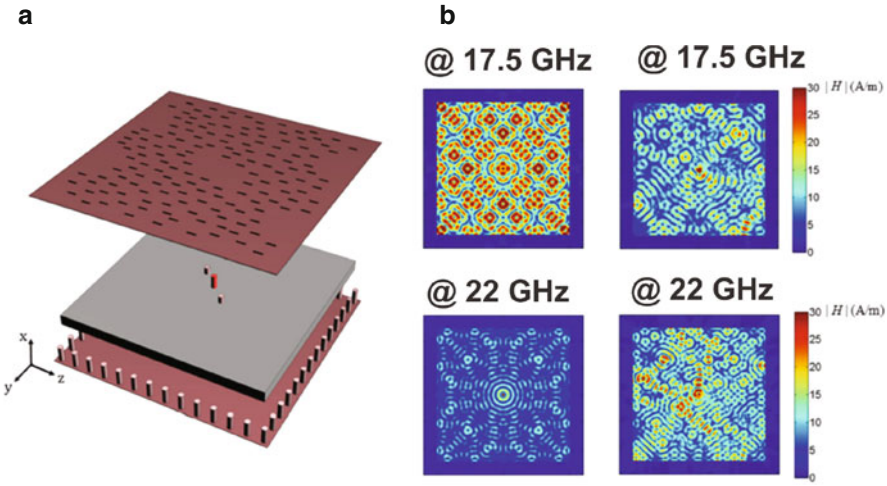

Fig. 11 (**a**) Schematic diagram of a printed cavity with radiating slots; (**b**) different modes generated inside the printed cavity at 17.5 GHz and 22 GHz, respectively (left column shows cavity modes without scattering vias, and right column shows cavity modes with scattering vias). (Adapted with permission from (F. Imani et al. 2016) © The AIP Publishing)

small holes and illuminates a target. Figure 10b shows a 3D reconstructed image of a target using the aforementioned cavity-backed antenna. A printed version of frequency-diverse metasurface antenna fed by frequency-diverse waveguide modes is shown in Fig. 11a. A planar cavity is formed by closely spaced metallic vias arranged around the edges of a substrate metallized on both sides (F. Imani et al. 2016). Narrow rectangular slots randomly distributed over the aperture of the cavity couple the energy of the resonant modes into free-space radiation as the frequency sweeps over the measurement bandwidth. The random field distribution of slots is

opted to reduce correlation between the radiated fields. The two distinct waveguide modes generated inside the printed planar cavity with and without the presence of subwavelength-scattering vias are depicted in Fig. 11b. As we can see, the introduction of scattering vias alters the magnetic field distribution inside the cavity. The field distribution inside the cavity varies as a function of frequency, and the diversity of cavity-resonant modes can be further increased by introducing subwavelength scatters inside the cavity (F. Imani et al. 2016).

A frequency-diverse metasurface antenna illuminates a given scene with spatially diverse radiation patterns by simple frequency sweeping. The information content of the entire scene is encoded into frequency domain, and the scene is reconstructed by applying image reconstruction algorithm. The number of measurement modes or distinct illumination patterns is proportional to the number of metamaterial elements. However, as we increase the number of radiators in the aperture, radiation losses increase which in turn results in low-quality factor and high-spatial correlation. Therefore, in order to obtain as many diverse patterns as possible within the available space bandwidth product, the frequency shift required to select a distinct radiation pattern needs to be reduced (Marks et al. 2017). The amount of frequency shift required to generate a unique radiation pattern is governed by the quality factor of the metasurface antenna. The quality factor (Q-factor) is defined as (Marks et al. 2017)

$$Q = 2\pi f_0 \frac{\text{Stored energy in component}}{\text{Power loss in component}} \tag{12}$$

$$Q = \frac{f_0}{B} \tag{13}$$

where f_0 is the center frequency of excitation signal and B is the operational radiation bandwidth. The amount of frequency shift required to generate a distinct radiation pattern is f_0/Q. For a given quality factor, the number of spatially diverse radiation patterns (N) that an antenna can produce is given by Marks et al. (2017):

$$N \approx \frac{QB}{f_0} \tag{14}$$

Therefore, the number of unique radiation patterns formed by an antenna can be increased by boosting the quality factor. The quality factor can be increased either by increasing stored energy or reducing the losses in the system. The stored energy is directly proportional to the volume of the antennas. So the simple solution is to increase the volume of the antenna. However, as the volume of a dielectric medium-filled metasurface antenna becomes larger, the associated dielectric losses also increase (Marks et al. 2017). Thus, designing a hollow metallic cavity-based metasurface antenna is a practical solution, since this cavity has minimum dielectric losses and high surface to volume ratio (Marks et al. 2017). In general, the quality factor of individual metamaterial resonators and/or that of a cavity feeding the metasurface is optimized as large as possible in order to obtain less correlated modes. Less correlated modes are essential to yield high-quality images. A hollow

cavity-backed metasurface antenna has large radiation bandwidth and high Q factor, correspondingly many spatially diverse radiation patterns. A significant increase in number of measurements can be achieved by using a pair of hollow metallic cavity-backed metasurface antennas. For such a transmit and receive antenna pair, the number of metamaterial radiators on the transmit antenna (N_{tx}) is related to the number of subwavelength radiators on the receive antenna (N_{rx}) as (Marks et al. 2017)

$$N_{tx}N_{rx} \approx \frac{QB}{f_0} \tag{15}$$

Furthermore, it is suggested that the use of multiple feeding points increases the number of measurement modes. And the feed positions are arranged far from each other to ensure less correlation between the guided modes (Hunt et al. 2014). A stringent trade-off between number of metamaterial elements and quality factor is to be maintained to design a high-performance metasurface-imaging system.

Dynamic Metasurface Antennas

The number of uncorrelated spatially diverse illumination patterns generated by a frequency-diverse metasurface antenna is limited by the maximum achievable quality factor of individual metamaterial elements, the number of radiator elements, signal to noise ratio, and the quality of measurements (Imani et al. 2020). For a frequency-diverse antenna, ultrawideband (UWB) operation is crucial in achieving a large number of measurement modes and excellent range resolution. The requirement of a UWB signal source and the associated RF circuitry makes the RF system highly expensive and complex. Moreover, the allocation of ultrawideband spectrum is limited in certain applications due to interference regulation (Imani et al. 2020). To alleviate these constraints, an alternative method based on electronic reconfiguration in the form of dynamic metasurface antenna (DMA) is suggested. Dynamic metasurface antenna has the same basic architecture as that of a frequency-diverse metasurface antenna. The difference is in the presence of semiconductor-switching elements introduced in each metamaterial radiator as shown in Fig. 12a. PIN diodes and varactor diodes are the commonly used switching elements to control the state of each element on the metasurface. PIN diode offers binary modulation, that is, it turns "ON" metamaterial element contributing to radiation or turns "OFF" radiating metamaterial element. It has less degree of freedom due to binary tuning while varactor diodes have more control over tuning of resonant elements (Yurduseven et al. 2018). Varactor diodes can tune the resonant frequency of each subwavelength resonator by gradually varying the applied dc voltage. However, a varactor diode cannot turn "ON" or "OFF" a metamaterial element completely which leads to a gradual attenuation of the guided mode. In addition, PIN diodes are current driven elements with high-power consumption in comparison to voltage-driven varactor diodes. In most of the reported works on dynamic metasurface antennas, PIN diodes are generally preferred due to their simple biasing circuitry compared to the biasing circuitry of varactor diodes which allows the application of different dc levels (Imani

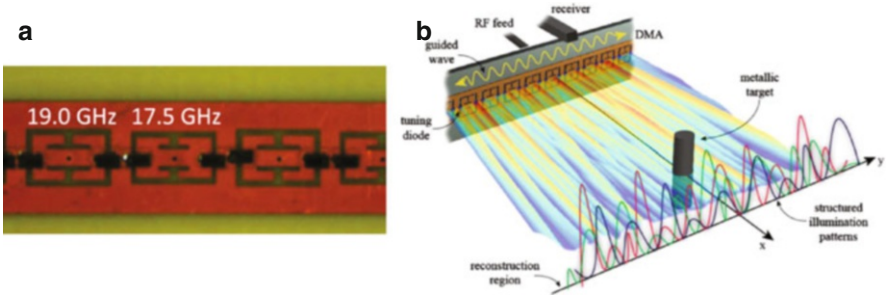

Fig. 12 (**a**) One-dimensional dynamic metasurface antenna loaded with PIN diodes. (Adapted with permission from (Sleasman et al. 2016a) © The Optical Society); (**b**) microwave imaging using 1D-dynamic metasurface antenna. (Adapted with permission from (Diebold et al. 2018) © The Optical Society)

et al. 2020). By using PIN diodes, different subsets of metamaterial elements can be turned "ON" to produce diverse illuminating patterns even at a single frequency as illustrated in Fig. 12b. Thus, the introduction of switching elements in each metamaterial radiator gives more degree of freedom in controlling the properties of each subwavelength radiator which facilitates the synthesis of more distinct radiation patterns. The operating principle of dynamic metasurface antenna is different from that of spatial reconfiguration using metasurface reported in literature (Sievenpiper et al. 2002). In DMA, frequency response of each metamaterial element is controlled individually and is independent of other elements on the metasurface (Sleasman et al. 2016a). On the other hand, a switching or tuning circuitry modulates the effective impedance or reactance of the metasurface in case of a reconfigurable metasurface antenna.

Similar to frequency-diverse metasurface antenna architecture, a dynamic metasurface antenna can also be designed using 1D or 2D metasurface. For example, Fig. 12a shows a one-dimensional DMA where metamaterial elements are patterned on a microstrip transmission line, and each element is loaded with PIN diodes (Sleasman et al. 2016a). A given scene is illuminated with diverse radiation patterns by turning "ON" different sets of metamaterials individually. A two-dimensional DMA reported in Mizrahi et al. (2020) consists of an irregular planar cavity formed by a metallic via cage around the edges of a double-sided copper-clad substrate. The top of the cavity is loaded with a random distribution of metamaterial elements. Each metamaterial radiator is loaded with a PIN diode to enable individual tuning of the element. Random distribution of metamaterial elements on the aperture is utilized to reduce spatial correlation (Mizrahi et al. 2020). As a guided wave propagates along a waveguide (1D- DMA) or within a printed cavity (2D-DMA), the energy of the guided mode decays gradually. Hence, the size of metamaterial aperture, the number of metamaterial elements, and feeding locations are optimized in such a way that every element receives ample energy (Imani et al. 2020). In addition, the ohmic losses of PIN diodes and metamaterial elements under "OFF" state have to be considered in the designing of a dynamic metasurface antenna. Figure 13a illustrates

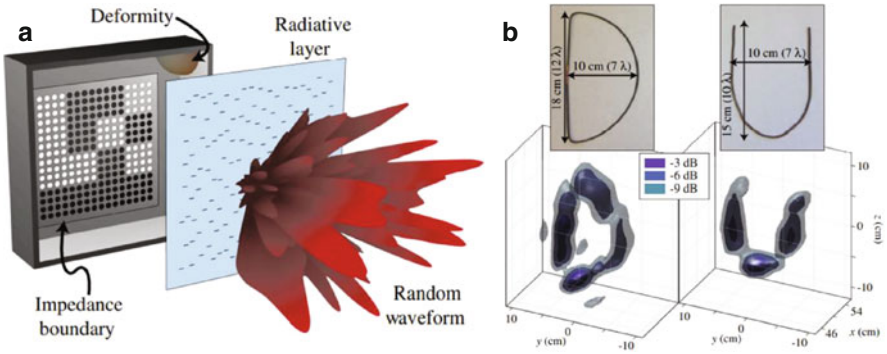

Fig. 13 (**a**) Cavity-backed dynamically tunable metasurface antenna; (**b**) reconstructed images of letters "D" and "U." (Adapted with permission from (Sleasman et al. 2016b) © American Physical Society)

the schematic diagram of a cavity-backed dynamically tunable metasurface antenna. It consists of a metallic hollow cavity with a spherical scatterer at one of its corners (Sleasman et al. 2016b). The scatterer perturbs the cavity modes and generates many diverse frequency-dependent cavity modes. A large number of slots are randomly distributed over the top plate of the cavity, and these elements radiate energy into free space. Pattern diversity is further improved by incorporating a tunable mushroom structures-based high-impedance surface on the face of the cavity (opposite to the face perforated with slots). The impedance of the surface is tuned by using varactor diodes which in turns generates abundant frequency-diverse cavity modes and ultimately many distinct radiation patterns (Sleasman et al. 2016b). The reconstructed images of the letters "D" and "U" using this proposed metasurface antenna are shown in Fig. 13b.

Figure 14a illustrates the schematic diagram of a dynamically reconfigurable metasurface antenna which can raster scan a given scene of interest electronically (Yurduseven et al. 2018). The device is composed of two printed circuit board (PCB) layers: bottom layer (PCB_1) and top layer (PCB_2) which are bonded together as shown in Fig. 14a to form the complete structure. The bottom-grounded PCB layer consists of a microstrip line loaded with complementary metamaterial elements and the PIN diodes connected across each metamaterial elements. The top layer carries the biasing circuitry associated with PIN diodes embedded on the bottom layer. This ensures a proper isolation of the dc-biasing circuitry from the RF circuit and thus stable operation of the antenna. A microwave signal is launched at one end of the microstrip line, and the other end is terminated with standard 50 Ω impedance. As a guided wave propagates through the microstrip line, energy gets coupled via metamaterial elements into free space. The states of different sets of metamaterial elements are controlled in such a way that it contributes to the formation of desired radiation pattern (Yurduseven et al. 2018). Figure 14b depicts three diverse radiation patterns corresponding to three different metamaterial element states.

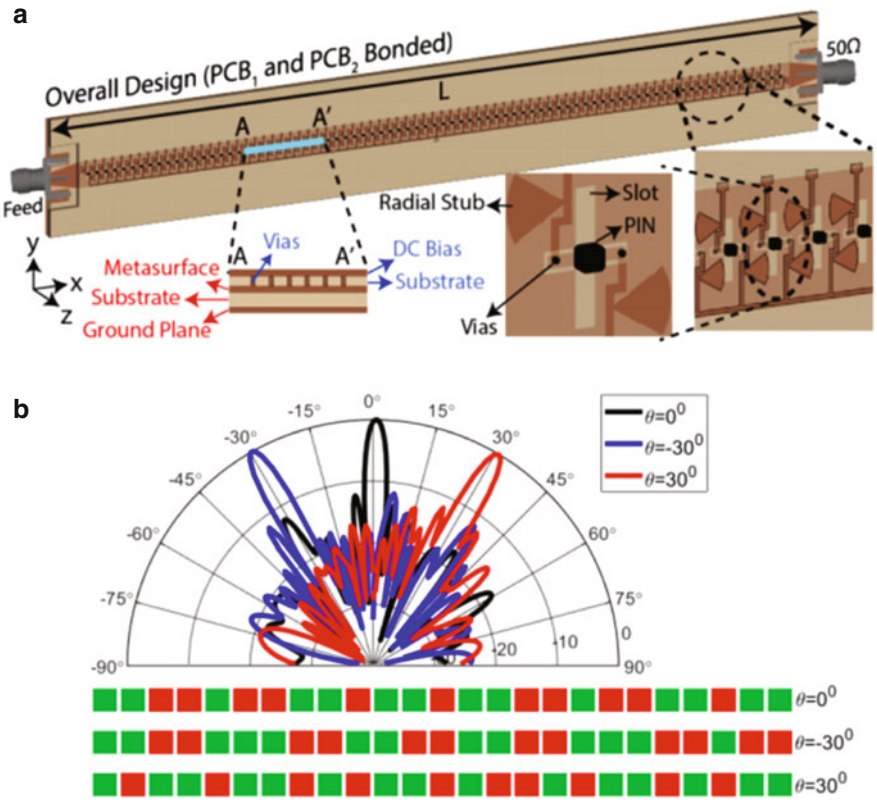

Fig. 14 (**a**) Schematic diagram of one-dimensional dynamic metasurface antenna; (**b**) radiation pattern corresponding different metamaterial element states (green: "ON" and red: "OFF"). (Adapted with permission from (Yurduseven et al. 2018)© The Optical Society)

Hybrid Metasurface Antennas

The research activities in microwave imaging are advancing at a higher pace. One of the recent developments includes the use of hybrid metasurface-imaging schemes to enhance the diversity of radiated fields from a metasurface antenna. In this approach, specific characteristics of different approaches are applied in combination. For example, the degree of freedom of a frequency-diverse metasurface antenna can be improved by rotation synthesis without increasing the complexity of an imaging system (Wu et al. 2016). In this method, a metamaterial antenna panel is rotated around the aperture axis which results in the generation of more pseudorandom radiation patterns or measurement modes in addition to the diverse patterns accomplished using simple frequency sweeping. Metasurface with frequency and polarization diversity has also been reported in literature for imaging applications (Zhu et al. 2017). The proposed metasurface contains metamaterial elements whose resonant frequencies are randomly distributed within the operational bandwidth.

The polarization attribute is introduced by rotating the metamaterial elements about its axis which makes the meta-aperture sensitive to incident wave polarization. This method significantly improves the number of uncorrelated measurement modes. Polarimetric imaging allows the collection more information about the scene of interest using the polarization aspect or vector nature of radiated fields. In conventional microwave imaging, scalar fields radiated from the transmitting and receiving aperture are used in the reconstruction of a given scene. However, with the advent of modern computers, it is possible to analyze the polarization characteristics of radiated fields. For example, in (Peng et al. 2020) a cavity-backed frequency-diverse metasurface, an antenna is investigated for polarimetric imaging. The antenna consists of a hollow metallic rectangular waveguide cavity with randomized holes on the top plate of the cavity. The cavity supports a number of modes depending on its physical dimensions. The introduction of a spherical scatterer at one of the cavity's corners perturbs the rectangular eigen modes and generates abundant diverse modes. The cavity with multitude of diverse modes excites the randomized holes which couple energy to free space as the excitation frequency is swept over measurement bandwidth. Thus, the cavity-backed antenna produces spatially diverse radiation patterns with different polarization states and low correlation to sense the target and multiplex the image data into a complex measurement mode.

Digital or Programmable Metamaterial Antennas

Digital or programmable metamaterial antennas are gaining huge interest in real-time microwave-imaging techniques. In this method, the radiated fields of an antenna are modulated by means of a digitally controllable metasurface. A metasurface, loaded with switching elements for controlling the phase and amplitude of an incident wave, is the essential part of a digital or programmable metasurface antenna-based microwave-imaging system (Li et al. 2016). Different rows or columns of metamaterial elements are turned "ON" or "OFF" instead of turning each metamaterial element "ON" or "OFF" independent of other elements. This significantly reduces the overall system complexity and cost as compared to a dynamic metasurface antenna which is capable of addressing each metamaterial element independently. Figure 15a depicts a programmable metasurface-based single sensor microwave-imaging system. This approach utilizes a programmable metasurface controlled by random binary codes to generate many diverse transmission radiation patterns (Li et al. 2016). Each metamaterial unit cell has two layers which contains metallic patches and their feeding structures. Switching diodes are connected to the top metallic patches via horizontal feed lines while the bottom patches are connected via vertical feed lines. Thus, each unit cell can represent four phase levels 0, $\pi/2$, π, and $3\pi/2$ which are denoted by digital codes "00," "01," "10," and "11," respectively (Li et al. 2016). Random binary codes are delivered to each row or column of the metamaterial elements using a field-programmable gate array (FPGA). FPGA applies positive and zero voltages in rows or columns to turn "ON" and "OFF" metamaterial elements, respectively. In this metaimager, individual metamaterial

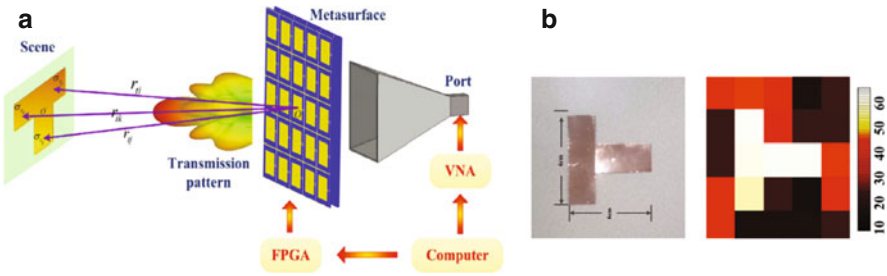

Fig. 15 (**a**) Schematic diagram of programmable metasurface-based single sensor microwave-imaging system; (**b**) reconstructed image of a metallic object using digital metaimager. (Adapted with permission (Li et al. 2016) © Springer Nature)

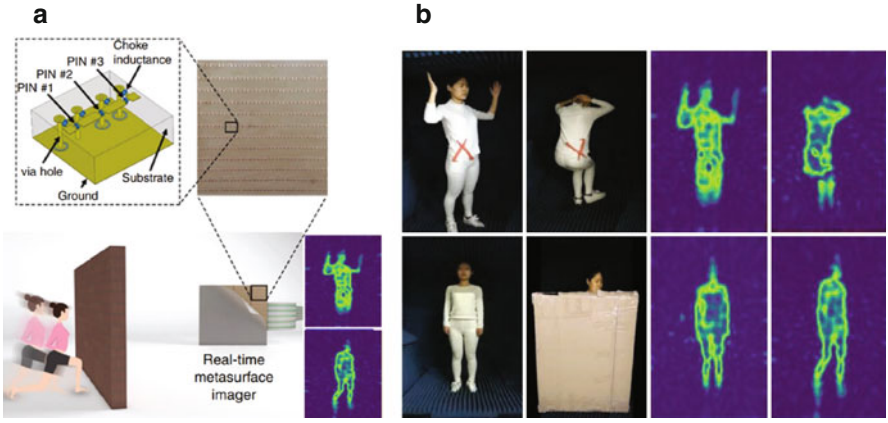

Fig. 16 (**a**) Machine learning assisted real-time microwave imaging of a moving target; (**b**) reconstructed images of a person. (Adapted with permission from (Li et al. 2019) © Springer Nature)

unit cells are not addressed by switching diodes, instead random 2 bit digital codes are applied to a row or column of metamaterial elements to spatially modulate the radiated fields (Li et al. 2016). Figure 15b shows a reconstructed image of a metallic object using the 2 bit-transmission-type-programmable metaimager. Digital metasurface imager in conjunction with machine learning can realize dynamic real-time imaging for surveillance applications (Li et al. 2019). A 2 bit-programmable metasurface is used to manipulate electromagnetic waves and generate arbitrary radiation patterns required by machine learning (Li et al. 2019). The programmable metaimager is trained with a large number of training samples in order to realize real-time imaging of moving targets. The schematic diagram of a metamaterial unit cell used in this programmable metasurface is shown in Fig. 16a. Each unit cell contains three metallic patches, and the adjacent metallic patches are connected with PIN diodes. Choke inductors are used to isolate dc circuitry from RF hardware (Li et al. 2019). Since each PIN diode can be turned "ON" or "OFF" by applying positive or zero

voltage, a metamaterial element can represent four binary states ("00," "01," "10," and "11"). FPGA is used to deliver digital codes to metamaterials to obtain desired radiation patterns. First, the reprogrammable metasurface imager is trained by desired radiation patterns using machine learning technique. Then, the digital codes required to synthesize the desired patterns are optimized. Finally, a target is illuminated with spatially diverse radiation patterns generated by the reprogrammable metasurface based on the machine learning-optimized digital codes or measurement modes (Li et al. 2019). The back scattered fields are processed to reconstruct images of a moving target. Figure 16b shows the reconstructed images of a person in different body positions and in a situation where the person is inside a cardboard box. The proposed metaimager can perform real-time microwave imaging of moving targets with high-spatial resolution. In addition, this approach can reduce image acquisition time and post image data-processing time drastically with the help of reprogrammable metasurface in combination with machine learning techniques (Li et al. 2019). Moreover, it can handle enormous amount of raw image data.

Metamaterial or FSS-Loaded Image-Sensing Antennas

Microwave imaging has a wide variety of applications. Some applications require only the shape or position information of a target. For example, detection of hidden crevasse in glacier terrain and nondestructive evaluation and testing such as detection of cracks or defects in the material under test. On the other hand, medical screening requires complete information of a given scene or a target under investigation which is related to the spatial distribution of dielectric properties of the target. Microwave tomography is generally used as a diagnosis method to screen cancerous growth. The target is sequentially illuminated by an antenna moving around the target or by multi-illumination method with several transmitting antennas (Pastorino 2010). The back scattered signals collected by receiving antennas are processed to reconstruct images. The generated image maps a reflectivity profile of a target under test based on difference in dielectric properties of healthy living tissues and malignant tissues. Depending on the application and the geometry of the target, arrangement of antenna elements varies. For example, a multistatic circular arrangement of antennas (Islam et al. 2019) is used to diagnose breast cancer as shown in Fig. 17. Antennas used for microwave tomography are required to have highly directive radiation beam in the entire frequency band of operation in order to ensure penetration of electromagnetic waves inside the target. Horn antennas, Vivaldi antennas, antipodal Vivaldi antennas, printed wideband directive monopole antennas, etc. are the commonly used high-directional antennas in imaging systems.

Artificially engineered materials such as frequency selective surfaces and metamaterials exhibit exceptional electromagnetic properties capable of manipulating the propagation of electromagnetic waves by bending, absorption, reflection, or guiding. A rich body of literature has investigated the use of frequency selective surfaces or metamaterial elements in improving the directivity, gain, bandwidth, and front to back ratio of antennas (Johari et al. 2017; Imran and Elwi 2017). Vivaldi or antipodal

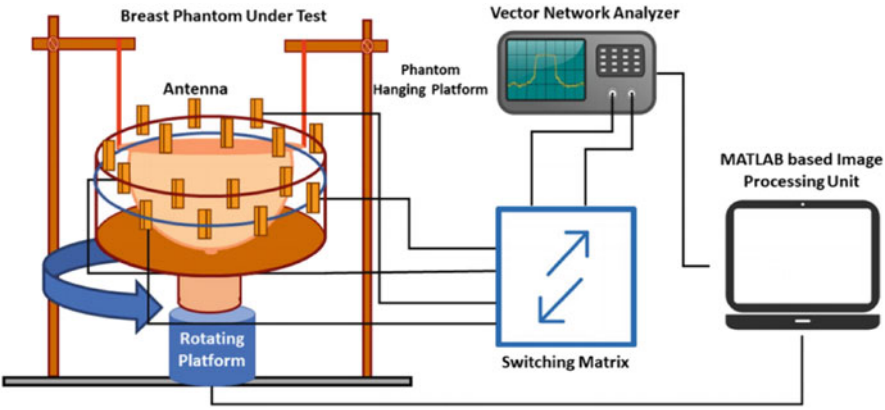

Fig. 17 Experimental setup for microwave imaging using metamaterial-loaded antennas. (Adapted with permission from (Islam et al. 2019) © Springer Nature)

Vivaldi antennas are preferred in microwave tomography due to their high directivity, wider operating bandwidth, planar geometry, and compact size (Islam et al. 2019). Conventional ultrawideband antennas with directional radiation pattern suffer from low gain in the lower operating frequencies (Gopikrishnan et al. 2020). A promising solution is to load near-zero index metamaterial elements on the radiating aperture. Near-zero index metamaterials have near-zero permittivity or permeability which results in near-zero refractive index (Johari et al. 2017). They reduce overall reflection and induce the propagation of waves in a specific direction. The placement of metamaterial elements on the antenna aperture is optimized based on the antenna geometry, direction of main radiated beam, and the mode of excitation. In the Vivaldi antenna proposed in Islam et al. (2019), the metamaterial elements are loaded in a region of maximum electric field to enhance the interaction of metamaterials with the radiated field as shown in Fig. 18. The loading of anisotropic near-zero index metamaterial unit cells increases the electrical length of the antenna which improves the antenna gain without increasing the overall dimension of the antenna (Islam et al. 2019). It also helps in focusing the radiation in the main lobe direction and reduces back scattering of waves. Thus, the use of metamaterials with near-zero refractive index increases the directivity and gain of the antenna. Another approach is to use frequency-selective surfaces to control and manipulate the propagation of electromagnetic waves. They are employed in antenna design to improve its overall performance including directivity, gain, bandwidth, and front to back ratio. A frequency selective surface is used as reflective layer to improve front to back ratio of a UWB antenna reported in Imran and Elwi (2017). A UWB antenna conformed over a cylindrical surface has been proposed for microwave radiology imaging. An FSS layer is introduced behind the antenna, and it acts as a reflective layer for the back scattered fields. The FSS layer reflects the back scattered waves in phase which adds constructively to contribute more radiation in broad side direction (Imran and Elwi 2017). Thus, the gain and front to back ratio of the FSS-based

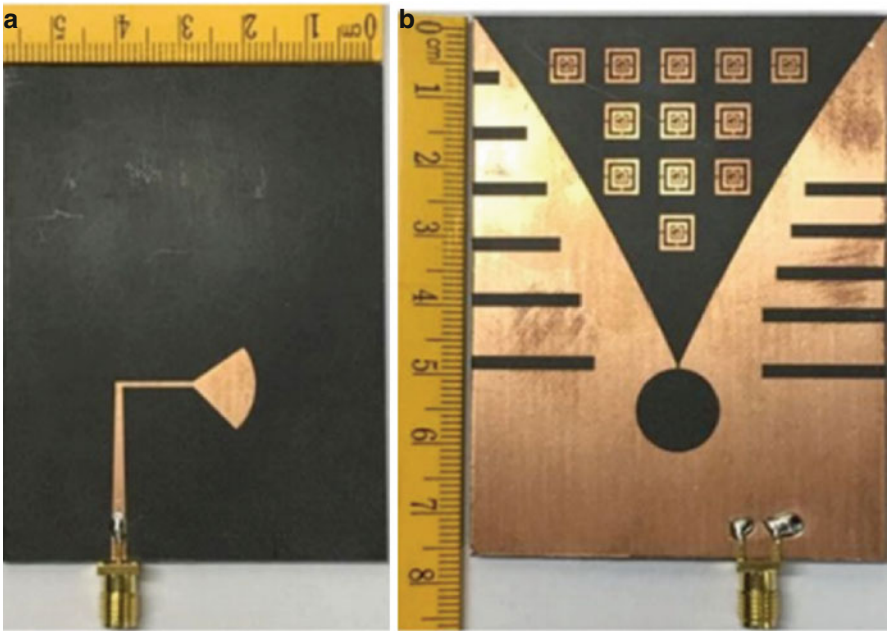

Fig. 18 Metamaterials-loaded Vivaldi antenna: (**a**) bottom view; (**b**) top view. (Adapted with permission from (Islam et al. 2019) © Springer Nature)

antenna is enhanced. Higher the directivity and gain of the sensing antenna is, better the accuracy of scanning which in turn results in the reconstruction of high-quality medical images.

Subdiffraction Imaging Using Metamaterials

Subdiffraction imaging allows the reconstruction of images with finest spatial details of a given source. The spatial information of a source is indexed in the spectral components (wavenumbers or spatial frequencies) associated with the imaged source (Iyer and Eleftheriades 2009). Therefore, the resolution of a reconstructed image ultimately relies on the precise reconstruction of spatial frequencies of a source at the image plane. The spatial resolution of conventional imaging systems is limited by classical diffraction limit. This arises due to the inevitable losses associated with the imaging systems and the maximum achievable resolution limits to an order of half the wavelength ($\lambda_o/2$), where λo represents the wavelength corresponding to the frequency of illuminating signal (Iyer and Eleftheriades 2009). In conventional lens-based imaging, only the interference of propagating spectral components is considered in the image reconstruction and does not account for evanescent waves. The evanescent waves decay exponentially as they travel away from the source. The elimination of evanescent wave components, which carry

the information of finest spatial details of a source, results in reconstructed image degradation. On the contrary, in subdiffraction imaging, metamaterial-based superlenses are employed due to their capability in restoring the evanescent and propagating spectral components of a source in the image plane (Iyer and Eleftheriades 2009). This allows microwave imaging beyond classical diffraction limit with a spatial resolution of several orders of magnitude higher than conventional diffraction-limited resolution. It finds applications in the imaging of closely spaced objects or targets such as microwave radio imaging of tumors, detection of land mines, and precise localized heating of cancerous tissues located in electrically small regions in microwave hyperthermia applications (Iyer and Eleftheriades 2009).

A negative refractive index metamaterial lens for free space imaging beyond classical diffraction limit has been reported in Iyer and Eleftheriades (2008). The proposed metamaterial superlens is composed of 43 metamaterial layers, and each layer consists of negative refractive index transmission line metamaterial unit cells as shown in Fig. 19. The properties of the metamaterials used in the design are strictly optimized in such a way that it exhibits low losses and simultaneous negative permittivity and permeability. The magnitude of these parameters matches with free space permittivity (ε_0) and permeability (μ_0), respectively. In other words, the metamaterials exhibit a negative refractive index of -1, permittivity $\varepsilon = -\varepsilon_0$ and permeability $\mu = -\mu_0$ (Iyer and Eleftheriades 2008). Thus, the negative refractive index metamaterial transmission line operates as a Veselago-Pendry superlens which is capable for restoring the amplitude of evanescent waves and phase of propagating waves at the image plane. The waves crossing the interface between free space double negative medium experience, an opposite sign of refractive index which changes the slopes of tangential magnetic and electric fields at the boundary (Bilotti and Sevgi 2012). It is possible to generate resonant modes at the boundary by adjusting the electrical and magnetic properties of metamaterial layer and the spacing between different metamaterial layers. These resonant modes assist in the enhancement of evanescent spectral components and hence resolution beyond

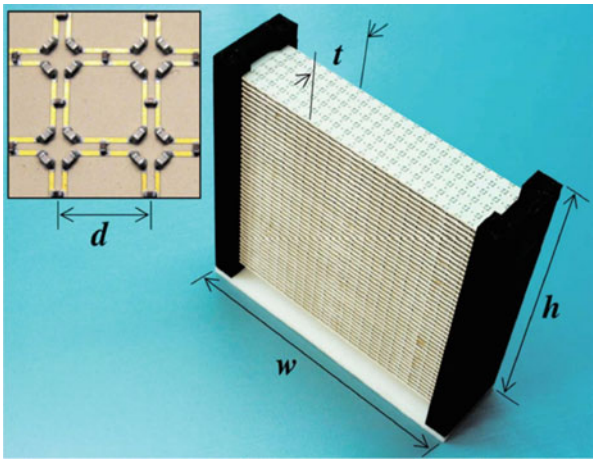

Fig. 19 Veselago-Pendry negative refractive index transmission line superlens for subdiffraction imaging. (Adapted with permission from (Iyer and Eleftheriades 2008) © The AIP Publishing)

diffraction limit. The experimental validation is provided by the imaging of two coherent loop antennas located at distance of t/2 from the front face of the superlens (Iyer and Eleftheriades 2008). The source antennas are separated by a transverse distance of $\lambda_0/3$. The detector antenna located behind the lens, which connected to a network analyzer, measures the field components. The measured transverse field profile showed a clear distinction of imaged sources with super resolution at 2.08 GHz and 2.40 GHz (Iyer and Eleftheriades 2008).

Summary

This chapter briefly discussed various metamaterial-FSS-based antennas for microwave image-sensing applications. An overview of microwave-imaging technique and the role of antennas in image sensing are presented in the beginning of the chapter. The ability of metamaterial or FSS in controlling the amplitude and phase of electromagnetic waves is the principle idea behind using these artificial structures in the antenna design for improved radiation characteristics. The metamaterial-FSS image-sensing antennas are broadly classified into three categories: image-sensing metasurface antennas, digital or programmable metasurface antennas, and frequency-selective surfaces-/metamaterials-loaded image-sensing antennas. The metasurface antennas capable of generating spatially diverse complex radiation patterns enable the multiplexing of image data into a few measurement modes. This allows image compression at the physical layer of the imaging system itself. Thus, the complexity of image reconstruction process is shifted from hardware to software level. The advanced signal-processing capabilities of modern computers aid in fast image reconstruction. In addition, these metasurface antennas eliminate the need of electronic or mechanical scanning of the aperture which in turn reduces the hardware complexity, cost, and image data acquisition time. The digital or programmable metasurface-based image-sensing antennas utilize a programmable metasurface to spatially modulate the fields illuminating the target. Semiconductor-switching elements are incorporated in the metasurface to control the phase and amplitude of the incident waves. The third category of image-sensing antennas utilizes frequency selective surfaces/metamaterial elements to improve their directivity, gain, and front to back ratio. The loading of these artificial structural units which have anomalous electromagnetic properties enhances the radiation characteristics of conventional directive antennas by focusing or guiding radiating fields in a specific direction. The operating principle, design considerations, and advantages of these three categories of image-sensing antennas are discussed in detail. Finally, a brief overview of metamaterial-based sub-diffraction imaging is given at the end of the chapter.

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A High-Isolated CPW-Fed Multi-Band Metamaterial Inspired MIMO Antenna for Wireless Applications

18

Harleen Kaur and Hari Shankar Singh

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Abstract

In this chapter, a CPW-fed metamaterial-inspired antenna with asymmetrical bell-shaped ground is designed and investigated for wireless applications. The overall size of single-port antenna is $50 \times 36 \times 0.3$ mm^3 ($0.32\ \lambda_l \times 0.2\lambda_l \times 0.002\lambda_l$, where, λ_l is at lowest cut-off frequency). The top layer of flexible FR-4 dielectric substrate contains an array of 3×4 Hilbert-shaped unit cells and asymmetrical bell-shaped partial ground. The proposed antenna offers impedance bandwidth of 1.92 GHz to 2.83 GHz, 3.22 GHz to 3.87 GHz and 5.4 GHz to 5.95 GHz (covering Wi-Fi/Wi-MAX/WLAN standards). Further, to combat the multipath fading or enhance the channel capacity, multiple-input multiple-output (MIMO) antenna with two elements is analyzed and tested. However, maintenance of isolation between antenna elements is another matter of concern. In view to this, a reverse T-shaped rectangular strip is placed exactly at the middle part of the substrate that inter-connects both bell-shaped CPW grounds. This greatly enhances the inter-port

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isolation between elements i.e. $S_{21} \leq -17.5$ dB over operating frequencies 1.92 GHz to 3.3 GHz and 4.6 GHz to 6 GHz (covering Wi-Fi/WLAN standards). Moreover, the measured and simulated results of the proposed 2-element MIMO radiator show good agreement with each other. Furthermore, simulated study of 4-element MIMO antenna with or without actual communication environment has also been analyzed and compared which inevitably makes the proposed antenna a suitable candidate for modern wireless applications.

Introduction

The advancement in wireless communication has been rapidly increasing with the growing needs of the users. The demand for high data rates and less fading is still a major concern which is why nowadays industrialists and researchers have been exploring multiband antennas. It has gained a lot of attention due to its rapid demand in upcoming wireless devices (Lee and Seo 2009; Pei et al. 2011; Sun et al. 2012). Over the past two decades, microstrip antenna has been preferred as it has many advantages such as low profile, light weight and ease of integration (Volakis et al. 2007). The major disadvantage of this method is narrow impedance bandwidth (IBW), low radiation efficiency, high cross-polar radiation, and less gain which makes it unfit for modern wireless devices. To enhance the antenna bandwidth, several schemes have been reported in the past few decades like varying the height of the substrate (Rayno and Sharma 2012), modifying the shape of radiator and ground via etching slots or different geometries (Liu et al. 1997; Huang and Yu 2011; Guha et al. 2011), and different feeding techniques (Chung et al. 2005; Qu et al. 2006). Recently published manuscripts show that the incorporation of meta-material (MTMs) unit cells in the printed circuit antennas leads to unprecedented advancement in terms of antenna's bandwidth performance, miniaturization and radiation pattern (Pandey et al. 2014; da Silva Paiva et al. 2019). MTMs are the electromagnetic (EM) periodic structures which do not exist in nature in their own. The most important characteristic of an MTM is that both of its permeability ($\mu < 0$) and permittivity ($\varepsilon < 0$) values should be negative concurrently. This unusual behavior such as left-handed (LH) EM wave propagation, negative refractive index, antiparallel phase and group velocity, zero propagation constant at ZOR is not possible in case of naturally occurring right-handed (RH) materials. MTM inspired antennas can be designed by utilizing the transmission lines like composite right/left-handed (CRLH), epsilon negative and mu-negative. Another significant approach is by loading the resonators such as split-ring (SRR) and complementary split-ring (CSRR). But the surrounding environment and conducting ground plane may significantly affect the antenna performance (Sharawi 2017). To date, numerous components based on metamaterials (MM) have been extensively deployed due to their remarkable characteristics when interacting with EM waves. Some of its applications include near-zero index (Liberal and Engheta 2017), polarization conversion (Hao et al. 2010), cloaking (Alitalo and Tretyakov 2009), and perfect absorption performance (Li et al. 2010). In line of this, researchers have been working on enhancing the system's efficiency by utilizing MIMO

technology but maintenance of inter-element isolation is highly challenging due to rising demands of small sized devices. For the past few years, a vast amount of research on improving the isolation between antenna elements has been reported by several academician and industrialists (See et al. 2012; Anitha et al. 2016; Malviya et al. 2017; Kaur et al. 2018). Different isolation enhancement techniques include decoupling networks (Kaur et al. 2019), defected ground structures (DGS) (Luo et al. 2015), neutralization lines (Zhang and Pedersen 2015), parasitic elements (Payandehjoo and Abhari 2014) and meta-materials (MTMs) (Payandehjoo and Abhari 2015; Liu et al. 2019). Generally, the orthogonal and adjacent arrangements are preferred to increase the isolation between antenna elements (Sipal et al. 2018; Kaur et al. 2021a). In (Zhu et al. 2010), authors Zhu et al. present a compact CPW-fed triple-band monopole radiator that utilizes the concept of reactive loading with a defected ground-plane for Wi-Fi and Wi-MAX applications. Likewise in (Huang et al. 2015), authors Huang et al. designed a compact single-element multi-band metamaterial-loaded antenna for WLAN and Wi-MAX applications. The top of the FR-4 substrate has a microstrip-fed monopole with an inverted L-shaped slot and rectangular patch whereas the bottom side of the substrate contains another rectangular patch and a thin short circuited strip. The surface current distribution at WLAN band shows that the waves appear at single cell MTM loading creates an energy loop which consequently leads to the effective radiation. Meanwhile MIMO antennas based on MTM concept were also rapidly gaining attention as authors in (Panda et al. 2017) presented a compact high-isolated aperture-coupled dual-band MTM based MIMO antenna for Long term Evolution (LTE) and WiMAX applications. Each element is composed of complementary split ring resonator (CSRR) and square-ring slot radiator (SRSR) to obtain two different resonant frequencies at 2.4 and 3.5 GHz. Further, an MTM decoupling element comprising of complementary two-turn spiral resonator (CSR2) unit cells are integrated between the radiating elements to reduce the mutual coupling. Further in (Wang et al. 2018), miniaturized CPW-fed dual band-notched ultra-wide band (UWB) MIMO antenna with MTM inspired isolator for mobile device applications has been discussed. Here a novel single unit cell SRR has been utilized to achieve high inter-port isolation. Likewise authors in (Singh et al. 2019) designed a compact CPW-fed MTM inspired antenna for ultra-wideband applications. The top layer of substrate contains a radiating element that is made up of three split-ring resonators (SRR) which are positioned along x-axis. Each SRR consists of two oppositely situated concentric split rings, which is made of metal. The chosen SRR geometry leads to miniaturization in the design and offers stable radiation patterns. Also, the modified ground plane improves the impedance matching over operating frequencies. Recently in (Moussa et al. 2021), authors Moussa et al. designed the slotted meta-material decoupling slab (S-MTM-DS) with dual reflexes and slotted E-shapes connected with an inductive stub for wireless applications. Its main objective was the inter-element isolation enhancement, attainment of miniaturized design, improvement of fractional bandwidth and concealment of return losses.

In this work, initially a CPW-fed multiband MTMs-inspired antenna is proposed for wireless applications. The unit cells are loaded in the radiating patch and a bell shaped CPW partial ground is designed near the feed which improves the impedance matching.

Here, the CPW feeding ground plane not only lowers the radiation loss but also less sensitive to the thickness of substrate. Moreover, the conventional coaxial probe feeding method is more complex than coplanar waveguide (CPW) fed technique as it hampers the conformability in structures. The measured and simulated results show that the proposed radiator can operate over multiple wireless standards such as Wi-Fi, Wi-MAX and WLAN. Further, the performance of 2-element MIMO antenna is analyzed and tested. High isolation of more than −17.5 dB over operating band of frequencies is attained via reverse T-shaped strip attached to CPW grounds of each element. Measured and simulated S-parameters results are in good agreement with each other which makes it eligible candidate for practical Wi-Fi and WLAN applications. Likewise, the simulated study of 4-element MIMO antenna with free space and with actual communication environment is also illustrated. All the simulations have been performed using electromagnetic (EM) simulator named computer simulation technology microwave studio (CST MWS) (CST Microwave n.d.). The rest of the paper is divided into five sections. Section "Antenna Design and Methodology" discusses the methodology involved in designing the proposed antenna and their effect on S-parameters; it also includes some parametric variation study along with the effect of varying the orientation and position of proposed radiator, whereas section "Characterization of Proposed Single-Element Antenna" contains all the results and discussion of some important antenna parameters such as surface current distribution, radiation pattern, and diversity parameters. Further, in section "Analysis of Proposed Antenna with Different Configurations Design," the performance of MIMO antenna in actual communication environment is evaluated. Finally, the work has been concluded in the last section.

Proposed Antenna Design and Methodology

The layout of the proposed MTMs inspired antenna is shown in Fig. 1a whose size is 50×36 mm^2. The radiating patch is made up of copper material and loaded onto the top layer of FR-4 flexible substrate of thickness 0.3 mm (with relative permittivity of 4.3 and loss tangent of 0.017). It consists of 3×4 Hilbert-shaped unit cells. For proper impedance matching, the proposed antenna is excited via 50Ω CPW feed line with two asymmetric bell-shaped ground planes. Figure 1c shows the simulated reflection coefficients of proposed antenna whose values are less than −10 dB around frequencies from 1.92 GHz to 2.83 GHz, 3.22 GHz to 3.87 GHz and 5.4 GHz to 5.95 GHz (covering Wi-Fi, Wi-MAX and WLAN bands). Table 1 illustrates the dimensions of all parameters that were involved while designing the proposed antenna.

Characterization of Proposed Single-Element Antenna

Figure 2a illustrates the layout design of unit cell used in proposed antenna along with its dimensions. The design is formed by merging twelve square-shaped patches (each having different area) together and placed inside a square-shaped ring. Figure 2b depicts the simulation setup where it is terminated with perfectly-matched boundary

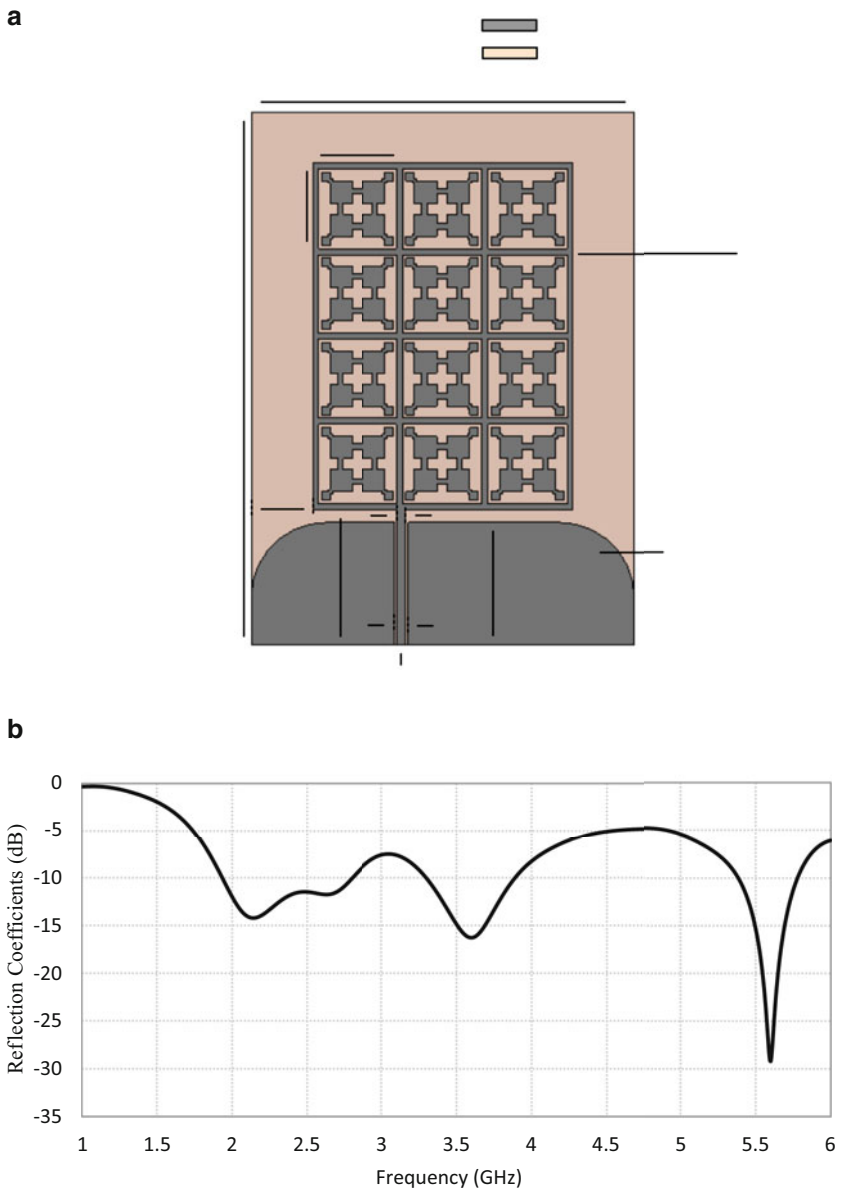


Fig. 1 (**a**) Layout of proposed CPW-fed MTMs inspired antenna; (**b**) Refection coefficients (in dB) of proposed antenna

conditions whereas Fig. 2c shows the simulated *S*-parameters of proposed unit cell. Here, the small patches (resonator) generate the inductive effect and the gaps created via ring form the capacitive effect. Due to their overall effect, the proposed unit cell

Table 1 List of parameters of proposed antenna design

Parameters	Dimensions (mm)	Parameters	Dimensions (mm)	Parameters	Dimensions (mm)
L_x	36	w	0.9	n_1	0.5
L_y	50	U_x	8.5	n_2	2.1
g_1	11.5	U_y	8.5	n_3	0.8
g_2	12.75	a_1	0.85	n_4	0.4
g_3	5.75	a_2	1.45	n_5	1
G	1.3	a_3	0.75		

resonates at frequency 8.7 GHz. Further, to study the electromagnetic characteristic of the unit cell, a method known as Nicolson–Ross–Weir has been used to compute its permittivity and permeability values. Figure 2d and e show the plot curves of effective permittivity (ε_{eff}) and effective permeability (μ_{eff}), respectively. The negative values of permittivity and permeability are noticed around 8.7 GHz which confirms the nature of metamaterial (double negative material). The negative value implies that the proposed metamaterial unit cell structure can be utilized to design a metamaterial inspired radiator so called metamaterial inspired antenna.

Analysis of Proposed Antenna with Different Configurations Design

In this section, design evolution process of proposed single-element MIMO antenna has been described and compared. Figure 2a shows different design steps used to evolve the proposed design. Its corresponding S-parameters are shown in Fig. 2b. The work is initiated by Design-1 stage in which 2 × 2 unit cells have been deployed as a radiating element. It is noticed that the Design-1 provides multi-resonance over the frequencies from 2.65–3.4 GHz and 4.4–5.5 GHz. Further to increase the electrical path length of radiator, Design-2 exploited 3 × 3 unit cells. With these extra unit cells in the radiating element, a new resonance has been created around 4 GHz. The return loss characteristics of Design-2 show that operating bandwidth lays within 3.4 to 4.3 GHz (Fig. 3).

Finally, in the proposed design, 3 × 4 unit cells have been used to achieve multi-band characteristics over wireless standards such as Wi-Fi, C, and WLAN band. This has been attained after performing several parametric variations on width of the feed and some key design parameters of unit cell. In addition, the size of CPW ground has also been tuned. Therefore, with proper optimization, the proposed single-element antenna can be able to cover frequencies from 1.9–2.8 GHz (Wi-Fi), 3.3–3.9 GHz (C-band), and 5.3 to 5.8 GHz (WLAN).

Design and Analysis of Proposed 2-Element MIMO Antenna

The design of proposed 2-element MIMO antenna is shown in Fig. 4a. The overall size is 50 × 82 × 0.3 mm^3. Initially, the radiating elements of the proposed antenna are positioned in such a way that the Element-1 is mirror image to the Element-2 and

the separation between antenna elements is kept at some distance '*s* '. Further, Fig. 4b shows its simulated *S*-parameters in which the reflection coefficients (S_{11}) has value less than −10 dB over frequencies 1.92 GHz to 2.83 GHz, 3.22 GHz to 3.87 GHz and 5.4 GHz to 5.95 GHz with poor isolation level (S_{12}) of less than −8.5 dB. Here, the grounds of each radiators are not connected which limits its usage in practical communication environment. Thus, the grounds of radiating elements have been joined together via a small strip as depicted in Fig. 5a. It can be noticed from Fig. 5b that simulated S_{21} parameter with the shared ground plane

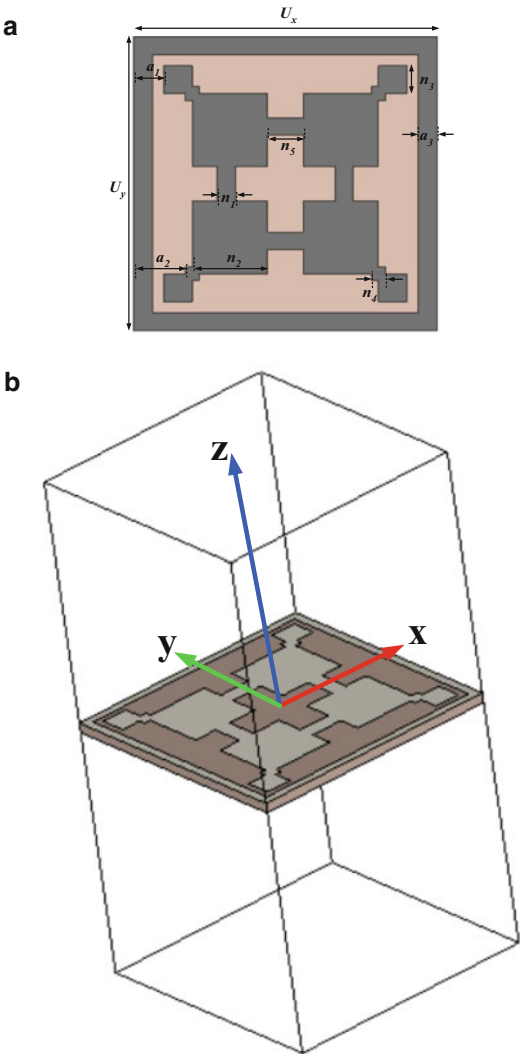


Fig. 2 (continued)

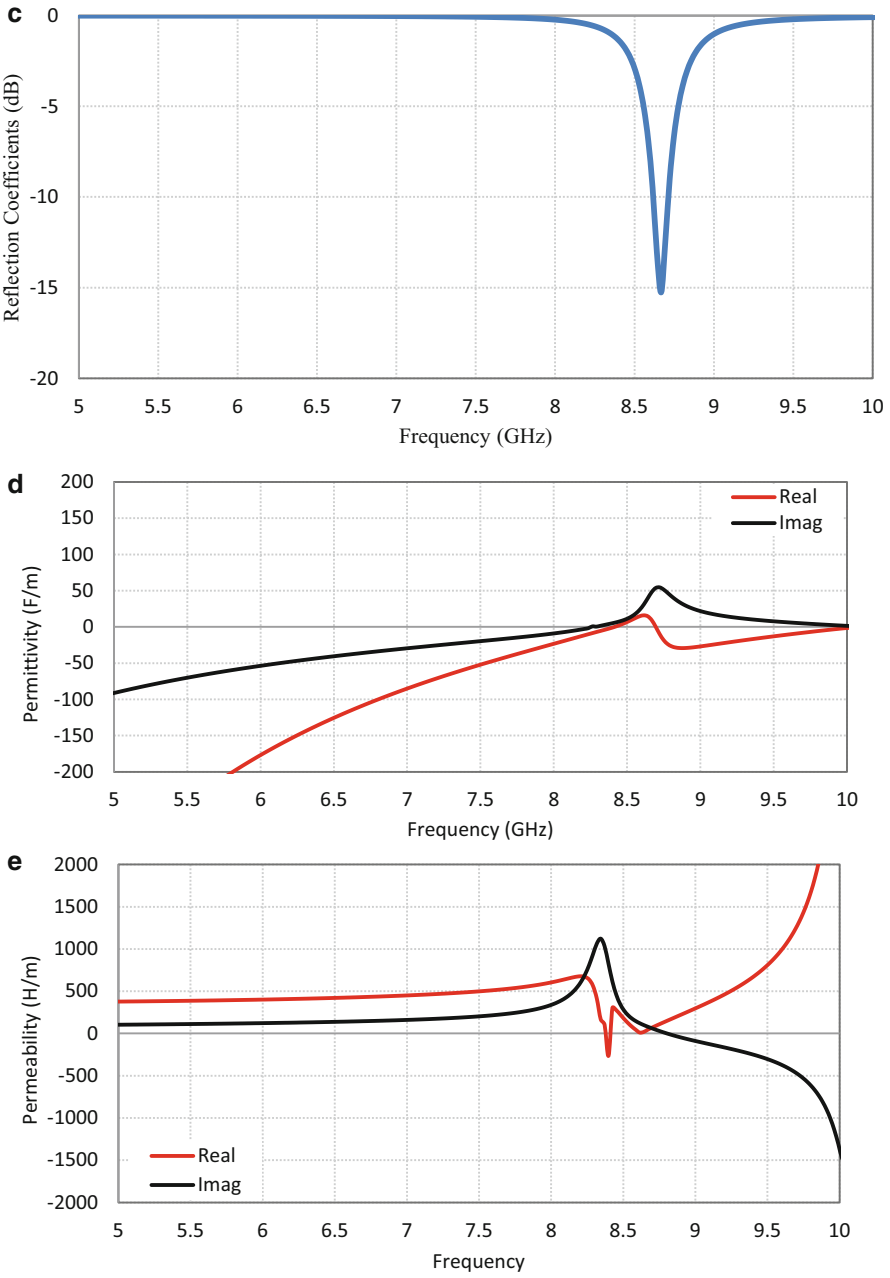


Fig. 2 (**a**) Single unit cell representation; (**b**) Refection coefficients (in dB) of proposed antenna; (**c**) Simulation setup of proposed unit cell placed within a waveguide; (**d**) Effective Permittivity of proposed unit cell; (**e**) Effective Permeability of proposed unit cell

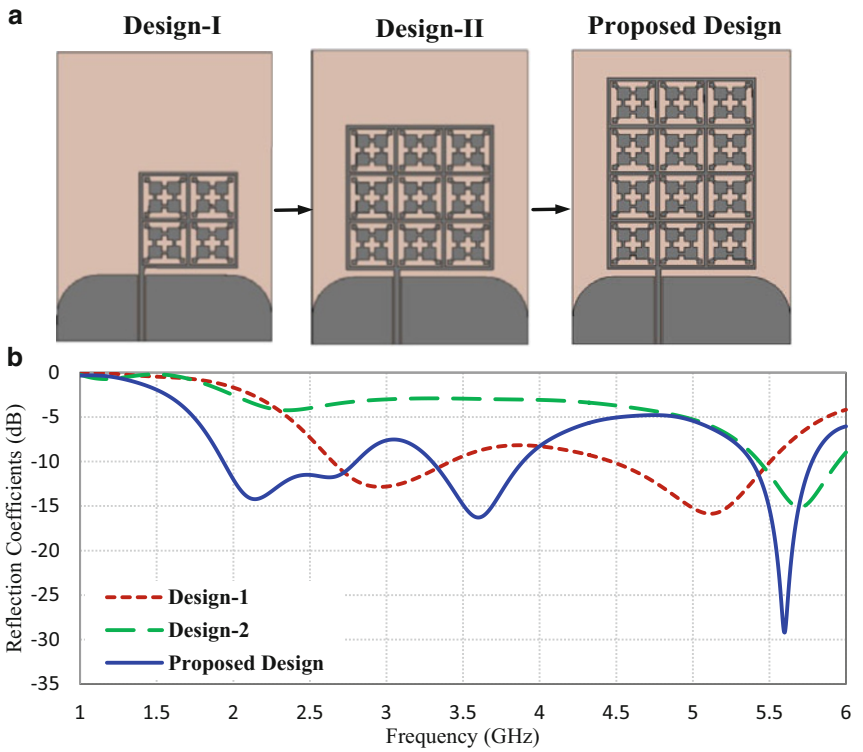


Fig. 3 (**a**) Design evolution process of proposed single-element CPW-fed MTMs inspired antenna; (**b**) Plot of simulated reflection coefficients (in dB) vs. frequency (in GHz)

has somewhat increased but still not offering desired level (≤ -15 dB) especially around the lower cutoff frequencies.

In view to this, a reverse T-shaped strip is introduced in Fig. 6a while maintaining the side-by-side arrangement between antenna elements. The strip connects the asymmetrical bell-shaped CPW grounds of each element. Further, the simulated S-parameters of the proposed antenna with or without strip are compared in Fig. 6b. By incorporating this strip rightly at the middle of these radiating elements, huge enhancement in inter-element isolation is achieved. However, antenna's operating frequency band now lies within 1.89 GHz to 3.2 GHz and 4.7 GHz to 6.2 GHz, covering only Wi-Fi and WLAN frequency bands.

For better understanding about enhanced isolation, Fig. 6c and d illustrate the surface current distribution of proposed 2-element MIMO antenna with and without reverse T-shaped strip (at 2.45 GHz) when port-1 and port-2 is excited, respectively. Only one port is excited at a time whereas the other port is matched terminated with 50 Ω loads. It is noticed that the current starts flowing from feeding point to the end of the radiating Element-1 which also confirms the resonance at 2.45 GHz. However, high coupling is observed between Element-1 and Element-2. Finally, to get the

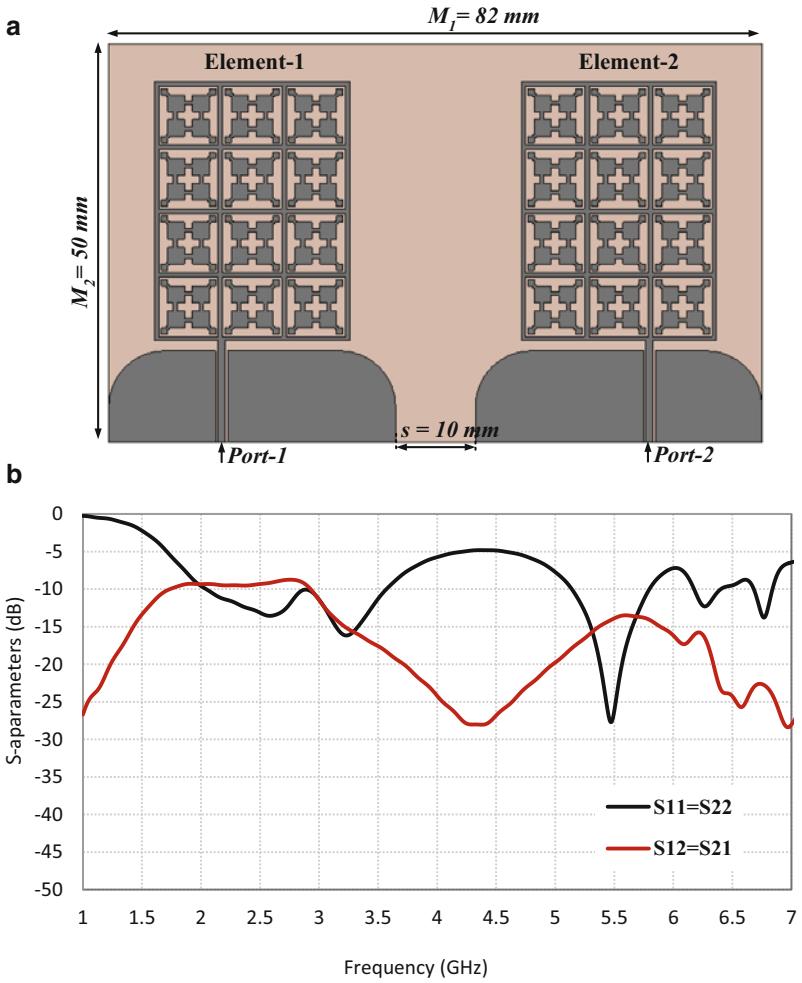


Fig. 4 (**a**) Design of proposed 2-element CPW-fed MTMs inspired MIMO antenna; (**b**) Plot of simulated S-parameters (in dB) vs. frequency (in GHz)

desired inter-element isolation, a combination of asymmetrical bell-shaped CPW ground and inverted T-shaped strip is attached. With this combination, overall electrical length of the antenna element is increased which not only shifts its resonant frequency corresponding to the $\lambda/4$ of the strip but also offers high isolation.

Experimental Validation

The planar and flexible dual-port proposed antenna's prototypes are fabricated as shown in Fig. 7a and b, respectively. The simulated results have been validated via vector network analyzer [Keysight E5063A (100 kHz-18GHz)] as depicted in

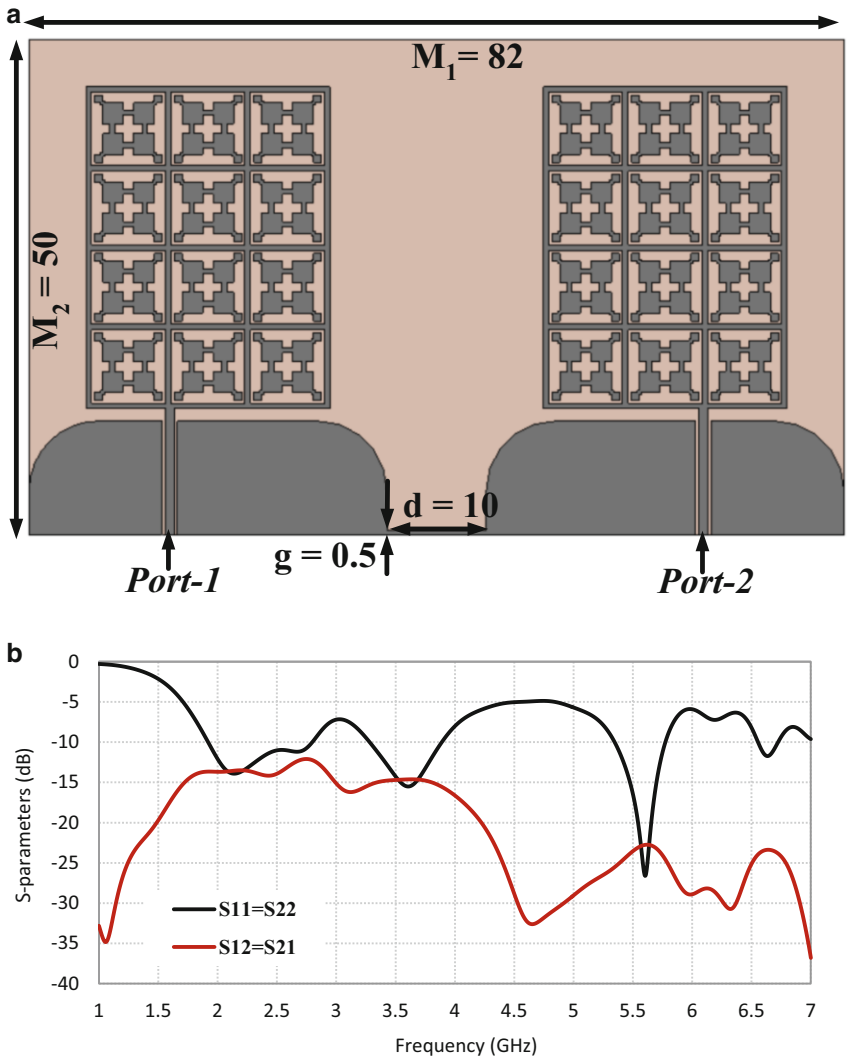


Fig. 5 (**a**) Design of proposed 2-element CPW-fed MTMs inspired MIMO antenna with orthogonal arrangement; (**b**) Plot of simulated S-parameters (in dB) vs. frequency (in GHz)

Fig. 7c. Further, Fig. 7(d) highlights the comparison of measured reflection coefficients of single-port antenna with its simulated ones. Furthermore, Fig. 7(e) illustrates the comparison between the simulated and measured Sparameters results of 2-element proposed MIMO antenna. Also, the measured reflection coefficients at each port, i.e. (S_{11}) and (S_{22}), do not offer same values even though both of the elements are designed in such a way that the Element-1 is mirror image to the Element-2. Because of the fabrication inconsistency, a slight variation is noticed between simulated and measured results. Likewise, measured transmission

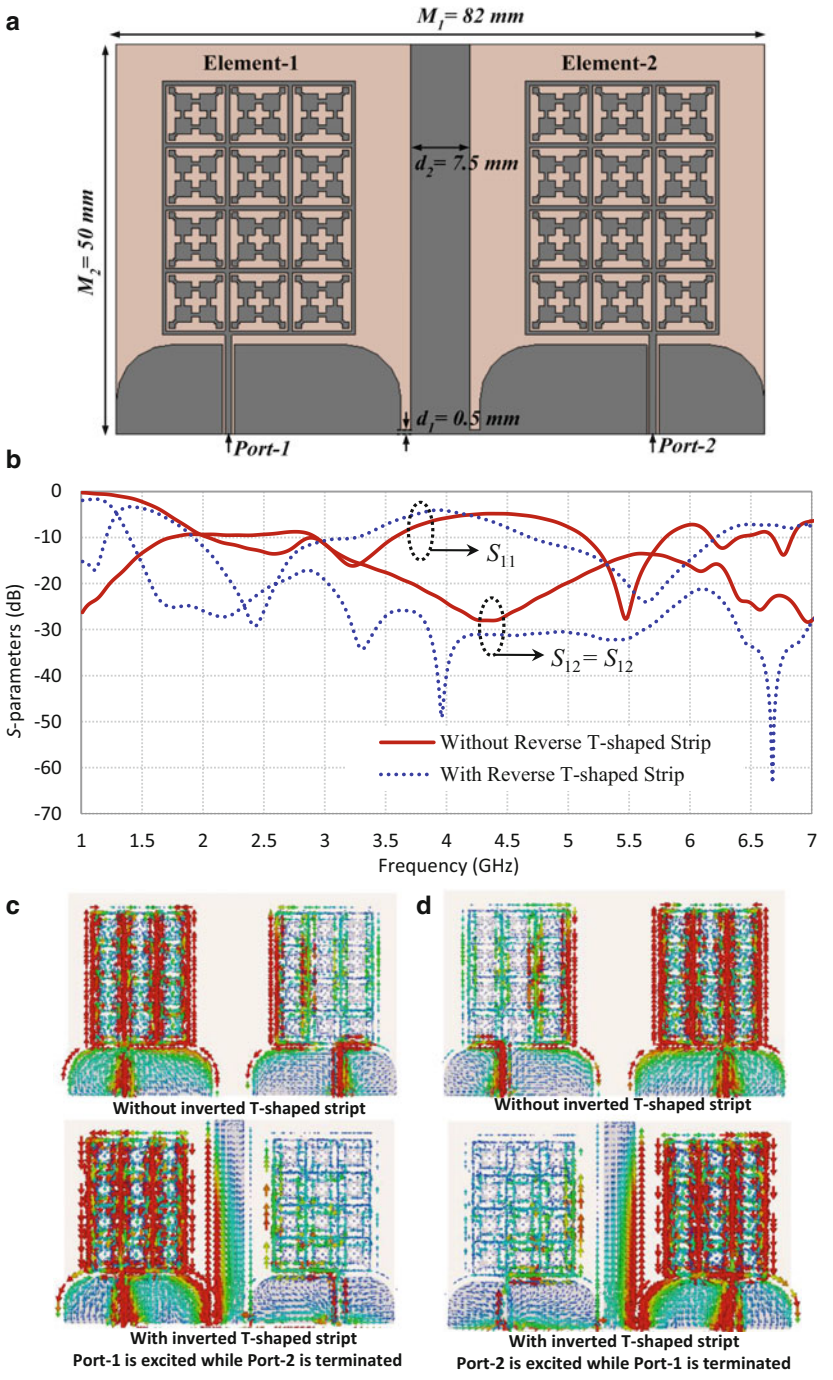


Fig. 6 (continued)

coefficients result at each port i.e. (S_{12}) and (S_{21}) closely matches with the simulated transmission coefficient.

Radiation Characteristics

In this section, the antenna's radiation characteristics along with its some key performance parameters have been discussed. Figure 8(a) shows the simulated 3D radiation pattern of proposed dual element antenna at two different frequencies (2.45 GHz and 5.5 GHz). Further, Fig. 8(b) highlights the simulated and measured 2D radiation patterns in horizontal-plane ($\theta = 90^{\circ}$) and vertical-plane ($\phi = 90^{\circ}$) at frequencies 2.45 GHz and 5.5 GHz, respectively. It is noticed that the radiation patterns of the Element-1 are complimentary to that of Element-2 which confirms pattern diversity, and consequently offers low correlation and high diversity performance in free space. Moreover, Fig. 9 depicts the simulated peak realized gain and total efficiency of the proposed antenna. All the simulations have been carried out in CST MWS electromagnetic software. The peak realized gain of the proposed antenna is around 5 dBi and total antenna efficiency varies from 80% to 93% over desired operating bands.

Diversity Performance

To evaluate the performance of proposed dual-element MIMO antenna, all the diversity parameters such as envelop correlation coefficients (ECC) (Zhao and Choi 2014), effective diversity gain (EDG) (Kaur et al. 2021b), mean effective gain (MEG) (Taga 1990), channel capacity loss (CCL) (Chae et al. 2017), and total active reflection coefficients (TARC) (Andrade et al. 2019) have been calculated. In regards to MIMO antennas, ECC is an essential parameter for the estimation of inter-dependence of channels. It quantifies the degree of correlation between antenna elements. Its value should be very low to ensure high capacity and diversity gain (DG). Generally, ECC in MIMO systems is computed using both far-field radiation patterns and S-parameters. The generalized formula for attaining ECC using S-parameters is described as the following equation:

Fig. 6 (**a**) Design of proposed 2-element CPW-fed MTMs inspired MIMO antenna with reverse T-shaped strip; (**b**) Plot of simulated S-parameters (in dB) vs. frequency (in GHz); (**c**) Surface current plots of proposed 2-element MTMs inspired MIMO antenna when only port-1 is excited; (**d**) Surface current plots of proposed 2-element MTMs inspired MIMO antenna when only port-2 is excited

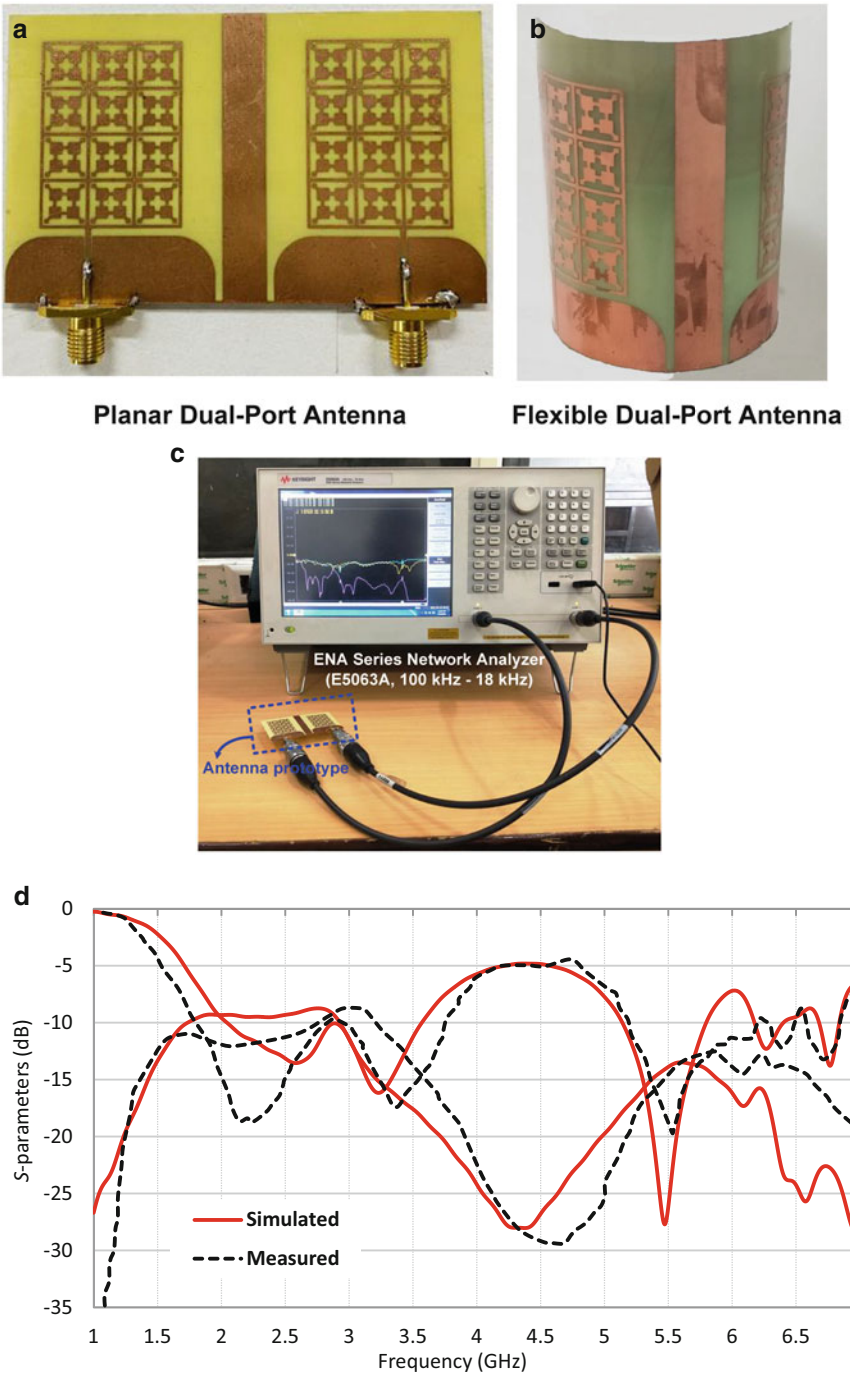


Fig. 7 (continued)

$$\mathrm{ECC} = \left| \frac{\sum_{n=1}^{N} s_{ni}^{*} S_{nj}}{\sqrt{\left(1 - \sum_{n=1}^{N} \left|S_{nj}\right|^2\right)\left(1 - \sum_{n=1}^{N} \left|S_{nj}\right|^2\right)}} \right|^2 \tag{1}$$

where S_{11} and S_{22} indicate the reflection coefficients (in dB) whereas S_{12} and S_{21} denotes the mutual coupling (in dB) of 2-element antenna array. The results using formula given in Eq. (1) is quiet simple to obtain but this method leads to degradation in radiation efficiency (Srinivas et al. 2014). The Sparameter method works accurately only when the total radiation efficiency of ports is 100%. Further, another version of this method has been taken into account in which non-ideal radiation efficiency of antenna elements is considered. The formula for which is describes as follows:

$$\mathrm{ECC} = \left| \frac{\sum_{n=1}^{N} s_{ni}^{*} S_{nj}}{\sqrt{\left(1 - \sum_{n=1}^{N} \left|S_{nj}\right|^2\right)\left(1 - \sum_{n=1}^{N} \left|S_{nj}\right|^2\right)\eta_{\mathrm{rad},i}\eta_{\mathrm{rad},j}}} + \sqrt{\left(\frac{1}{\eta_{\mathrm{rad},i}} - 1\right)\left(\frac{1}{\eta_{\mathrm{rad},j}} - 1\right)} \right|^2 \tag{2}$$

This method certainly improves the computation accuracy of ECC but only valid for efficiencies more than 60%. Moreover, both of the S-parameters methods neglect the directive angular power spectrum (APS) distributions by assuming only uniform APS distribution (Elshirkasi et al. 2019). Hence, another method is utilized for calculating ECC with the help of far-field radiation pattern, whose mathematical representation is defined as follows:

$$\rho_e(i,j) = \frac{\int_0^{2\pi}\int_0^{\pi} A_{i,j}(\theta,\phi)\sin\theta\, d\theta d\phi}{\sqrt{\int_0^{2\pi}\int_0^{\pi} A_{i,i}(\theta,\phi)\sin\theta\, d\theta d\phi} + \sqrt{\int_0^{2\pi}\int_0^{\pi} A_{\,j,j}(\theta,\phi)\sin\theta\, d\theta d\phi}} \tag{3}$$

$$A_{i,j}(\theta,\phi) = \mathrm{EPR}.E_{\theta i}(\theta,\phi)E^{*}{}_{\theta j}(\theta,\phi)P_{\theta}(\theta,\phi) + E_{\phi i}(\theta,\phi)E^{*}{}_{\phi j}(\theta,\phi)P_{\phi}(\theta,\phi) \tag{4}$$

where XPR denotes the cross-polarization ratio between vertical and horizontal power components, $E_{\theta i}$, $E_{\theta j}$, $E_{\theta i}$ and $E_{\theta j}$ represents the electric fields for the i^{th} and j^{th} antenna that varies with θ and ϕ directions, respectively. P_{θ} and P_{ϕ} are the angular power spectrum distributions in the direction of θ and ϕ, and they must satisfy the below given conditions:

Fig. 7 (**a**) Fabricated prototype of proposed planar dual-port CPW-fed MTMs inspired antenna; (**b**) Fabricated prototype of flexible dual-element CPW-fed MTMs inspired MIMO antenna with reverse T-shaped strip; (**c**) Measurement Setup used while testing; (**d**) Comparison of simulated and measured S-parameters of proposed dual-port antenna

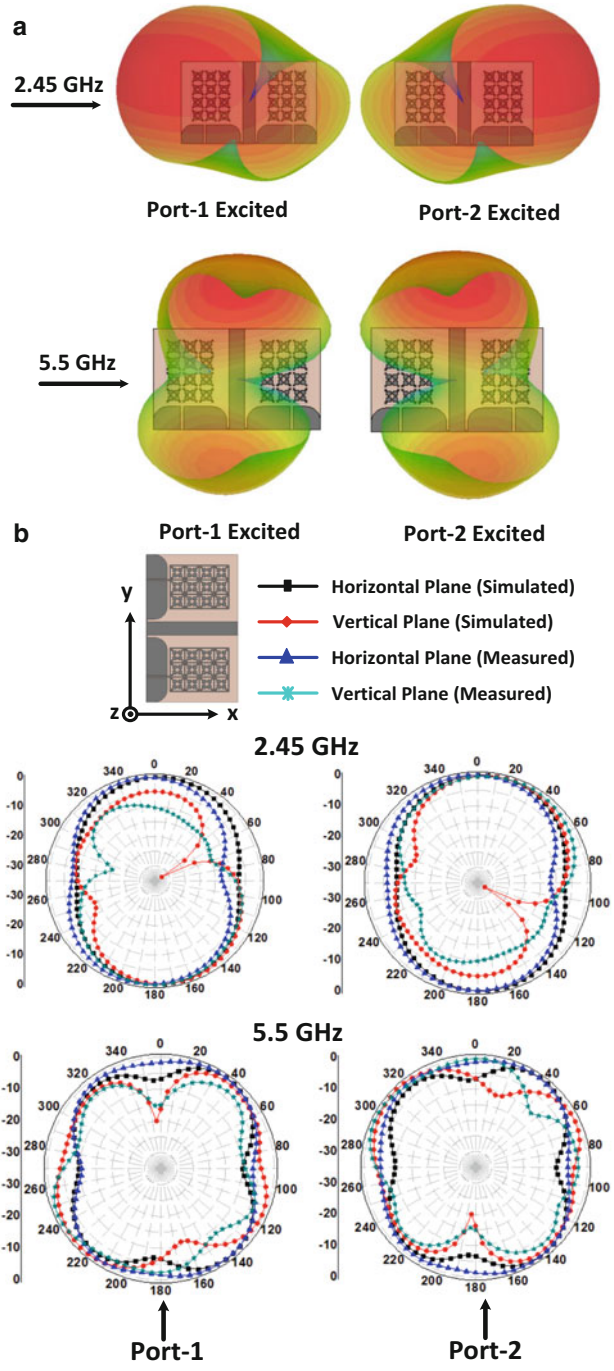


Fig. 8 (continued)

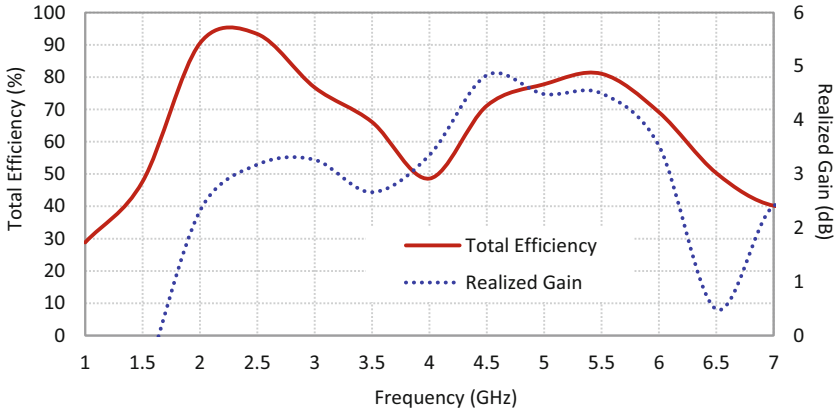


Fig. 9 Simulated peak realized gain and total efficiency of the proposed antenna

$$\int_0^{2\pi}\int_0^{\pi} P_\theta(\theta,\phi)\sin\theta d\theta d\phi = \int_0^{2\pi}\int_0^{\pi} P_\phi(\theta,\phi)\sin\theta d\theta d\phi = 1 \tag{5}$$

ECC using far-field radiation pattern approach provides the most accurate results but a lot more time consuming as compared to S-parameter approach. Figure 10a shows the simulated ECC of the designed antenna using both far-field and S-parameter method. The values of ECC should lie below 0.5 over the operating band of frequencies to ensure low correlation among radiating elements. Further, the strength of any diversity system is analyzed by the diversity gain parameter. The method for obtaining DG is defined as follows (Singh et al. 2014):

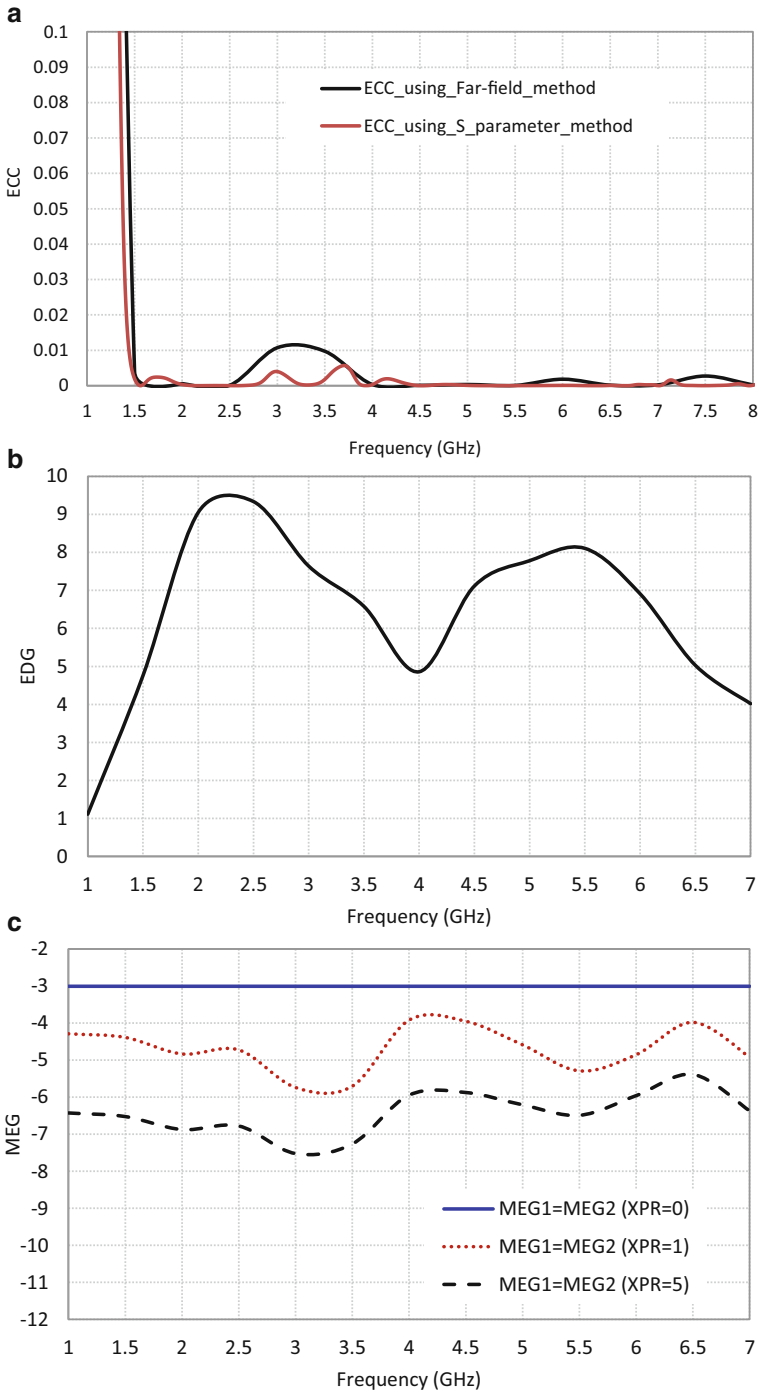
$$DG = 10\sqrt{1 - |\,\mathrm{ECC}\,|} \tag{6}$$

The above formula of DG does not consider the radiation efficiency which is why effective diversity gain (EDG) is observed whose formula is given as:

$$\mathrm{EDG} = DG^{*}\mathrm{Total\ \ Efficiency} \tag{7}$$

Figure 10b depicts EDG values which are computed with the product of diversity gain and total efficiency of antenna. The computed values offer reasonably good EDG at operating region of frequencies.

Fig. 8 (**a**) Simulated 3D radiation pattern of proposed dual element antenna at frequencies 2.45 GHz and 5.5 GHz, respectively; (**b**) Comparison of simulated and measured 2D radiation patterns in horizontal-plane ($\theta = 90°$) and vertical-plane ($\phi = 90°$) at frequencies 2.45 GHz and 5.5 GHz, respectively



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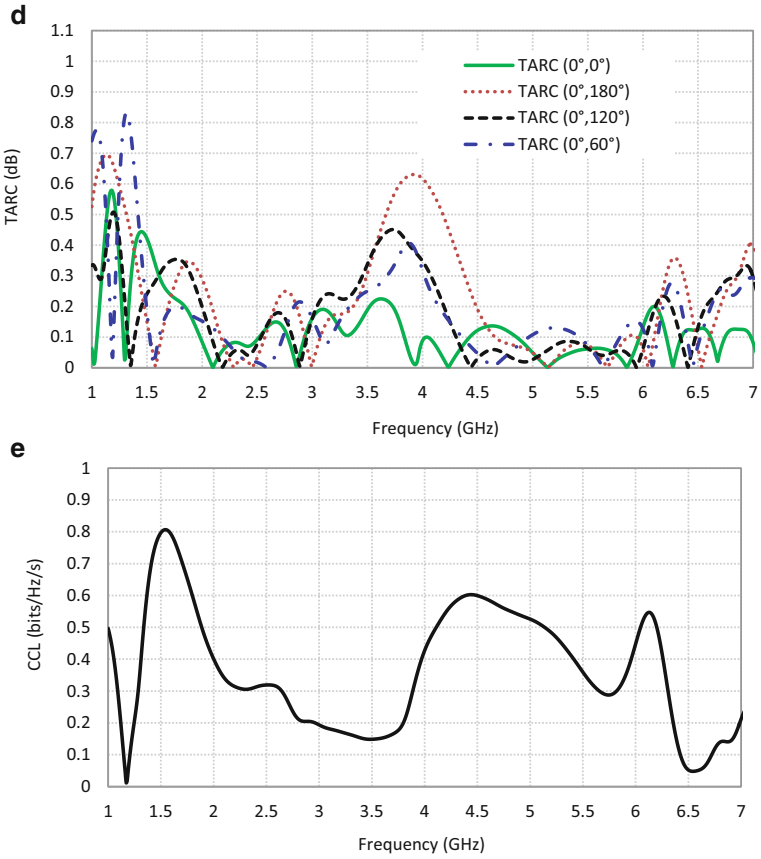

Fig. 10 (**a**) Comparison of simulated ECC with far-field method and with S-parameter; (**b**) Variation of simulated EDG with frequency; (**c**) Variation of MEGs with frequency at different XPR; (**d**) Simulated TARC with several phase difference combination; (**e**) Variation of simulated CCL with frequency

Furthermore, one of the most important performance parameter of diversity system is mean effective gain (MEG) which evaluates the effectiveness of multiple antenna elements system in terms of its average received signal. In other words, MEG is defined as the mean ratio of the power received by the antenna to the total mean incident power. Mathematically, MEG is computed with the equation given below:

$$\mathrm{MEG} = \int_0^{2\pi}\int_0^{\pi}\left[\frac{XPR}{1+XPR}G_\theta(\theta,\phi)P_\theta(\theta,\phi) + \frac{1}{1+XPR}G_\phi(\theta,\phi)P_\phi(\theta,\phi)\right]\sin\theta d\theta d\phi \tag{8}$$

$$\int_0^{2\pi}\int_0^{\pi}\left[G_\theta(\theta,\phi) + G_\phi(\theta,\phi)\right]\sin\theta d\theta d\phi = 4\pi \tag{9}$$

where, XPR is cross-polarization ratio; G_θ and G_ϕ denote the gain components corresponding to angle θ and ϕ; P_θ and P_ϕ are the channel models (angular density functions). The values of P_θ and P_ϕ are assumed as Gaussian in the elevation plane whereas uniform in the azimuth plane and their mean and standard deviation values are taken as 10° and 15°, respectively, for both antenna elements (Pedersen et al. 1997). Furthermore, Fig. 10c illustrates the MEGs for both the antenna elements of proposed antenna with different cross polarization ratio (XPRs) (0 dB, 5 dB and 1 dB). Another important metric that is often utilized to review the diversity performance of MIMO system is MEG. It is defined as the ratio of the received power to the impinging power of the antenna. Here various effects have been considered while analyzing the antenna like radiation power pattern, propagation of the wave and overall efficiency. The ratio of MEGs is almost equal to unity, hence following the equality criterion.

Thereafter, TARC is computed to characterize the proposed MIMO antenna's radiation efficiency. The ratio of square root of total power reflected by the antenna to the square root of total power incident on the antenna is defined as TARC. The mathematical expression for which is given as:

$$\mathrm{TARC} = \frac{\sqrt{\sum_{i=1}^{N} |b_i|^2}}{\sum_{i=1}^{N} |a_i|^2} \tag{10}$$

where a_i and b_i are the incident and reflected signals, respectively. For 2 × 2 MIMO antenna, the relationship between parameter 'a' and 'b' is defined in below equation:

$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \tag{11}$$

Here, the signals are assumed as Gaussian random variable that possess random phase and are independent and identically distributed (i.i.d). So, it can be expressed as follows.

$$b_1 = s_{11}a_1 + s_{12}a_2 = s_{11}a_0 e^{\,j\theta_1} + s_{12}a_0 e^{\,j\theta_2} = a_1\left(s_{11} + s_{12}e^{j\theta}\right) \tag{12}$$

$$b_2 = s_{21}a_1 + s_{22}a_2 = s_{21}a_0 e^{\,j\theta_1} + s_{22}a_0 e^{\,j\theta_2} = a_1\left(s_{21} + s_{22}e^{j\theta}\right) \tag{13}$$

Now putting the values of a and b in Eq. 10,

$$\mathrm{TARC} = \frac{\sqrt{|a_1(s_{11} + s_{12}e^{j\theta})|^2 + |a_1(s_{21} + s_{22}e^{j\theta})|^2}}{\sqrt{2|a_1|^2}} \tag{14}$$

Thereafter, TARC is computed which can be expressed by the following expression (Kaur et al. 2021a).

$$\mathrm{TARC} = \frac{\sqrt{\left|s_{11}e^{j\theta_0} + s_{12}e^{j\theta_1}\right|^2 + \left|s_{21}e^{j\theta_0} + s_{22}e^{j\theta_1}\right)\right|^2}}{\sqrt{2}} \tag{15}$$

The reason for the calculation of TARC parameter is because of the consideration the mutual coupling effects into account which offers more meaningful or accurate analysis of MIMO efficiency than with the conventional scattering matrix method. Figure 10d highlights TARC for 2-element MIMO antenna with several phase difference combinations. Initially, port-1 has been excited by $1e^{j0}$ while keeping the other port at different phases. The legend (0°, 0°) indicates that the elements are in-phase with each other whereas (0°, 60°) means the excited phase difference between radiating elements is equal to 60° and so on. It can be observed from the plots that the curves are quite less than −10 dB over operating band, hence ensuring high performance in actual communications channel environment.

Furthermore, Fig. 10e depicts the channel capacity loss (CCL) which is one of the most important parameter used for characterizing any MIMO system. It is mathematically expressed using the equation given below (Kumar et al. 2018):

$$C_{\mathrm{loss}} = -\log_2 |\psi^R| \tag{16}$$

$$\psi^R = \begin{bmatrix} \rho_{11} & \rho_{12} \\ \rho_{21} & \rho_{22} \end{bmatrix} \tag{17}$$

where ψ^R is the correlation matrix; ρ_{ij} are its elements whose values are computed using *S*-parameters, as indicated by the following equation:

$$\rho_{ii} = 1 - \left(|S_{ii}|^2 + |S_{ii}|^2\right), \rho_{ij} = -\left(S^*_{ii}S_{ii} + S^*_{ji}S_{ij}\right) \tag{18}$$

For stable performance of MIMO system, the expected CCL values must lie under 0.4 bps/Hz. It can be seen from the figure that the CCL values of proposed dual-element MIMO at operating bands has been quite less than that of expected values.

Analysis of Proposed Antenna with Large Ground Plane, USB and Housing

In this section, the simulation study of proposed 4-element MIMO antenna's performance in free space and in actual communication environment is compared. Figure 11a depicts the layout of 4-port MIMO antenna which has area equal to 100 × 82 mm^2 whereas Fig. 11b and c illustrates the top and bottom layer of the near field environment where the antenna is placed within housing among all other components such as USB connector, speaker, an extended ground, battery and LCD. The full view of the device is shown in Fig. 11d. The dimensions of all components are listed in Table 2. To sense the practical wireless communication environment,

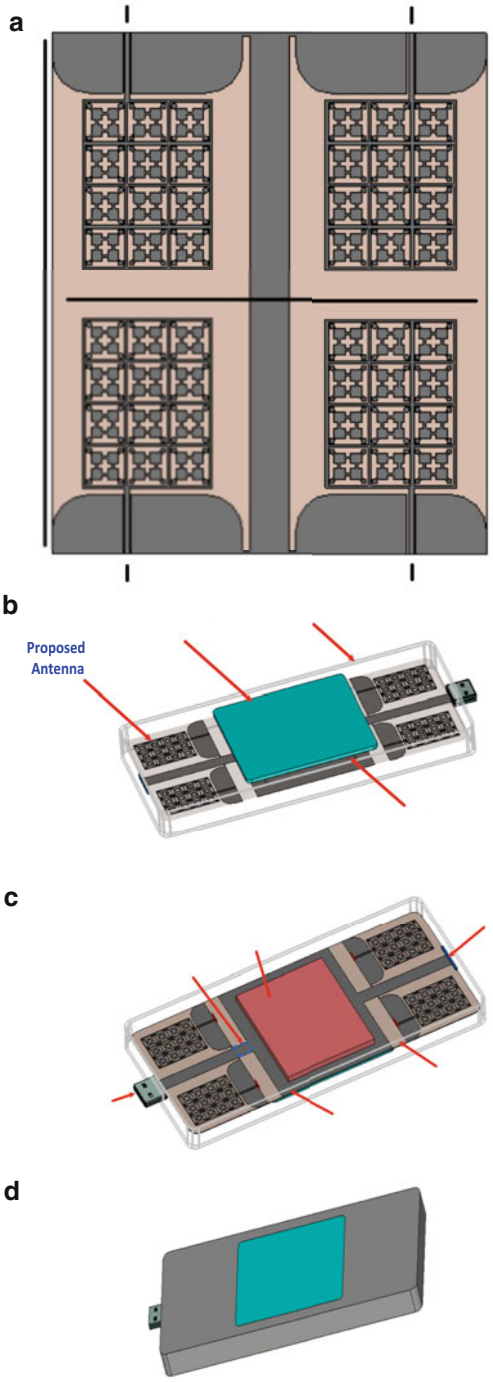

Fig. 11 (**a**) Design of 4-element proposed MTM-inspired antenna; (**b**) Top Layer of the design of 4-element proposed MTM-inspired antenna with actual communication environment; (**c**), Bottom Layer of the design of 4-element proposed MTM-inspired antenna with actual communication environment; (**d**), Overall device view of 4-element proposed MTM-inspired antenna

Table 2 List of parameters of proposed antenna design

Components	Dimensions (mm^3)
Housing	198 × 89 × 1
LCD	83 × 73 × 2
USB connector	12 × 15 × 3.75
Extended ground	74 × 83 × 0.035
Strip	7.5 × 9.5 × 0.035
Battery	67 × 58 × 3
Speaker	18 × 1 × 1.8

LCD and battery have been kept at 5.9 mm spacing but they are connected with the main PCB via a connector of radius 1 mm. The LCD is positioned within the extended ground section as it deteriorates the antenna performance. Then a trade-off between design complexity and size of the display is made by placing all the 4-elements at off-display area i.e. Element-1 and Element-2 on the top-left and top-right corner whereas Element-3 and Element-4 on the bottom-left and bottom-right corner of the upper layer of FR-4 substrate, respectively. Also, a speaker is placed on the same side where antenna elements are placed; however, the bottom layer of the substrate has LCD display. The materials of these components are assumed as copper annealed during simulations and the plastic box used for housing has thickness of 1 mm with dielectric constant of 2.2 and conductivity of 0.05 S/m. All components used in the designed circuitry are placed within the housing.

Further, the effect of free space environment and near-field communication environment on proposed antenna's reflection coefficients and transmission coefficients are highlighted in Fig. 12a and b, respectively. It can be noticed that a significant effect is observed both on the reflection and on the mutual coupling when near field environment is taken into consideration. Both impedance bandwidth and mutual coupling are improved especially at lower frequencies due to the placement of speaker near the antenna elements as the electrical length increases by the capacitive coupling between antenna structure and speaker, and consequently isolation level also get enhanced. Hence, the proposed antenna can be considered as a good candidate for wireless applications as almost negligible effect on antenna's performance in practical communication environment is observed.

Conclusion

In this chapter, MTM-inspired MIMO antenna has been proposed and tested. The study shows that both the simulated and measured results closely match with each other. The diversity characteristics of the proposed radiator are also found within the acceptable limits. Further, the effect of near field communication environment on the performance of proposed radiator is evaluated. To interpret the behavior of practical wireless communication, analysis of S-parameters has been performed on 4-element MIMO antenna which is placed within the device housing among several other components. The results show that the near-field environment has almost

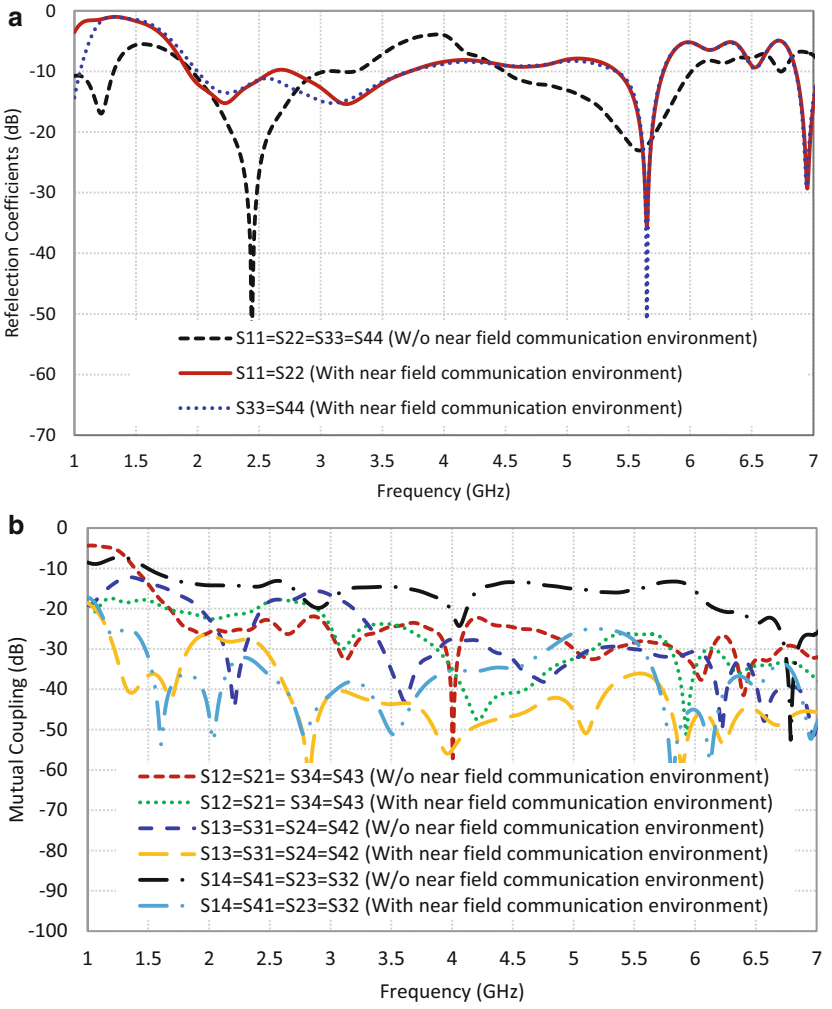

Fig. 12 (**a**) Comparison of simulated reflection coefficients (in dB) with and without near field communication environment; (**b**) Comparison of simulated mutual coupling (in dB) with and without near field communication environment

insignificant effect on the antenna performance, thus makes it a suitable candidate for modern portable wireless communication applications.

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Metamaterials for Antenna Applications

19

V. P. Sarin and K. Vasudevan

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Abstract

The invention of metamaterials or artificial materials with engineered electromagnetic responses and exotic material properties finds applications in a wide range of electromagnetic radiating systems. This chapter summarizes different techniques used to enhance the radiation performance of antennas using metamaterials. Here we discuss the four quadrants representing double positive (DPS),

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double negative (DNG), single negative (SNG), and near-zero-index regions of metamaterials and their corresponding applications in radiation performance of electrically small and standard radiating systems.

Keywords

Metamaterials · SNG metamaterials · DNG metamaterials · Split-Ring resonators

Introduction

Electromagnetic scattering from sub-wavelength particles is an exciting research topic over the last century. When an external electromagnetic wave impinges on an array of subwavelength particles, these particles are said to be polarized. The incident plane wave causes the separation of positive and negative charges on these composites, resulting in electric dipole moments. The circular loop current distributions excited on these composites create the magnetic dipole moments. Hence the resultant scattering will be the superposition of scattering contribution from these electric and magnetic dipole moments (Bohren and Huffman 1983). Figure 1 illustrates the pictorial representation of electric and magnetic dipoles in a material composite. We could achieve unprecedented control of electromagnetic scattering by tailoring the phase and magnitude of these moments.

We can calculate the induced moments by knowing the surface currents induced on the composite, as (Papasimakis et al. 2016),

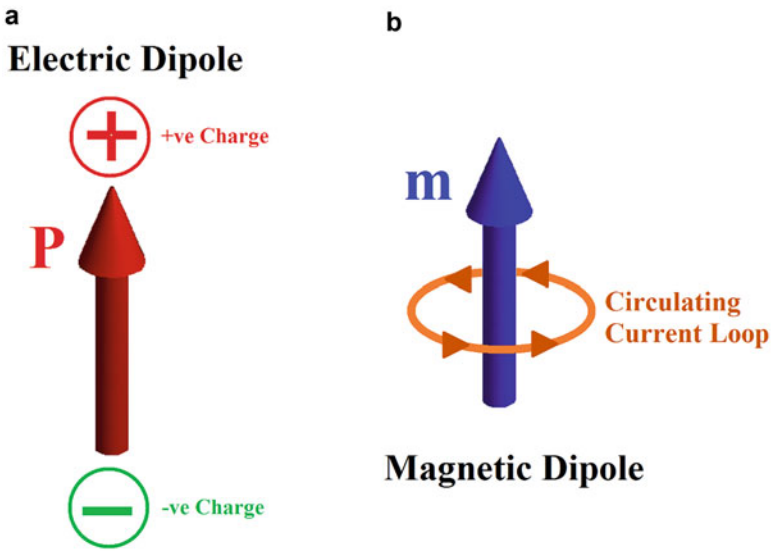

Fig. 1 Creation of dipole moments: (**a**) electric dipole moment and (**b**) magnetic dipole moment

$$\vec{P} = \frac{1}{i\omega}\int \vec{J} d^3 r \tag{1}$$

$$\vec{M} = \frac{1}{2c}\int \left(\vec{r} \mathrm{x} \vec{J}\right) d^3 r \tag{2}$$

where $\vec{P}$ and $\vec{M}$ are the electric and magnetic dipole moments, respectively, $\vec{J}$ is the surface current excited on the unit cell, ω is the angular frequency, c is the velocity of propagation, and $\vec{r}$ is the distance from the origin.

When the induced electric and magnetic dipoles are equal and perpendicular, an in-phase oscillation between them results in scattering in the forward direction with backward scattering suppressed (Bohren and Huffman 1983). The composite then acts as Huygen's secondary source. When these crossed dipoles oscillate out-of-phase, the resulting configuration suppresses scattering both in the forward and backward directions (Kerker et al. 1983). Hence, this composite acts like an invisible body. However, the practical realization of such magneto-dielectric materials was cumbersome due to the unavailability of natural magnetic materials at optical frequencies. In the microwave range, the available magnetic materials show high loss tangent. Hence, their usage is restricted only for low-frequency applications.

The real breakthrough occurred when J.B. Pendry from Imperial College London demonstrated artificial magnetism from an array of split-ring resonators (SRR) (Pendry et al. 1999). He concluded that the SRR array shows a negative permeability response at resonance. This discovery is considered a paradigm shift in the field of electromagnetics. By controlling the amplitude and phase of these moments, we could tailor the reflection/transmission phase and magnitude of electromagnetic waves independently. This discovery has opened a new field of research in electromagnetics of complex media known as "metamaterials." The term "meta" means beyond, and the name "metamaterial" represents artificial structures of metallic and dielectric implants whose electromagnetic properties go beyond that which are readily available in nature. Metamaterials are periodic inclusions of metallic and dielectric implants whose period is sub-wavelength compared to the free-space or guided wavelength. The external electromagnetic wave induces electric and magnetic dipole moments on these composites, modifying the macroscopic ε and μ of the composite. Figure 2 illustrates the classification of materials depending upon their signs of effective ε and μ (Kumar et al. 2020a).

The first quadrant represents double positive (DPS) materials. They exhibit simultaneous positive values of effective ε and μ. They are also known as regular dielectrics. The double negative (DNG) materials represented in the third quadrant exhibit negative values of constitutive parameters. This region supports backward wave propagation as envisioned by Vaselago in 1967 (Vaselago 1967). The epsilon-negative (ENG) and mu-negative (MNG) materials, represented in the first and fourth quadrants, respectively, do not support electromagnetic plane wave propagation. However, they support evanescent waves having imaginary wavenumber. We could also form an additional class of materials named epsilon-near-zero (ENZ) and mu-near-zero (MNZ) media. Here, the material parameters have values near zero. The wavelength of the electromagnetic wave will be infinite in these materials.

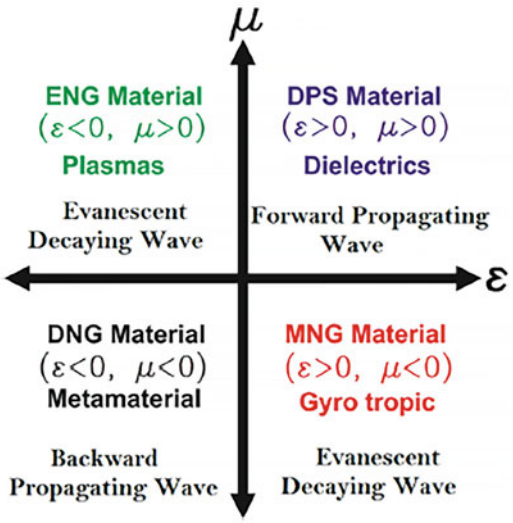

Fig. 2 Classification of metamaterials depending upon the sign of real part of permittivity and permeability. (With permission from Kumar et al. 2020a)

Transmission line representation is a straightforward approach to study electromagnetic wave interactions with metamaterials. In transmission line representation, we can represent each quadrant with the corresponding combinations of distributed inductance and capacitance. Figure 3 shows the transmission line representation of materials depending on their effective values of Re (ε) and Re (μ) (Alù and Engheta 2003; Naqui 2016).

Since electromagnetic wave propagation in a right-handed transmission line and free space (DPS material) are equivalent, the conventional L-C transmission line model is sufficient to represent the first quadrant (Ramo et al. 1994). It supports right-handed propagation, and the phase and group velocities are equal and parallel to the energy flow. The DNG medium in the third quadrant is represented using the dual transmission line form. The dual form is obtained by interchanging the inductance and capacitance of the conventional transmission line. Hence, we can describe DNG materials with the C-L transmission line segment. DNG materials are also known as left-handed media (LHM). The energy and Poynting vector in an LHM are antiparallel to each other as envisioned by Vaselago (Vaselago 1967). Similarly, we could represent ε-negative and μ-negative metamaterials using the L-L and C-C transmission line models, respectively.

ENG Metamaterials

In this section, we review the developments in ENG metamaterials for enhancing the radiation performance of antennas. Before going into detailed discussions, we review the theoretical concepts of ENG metamaterials and their practical demonstrations.

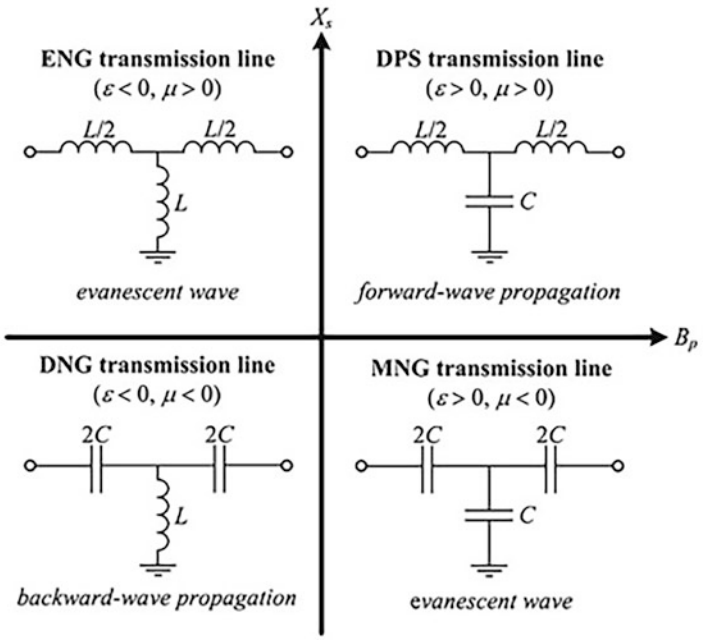

Fig. 3 Transmission line equivalent of materials for different signs of the real part of constitutive. (With permission from Naqui 2016)

ENG behavior is not uncommon in nature. A simple example is a plasma in which the permittivity is negative below the plasma frequency. The well-known examples of plasmas include the earth's atmosphere and noble metals like silver and gold in the infrared and visible frequency regions. We could write the effective permittivity in the form,

$$\varepsilon = \varepsilon_0 \left(1 - \frac{\omega_p^2}{\omega^2}\right) \tag{3}$$

which means that effective permittivity is negative for $\omega < \omega_p$, where ω_p represents the plasma frequency.

We could realize ENG media using parallel plate waveguides and wire inclusions (Rotman 1962). The parallel-wire structure acts as a high-pass filter to an electromagnetic wave whose electric field is oriented parallel to the wire axis. Below the plasma frequency, the effective permittivity is negative. Figure 4a illustrates such a cylindrical metallic array. Fig. 4b shows the dispersive permittivity of the wire medium. It is clear that, above the plasma frequency, the medium is transparent to an incident electromagnetic wave due to positive permittivity values. At plasma frequency, the effective permittivity is zero. We could term the medium as an epsilon-near-zero (ENZ) media near the plasma frequency.

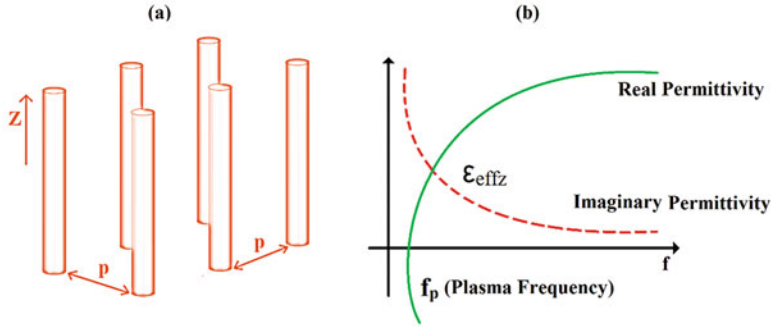

Fig. 4 Demonstration of artificial plasma using conducting wires: (**a**) array of running wires and (**b**) resultant effective permittivity

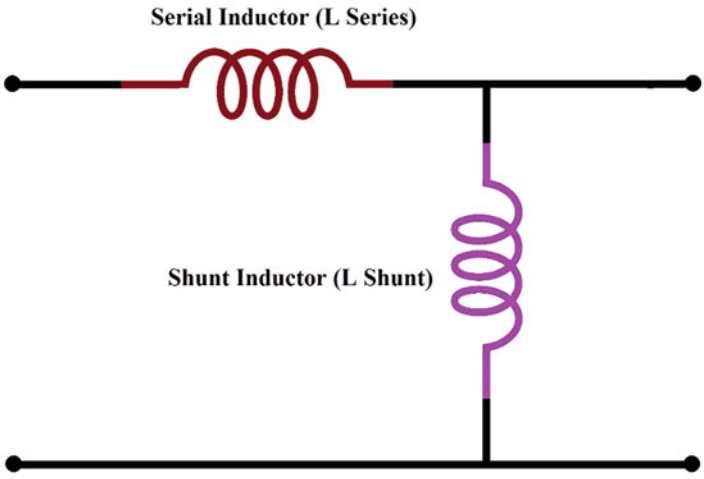

Fig. 5 *L-L* Transmission line model to represent ENG metamaterials

The L-L transmission line model is used to represent an ENG media in the electrical circuit domain. Figure 5 depicts an infinitesimal portion of the L-L transmission line under consideration. If the shunt and series inductance L_{shunt} and L_{series} are uncoupled, the mutual inductance is zero. Then, we can write the wave equations of current and voltage as (Alù et al. 2007),

$$\frac{d^2I}{dz^2} = \frac{L_{\text{series}}}{L_{\text{shunt}}} I \tag{4}$$

$$\frac{d^2V}{dz^2} = \frac{L_{\text{series}}}{L_{\text{shunt}}} V \tag{5}$$

This line does not allow the propagation of electromagnetic plane waves as long as L_{shunt} and L_{series} are positive. Similar to a standard L-L transmission line, it supports the propagation of only evanescent waves. This is the reason that we use the ionosphere for long-distance radio communications in the HF bands. The ionosphere acts like a dielectric plasma characterized by the plasma frequency. It is opaque to incident electromagnetic waves below plasma frequency and reflects the signal towards the ground station. Satellite communications are made possible by selecting the transponder frequencies above the plasma frequency for which the ionosphere is transparent.

ENG Shell for Efficient Electrically Small Antennas

Electrically small antennas (ESA) is a rapidly growing research topic due to their compact nature, which finds applications in today's emerging wireless communication systems, especially in wireless sensor networks and imaging applications. ESA's main drawback is their low radiation efficiency, and hence they are inefficient radiators of electromagnetic power. The small radiation resistance and the simultaneous large capacitive reactance result in poor impedance matching with the source. The design of an impedance matching circuitry is a complicated procedure. The incorporation of matching circuitry with the antenna system badly affects the radiation performance (Balanis 2016).

Before going into detail, let us briefly discuss an electrically small radiator and its characteristics. The power radiated from an electrically small dipole source could be written as (Ziolkowski and Kipple 2003),

$$P = \eta\left(\frac{\pi}{3}\right)\left|\frac{I_0}{\lambda}l\right|\left[1 - j\frac{1}{(kr)^3}\right] \tag{6}$$

where η is the wave impedance, I_0 is the driving current, λ is the operating wavelength, l is the length, k is the free space wavenumber, and r is the radius. It is clear that this radiator has a large capacitive near-field, and hence it is an inefficient radiator. Reactive near-field indicates the presence of electric and magnetic fields, which are spatially and temporally phase-shifted. In a capacitive near-field, the electric field lags the magnetic field by a temporal phase shift. If this temporal phase shift is 90°, then it represents an ideal evanescent source and shows exponential decay along the direction of propagation. In other words, this source is mismatched to free space.

Various ESA designs are available in the literature (Ziolkowski and Erentok 2006). An important parameter that better describes the operation of an ESA is its "ka" value. Consider an ESA surrounded by the smallest sphere of radius "a" with an operating frequency of f_0; then, the antenna is electrically small if its $ka \leq 1$ where $k = 2\pi/\lambda_0$ and λ_0 is the operating wavelength. If the antenna contains a ground plane, we can use the limit $ka \leq 0.5$ and consider only half of the criterion's radiation sphere.

A convenient way to match ESA's high capacitive reactance is to use matching networks, as shown in Fig. 6. The high capacitive reactance could be conjugate matched using an inductive reactance to bring the resonant zero impedance condition (Smith 1977). The matching network should also ensure that the total resistance of the system is 50 Ω. We could construct the matching network using passive lumped elements. However, the high current and voltages in these network elements cause a significant power loss yielding a low radiation efficiency and bandwidth.

We could avoid the use of passive and active components for impedance matching with metamaterials having ENG response. The ENG shell distributed around an infinite small dipole source can overcome high capacitive reactance, thereby bringing resonance due to the cover's high inductive reactance. Richard Ziolkowski used an ENG shell to match an electrically small antenna with free space (Ziolkowski and Erentok 2006). Their ultimate goal is to achieve near-unity radiation efficiency and maximum radiated power from the system.

Figure 7 illustrates the resulting dipole-metamaterial configuration. Here the electrically small electric dipole is situated in a homogeneous region at the center of the ENG shell represented using the red shaded area. The spherical shell has a negative real part of permittivity. We need to introduce one important parameter to quantify this composite known as the radiated power ratio (RPR). RPR is the ratio of far-field radiated power with and without the shell expressed in decibels. We can describe it as,

$$RPR = 10\log_{10}\left[\frac{P_{\text{with shell}}(1A \text{ input current})}{P_{\text{without shell}}(1A \text{ input current})}\ (dB)\right] \tag{7}$$

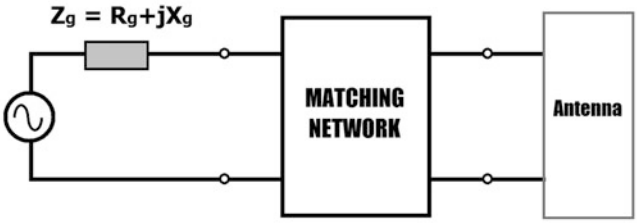

Fig. 6 Schematic of an antenna impedance matching network

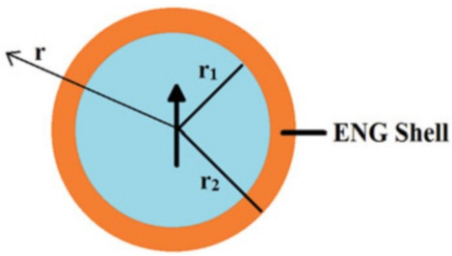

Fig. 7 Geometry of the dipole-ENG shell configuration. (With permission from Ziolkowski and Erentok 2006)

Figure 8 illustrates the variation of RPR against the real part of the dielectric constant of the surrounding shell. For all these variations, the length of the antenna element is as $l = 10$ mm. As the real part of permittivity increases, we need a lower radius of the outer shell to achieve maximum far-field radiated power.

As already mentioned, the ESA in effect acts as a capacitive element, and the cylindrical ENG shell surrounding the antenna acts as an inductive element. Figure 9

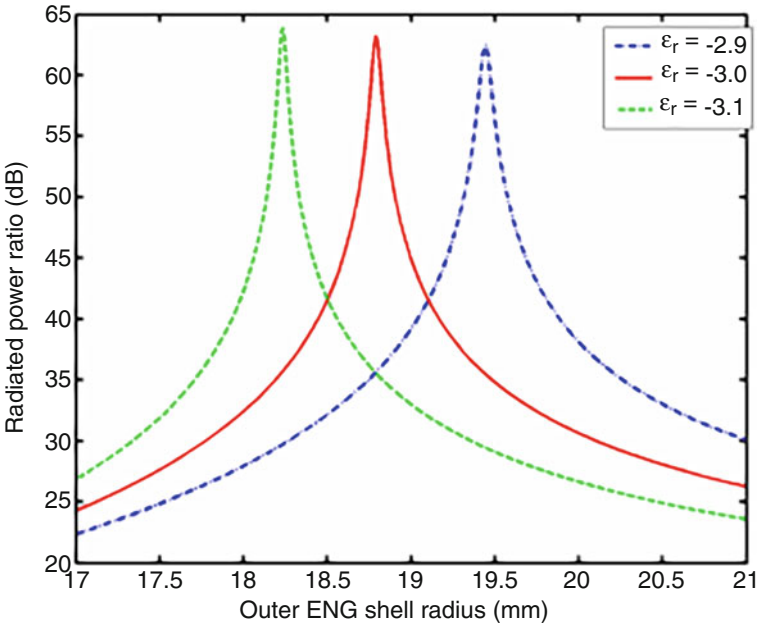

Fig. 8 Radiated power for different ENG shell radius for different dielectric constants configuration. (With permission from Ziolkowski and Erentok 2006)

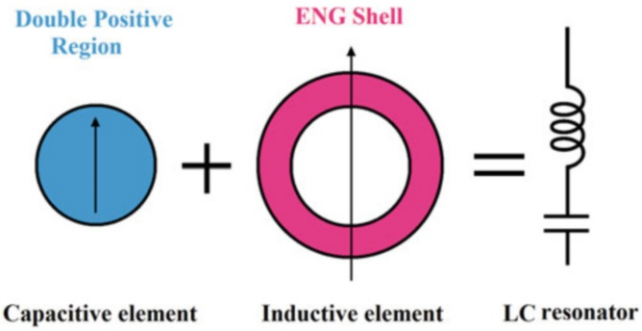

Fig. 9 Equivalent circuit representation of the ESA-ENG metamaterial configuration. (With permission from Ziolkowski and Erentok 2006)

shows the equivalent circuit representation of this configuration. The ENG shell offers a reactance,

$$X_{\text{shell}} = j\omega_0 L_{\text{eff}} \propto j\omega_0 \frac{1}{\varepsilon \Delta r \omega_0^2} \tag{8}$$

where Δr represents the shell's thickness, and it suggests that the ENG shell offers an inductive reactance. We can control its magnitude by varying the shell's thickness, permittivity, and operating frequency. The capacitive reactance of the ESA in the presence of the inductive reactance offered by the ENG shell forms a series resonance LC circuit, and its resonant frequency is,

$$\omega_{\text{res}} = \frac{1}{\sqrt{L_{\text{eff}} C_{\text{eff}}}} \propto \omega_0 \sqrt{\frac{|\varepsilon| \Delta r}{C_{\text{eff}}}} \tag{9}$$

For the practical demonstration of the above mentioned theoretical technique, it is better to use a monopole antenna over a ground plane instead of the line current element. The advantage is that we can model a coaxial feed line having 50 Ω characteristics impedance in simulation and can be easily implemented in experiments. Figure 10 shows the geometry of this configuration. The monopole element length is $l = 5$ mm, and in simulation, TEM waves are fed to the waveguide port of the coaxial probe.

The coaxial line's center conductor connects the cylindrical monopole element having a radius of $r = 0.1$ mm. The outer conductor of the coaxial connector connects the flat horizontal ground planes. The inner radius of the ENG shell is $r_1 = 10$ mm, and the outer radius is $r_2 = 18.45$ mm. The frequency-independent constitutive parameters of the cover are $(\varepsilon, \mu) = (-3\ \varepsilon_0, \mu_0)$. This antenna's resonance behavior can be analyzed using the feed port's reflection coefficient, as depicted in Fig. 11. The antenna shows a well-defined resonance and achieves a maximum

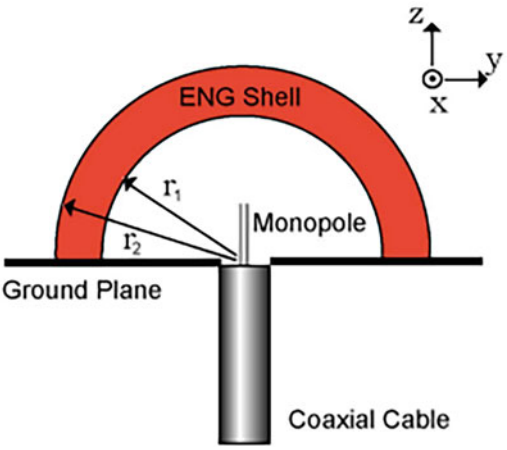

Fig. 10 Practical design of ENG shell incorporated monopole antenna. (With permission from Ziolkowski and Erentok 2006)

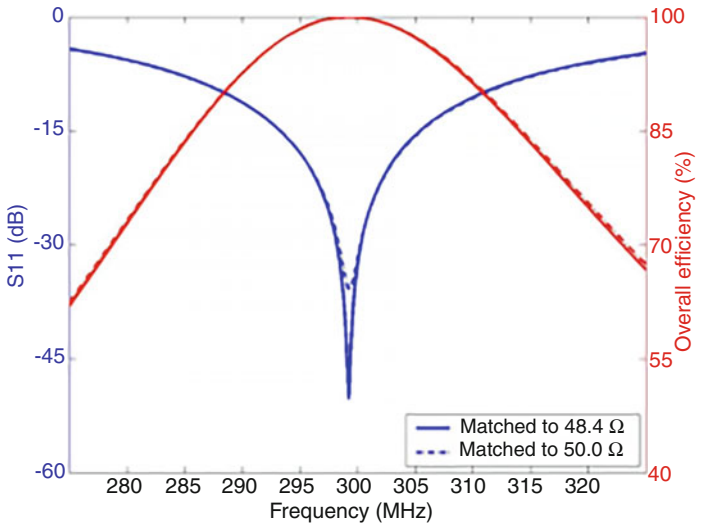

Fig. 11 Characterization of the monopole-ENG shell configuration. (With permission from Ziolkowski and Erentok 2006)

efficiency of 100%. In simulations, the ENG shell surrounding the antenna is loss-free and dispersionless. There would be finite losses in practical situations, and hence one could not achieve near-unity radiation efficiency. The radiation patterns of the antenna match well with standard monopole radiators backed by a metallic conductor. The added advantage is that the radiator's efficiency is enhanced significantly. The presence of losses and dispersion in the ENG shell causes a narrower frequency response compared to the loss-free case.

ENG Shell for Cloaking a Receiving Antenna or a Sensor

Another vital application of ENG metamaterials is the scattering reduction from standard electromagnetic radiators and sensors. For this, we need to cover the radiator or the sensor with specially designed plasmonic cloaks. The impinging electromagnetic waves will be penetrated through the cover so that the receiving/sensing capability will not be disturbed. A specially designed plasmonic cover suppresses scattering from the receiver and is undetectable from the outer world. However, the receiver can "sense" the surroundings while being invisible to the external world. Figure 12 illustrates such a sensing strategy (Alù and Engheta 2009). The primary observation is that the conventional receiving antenna receives power from the impinging electromagnetic plane waves. It scatters a part of the incoming electromagnetic energy to its surroundings. So, a radar can detect the antenna from the far-field. In the cloaked scenario, the impinging electromagnetic plane wave penetrates the cloak, as depicted in Fig. 12b. However, the scattering from the antenna or sensor is suppressed. This technique primarily differs from other cloaking

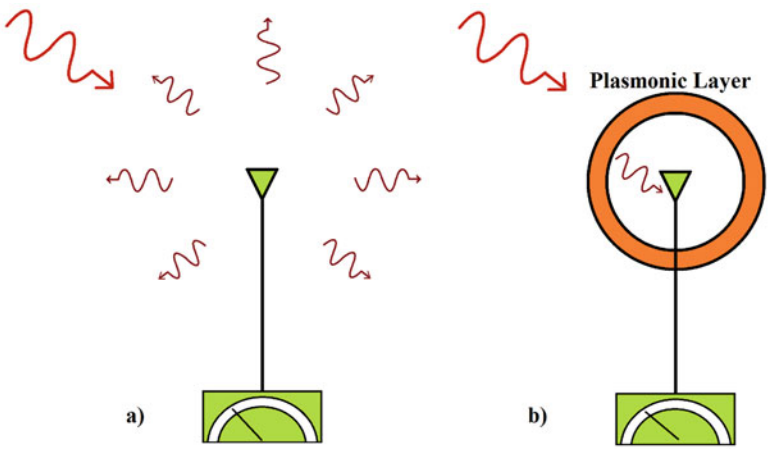

Fig. 12 Scattering characteristics of antennas in receiving mode: (**a**) conventional receiving antenna and (**b**) cloaked receiving antenna. (With permission from Alù and Engheta 2009)

technique is that in conventional cloaking techniques based on coordinate transformation, the impinging electromagnetic wave on the cloak is routed around the target, and there is no direct interaction between the wave and the target (Leonhardt 2006).

The receiving antenna or the sensor is characterized by its polarizability "α," which determines how electric dipole moment "P" is created from the incident electric field "E." We can relate them as P=αE. A higher "α" implies a better receiving/sensing and transmitting capability of the antenna. We could write the polarizability of a loaded short dipole as (Tretyakov et al. 2003)

$$\alpha = i\frac{4Z_{\text{in}} + Z_L}{Z_{\text{in}} + Z_L}\frac{l^2}{l2\omega Z_{\text{in}}} \tag{10}$$

where Z_{in} is the input impedance of the dipole, l is its length, and Z_L is the load impedance. When $Z_{\text{in}} = Z_L^*$, then the conjugate impedance matching condition is achieved. Then the receiving antenna will extract maximum power from the impinging electromagnetic field. This condition gives the largest value for α. The problem is that this condition provides the maximum scattering cross section (SCS). Under this condition, a receiving antenna can absorb and scatter maximum power, and hence it is visible to its surroundings. One technique to reduce resonant scattering from the dipole antenna is modifying the load reactance, but this technique reduces the received power.

So plasmonic cloaking in an efficient way to cloak a receiving antenna. In plasmonic cloaking, we can design the effective dielectric constant of the cloaking layer such that the incident electric field will induce a significant dipole moment P_C on the cover such that $P_C = -P_s$, where P_s represents the induced moment on the receiving antenna. The scattered power from the antenna and the cover will then be out of phase, reducing the scattering cross section. Hence, the antenna will be

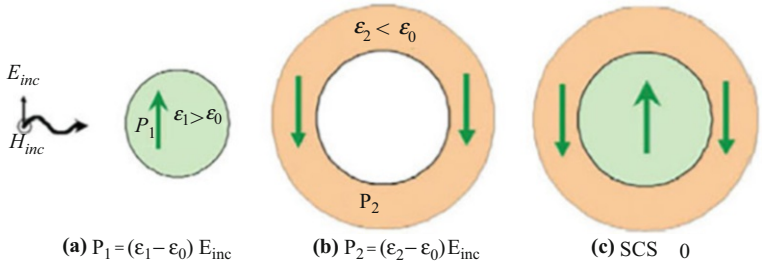

Fig. 13 Basic idea of plasmonic cloaking scheme. (With permission from Rezaei et al. 2021)

invisible from the surroundings. So plasmonic cloaking is a scattering cancellation-based cloaking technique in which the antenna's receiving capability is undisturbed.

The physical mechanism characterizing the plasmonic cloaking scheme is depicted in Fig. 13 (Alù and Engheta 2005). Three cases are considered here for study. The first one is the scattering from a dielectric sphere characterized by its permittivity ε_{DPS}, the second one is an ENG shell (ε_{DPS}), and the third is the combination of the DPS sphere and ENG shell. The regular dielectric sphere scatters electromagnetic waves due to the creation of electric dipole moments in it. Its polarization vector is given by $P = (\varepsilon - \varepsilon_0)E_{inc}$. Similarly, the ENG cover with $\varepsilon < \varepsilon_0$ scatters electromagnetic waves due to the induced polarization vector $P = (\varepsilon - \varepsilon_0)E_{inc}$. However, when both these materials are combined, the total scattering coefficient becomes zero due to the layers' antiparallel scattered field contribution. The ENG cover radius can be adjusted to suppress scattering completely, as shown in the final diagram (Rezaei et al. 2021).

For the practical demonstration, they constructed two prototypes. The first one is the uncloaked dipole having resonance at 1 GHz and is operated in the receiving mode in simulations. In the second design, the plasmonic cloak shell in which a Drude dispersive material is placed around the dipole antenna.

Both the structures are illuminated with a plane wave with polarization oriented along the axis of the antenna. The Poynting vector distributions observed are illustrated in Fig. 14. It is clear that the bare dipole antenna scatters the incident electromagnetic wave, and a turbulent distribution is observed around the antenna. In the case of a cloaked cylinder, a smooth flow of electromagnetic power flow is observed. In the case of the cloaked cylinder, the electromagnetic power flow inside the plasmonic shell is not perturbed, and it means that the receiving capability of the dipole is unaffected.

Other important parameters to be considered for scattering studies are the radar cross section (RCS) and scattering cross section (SCS). RCS or alternately the backscattering cross-section is defined as

$$\sigma = \lim_{R \to \infty} 4\pi R^2 \frac{|E_{sca}|^2}{|E_{inc}|^2} \tag{11}$$

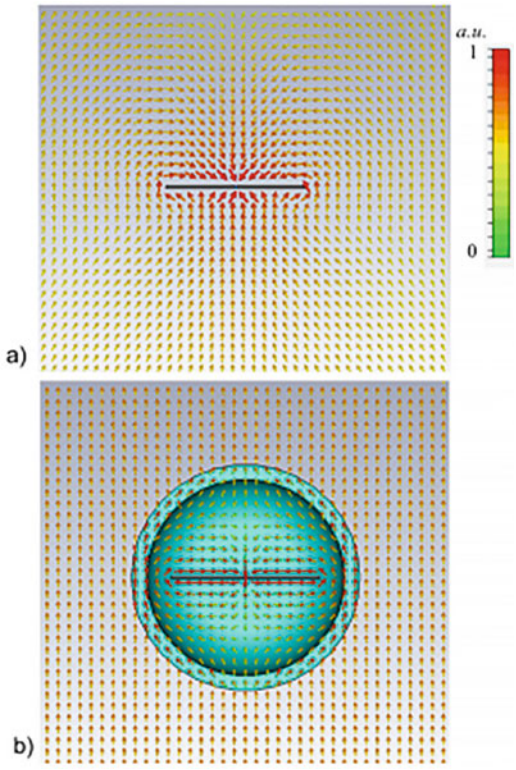

Fig. 14 Poynting vector distributions in the computational domain: (**a**) uncloaked dipole and (**b**) cloaked dipole. (With permission from Alù and Engheta 2009)

where R is the distance from the antenna to the far-field observation point, E_{inc} is the incident electric field measured at the position of the antenna, and E_{sca} is the scattered electric field measured at the far-field point. If the RCS of the antenna is zero, the scattered power from it at the far-field observation point is zero. In RCS measurements, only one far-field observation point is considered at a time. A more useful characterization of far-field scattering is the SCS of the target. It is defined as

$$\sigma_{\text{total}} = \frac{1}{4\pi} \int \sigma d\Omega \tag{12}$$

where the integration of σ is performed over the whole solid angle around the target. The scattering patterns of the cloaked and uncloaked antennas are shown in Fig. 15. More than 1500 folds reduction in the SCS of the cloaked antenna is observed compared to a bare dipole at resonance. A significant reduction in scattering is achieved, which means that the antenna covered with the plasmonic cloak is invisible from its surroundings.

The practical realization of scattering reduction from Yagi-Uda type antennas operating at UHF frequencies is also made possible by adopting the mantle cloaking

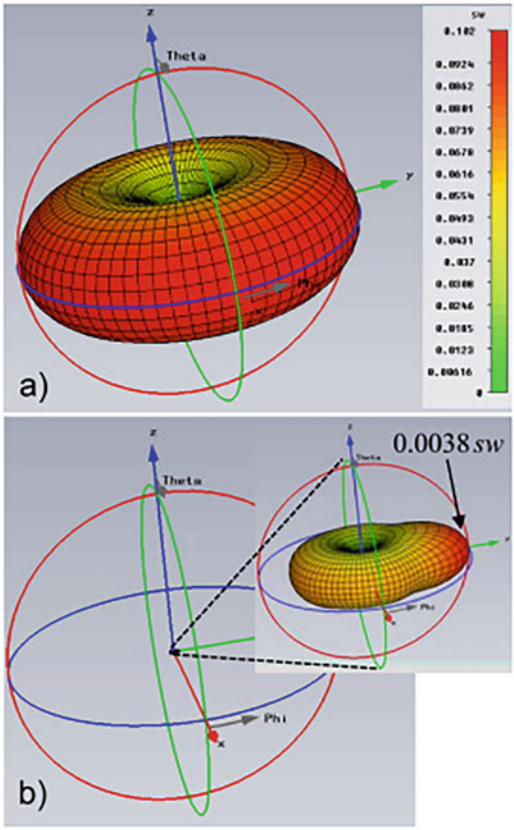

Fig. 15 Scattering patterns of the antennas: (**a**) bare dipole and (**b**) cloaked dipole. (With permission from Alù and Engheta 2009)

technique. The application of a cloaking layer over the antennas significantly reduces scattering in the S-band frequencies. The first step involved is the design of the Yagi–Uda antenna (Monti et al. 2016). This study is performed on a three-element Yagi design having parameters $L_r = 0.52\lambda_0$, $L_c = 0.47\lambda_0$, and $L_d = 0.43\lambda_0$ where λ_0 is the wavelength. The second step is to design the mantle cloak that is to be covered over the three antenna elements to suppress scattering. The working principle of mantle cloaking is to develop proper surface impedance of the cloaking layer to suppress scattering. For metallic objects, the cloaking layer should offer an intrinsic capacitive surface reactance to nullify scattering. For achieving capacitive surface reactance, one needs to arrange metallic layers around the three antenna elements. Some optimization procedures must be done after implementing the cloak layer around the antenna to restore antenna resonance and reduce the matching of unwanted resonances. The corresponding uncloaked and cloaked antennas are depicted in Fig. 16.

The measured reflection and radiation characteristics of the cloaked and uncloaked designs are shown in Fig. 17. Incorporating the cloaking layer does not

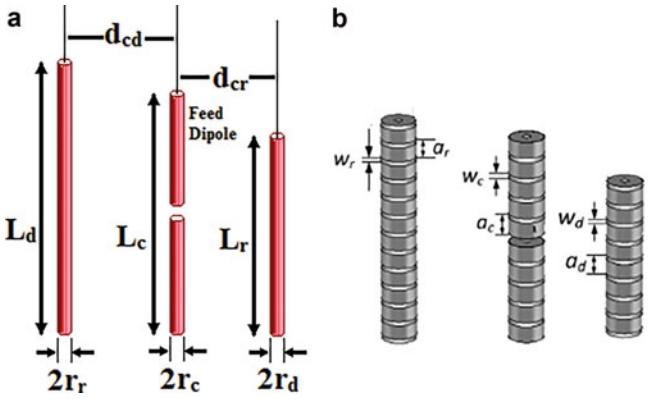

Fig. 16 Practical demonstration of Yagi-Uda antenna cloaking: (**a**) uncloaked Yagi and (**b**) cloaked Yagi. (With permission from Monti et al. 2016)

change the resonant frequency after the optimization process. The dotted line in Fig. 17a shows the −10 dB matching threshold for an antenna. The antennas' gain in the forward and backward direction also remains unaltered for the cloaked antenna, as indicated in Fig. 17b. The SCS of the cloaked and uncloaked antennas are also taken for a far-field illumination plane wave with polarization parallel to the metal wires. The SCS of the antennas is illustrated in Fig. 17c. Since the cloak is designed around 3.2 GHz in the S-band, a scattering reduction of 14.6 dB order is obtained at the designed frequency. The −10 dB SCS reduction bandwidth is 400 MHz.

Plasmonic Materials for Optical Nanoantennas

Nano sources like quantum dots and other emitters exhibit radiation characteristics like small electric dipoles. Electronic transitions in these systems are purely electric dipolar by nature. The contribution from magnetic dipole and quadrupole dipoles are significantly weaker in nano sources. Magnetic dipole transitions are approximated to be 10^5 times weaker than electric dipole transitions (Taminiau et al. 2012). So, the optical frequency regime is considered as a world of electric fields and electric dipoles. The major problem with electric dipole type emitters is that they are perfectly mismatched with free-space due to their sub-wavelength size. Figure 18 shows the equivalent circuit of an electric dipole source (Chu 1948). It represents a high-pass filter circuit. As the radius "a" of the spherical surface surrounding the antenna becomes smaller than the wavelength, the circuit becomes purely reactive due to the short circuit offered by the inductive stub. So conventional nano sources are mismatched to air due to the high capacitive reactance. So, only a small portion of the incident power is delivered to the load resistance.

So an intermediate element is needed to compensate for the high capacitive reactance. An optical nanoantenna acts as the compensation element. The inductive reactance due to the plasmonic material cancels the excess capacitive reactance.

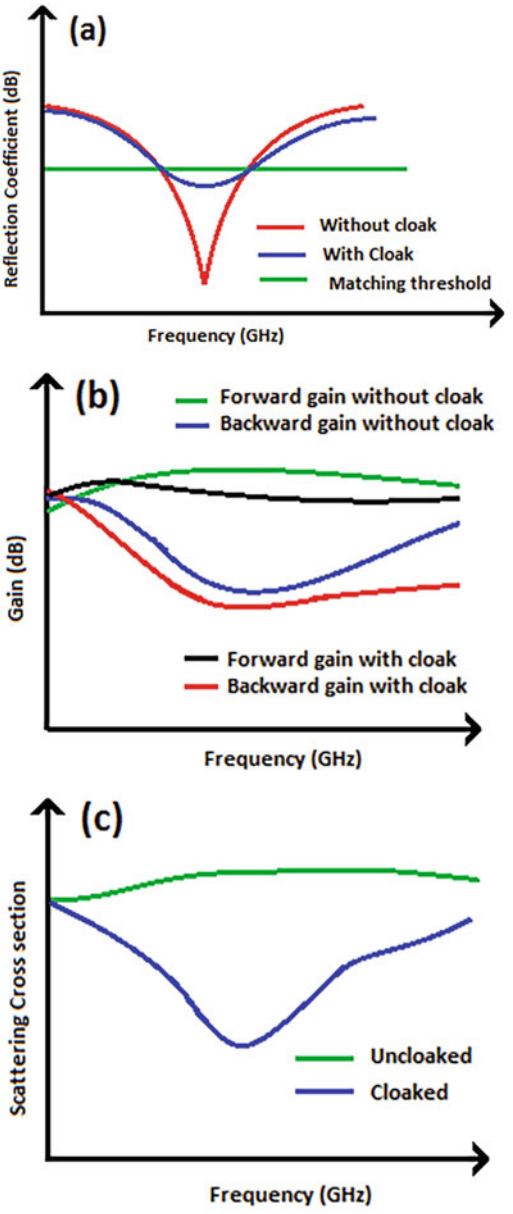

Fig. 17 Characterization of cloaked and uncloaked antennas: (**a**) reflection coefficients, (**b**) gain, and (**c**) scattering cross section. (With permission from Monti et al. 2016)

A nanoantenna thus converts the localized or guided optical fields into efficient far-field radiation.

In the microwave regime, metallic conductors are used for the construction of dipole sources. However, in the optical regime, the situation is quite different.

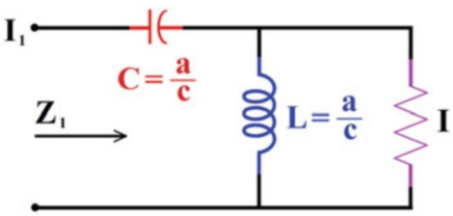

Fig. 18 Equivalent circuit of an electric dipole source. (With permission from Chu 1948)

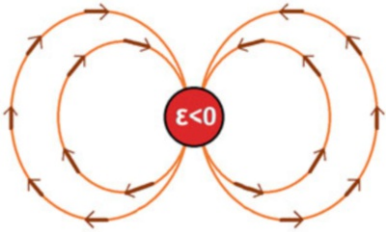

Fig. 19 Equivalent circuit of a plasmonic nanosphere. (With permission from Engheta et al. 2005)

Ordinary metals obey the Drude dispersion model in the visible and UV ranges. Near the plasma frequency, the real part of permittivity will become small and negative. So, the fields will penetrate the plasmonic material creating an enhanced reactive response. The inertial mass of the electrons contributes additional kinetic inductance at these frequencies. So, a nanoantenna can resonate even though its dimensions are deep sub-wavelength. This kinetic inductance dominates over the "Faraday inductance" offered by current paths.

For a sub-wavelength nanoparticle, the optical impedance is defined as the ratio of local potential difference $V = |E|l$ and displacement current $I_d = i\omega\varepsilon\varepsilon_0 |E|S$ through the particle (Monticone and Alù 2017; Monticone et al. 2017). This parameter is independent of the surrounding environment. Hence the nano-capacitance offered by a dielectric nanosphere of radius "a" can be written as

$$C_{\text{sphere}} = \pi a \varepsilon_0 \,\text{Re}\,(\varepsilon) \tag{13}$$

A plasmonic nanosphere could be treated as equivalent to a nano inductor in which the inductance is given by

$$L_{\text{sphere}} = \left[-\omega^2 \pi a \varepsilon_0 \,\text{Re}\,(\varepsilon)\right]^{-1} \tag{14}$$

The plasmonic nanosphere is represented using the equivalent nanocircuit form, as shown in Fig. 19. Here the nanoinductor represents the particle, and the parallel capacitance denotes the surrounding dipolar fringe field. The resistance is included in the figure to account for losses, and the current source represents the local applied field. This equivalent circuit representation predicts a resonance due to the parallel

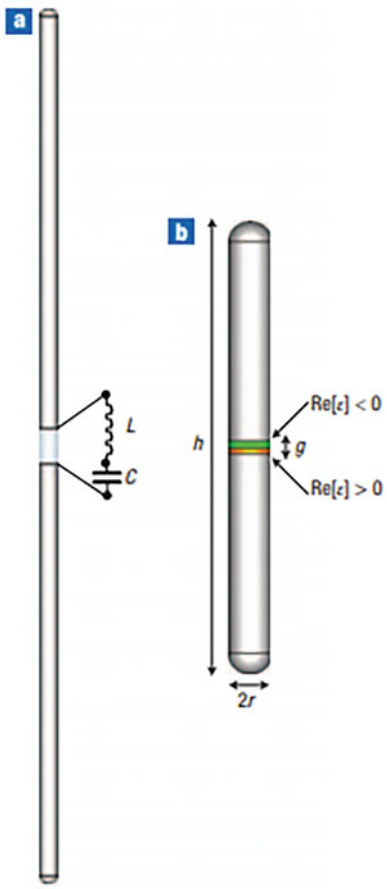

Fig. 20 Analogy between an RF dipole and a plasmonic nanoantenna: (**a**) LC loaded RF dipole and (**b**) nanoantenna loaded with a plasmonic particle at the feed gap (Alù and Engheta 2008)

combination of inductance and capacitance. It is worth noting that a pure sub-wavelength dielectric nanoparticle could not resonate due to the high capacitive reactance.

This technique can be readily applied for creating resonance and tuning of plasmonic nanoantennas (Alù and Engheta 2008). Figure 20 shows the analogy between the loading of an RF dipole antenna and a nanoantenna loaded with plasmonic particles. In a conventional RF dipole antenna, the scattering can be controlled by loading reactive components at the feed gap. Since it is impossible to create lumped elements at the optical domain, loading is achieved by using optical nano inductors and capacitors formed by plasmonic materials. These materials are loaded at the nanoantenna's feed gap, whose parameters can be varied to control impedance matching. The specialty is that we can thus control scattering from these nanoantennas.

MNG Metamaterials

In electromagnetic theory, a circulating current loop is considered to be the fundamental cause behind magnetism. The magnetic dipole moment is defined as the product of the magnitude of the loop's current and area. When an external electromagnetic field is allowed to impinge on an array of such metallic loops, circulating currents will be excited. The resulting magnetic flux will oppose the externally applied field. It means that such an array shows diamagnetic behavior (Weber 1852). The magnetic polarizability of a conducting ring place along the XY plane of the coordinate system is given by

$$m_z = \alpha_{zz}^{mm} B_z^{ext} \text{ where } \alpha_{zz}^{mm} = -\frac{\pi^2 r^4}{L} \tag{15}$$

Here L is the self-inductance of the ring, r is its radius, B_z^{ext} is the Z-component of the external applied magnetic field, and α_{zz}^{mm} is the polarizability. But the diamagnetism offered by this closed ring is negligibly small to create an effective negative μ. However, the ring's polarizability can be significantly enhanced by loading this loop with a capacitor (Shelkunoff and Friis 1966). The modified polarizability could be written as

$$\alpha_{zz}^{mm} = \frac{\pi^2 r^4}{L}\left(\frac{\omega_0^2}{\omega} - 1\right)^{-1} \tag{16}$$

From the equation, it is clear that loading the capacitor results in resonance, whose resonant frequency could be found out using the equation $\omega_0 = \frac{1}{\sqrt{LC}}$. The frequencies just above the resonant frequency show negative values of polarizability. However, in the microwave regime, the best possible method to achieve a negative μ response is to construct SRR arrays using simple planar photolithographic manufacturing techniques. The split-ring resonator array exhibits strong permeability resonance and its unit cell is shown in Fig. 21. The split-ring resonator consists of two concentric metallic split-rings engraved on a suitable substrate. When a time-varying electromagnetic wave with a magnetic field aligned parallel to the SRR, then resonant currents will be excited on the rings. It will create a magnetic field that will either suppress or enhance the incident field. The presence of the splits makes the resonant frequency subwavelength as compared to the incident wavelength due to the enhanced capacitance offered by the splits. SRR can be excited in two ways; one

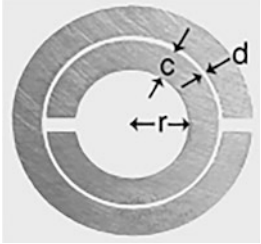

Fig. 21 Unit cell of the split-ring resonator for obtaining negative permeability response. (With permission from Smith et al. 2000)

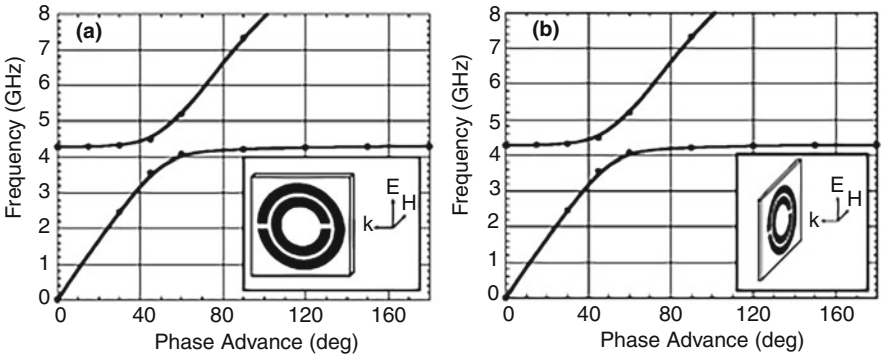

Fig. 22 Different excitation scenarios of SRR array and corresponding dispersion diagrams: (**a**) $H_{\|}$ excitation and (**b**) $H_{\perp}$ excitation response. (With permission from Smith et al. 2000)

is called the $H_{\|}$ excitation. Here, incident E-field is oriented parallel to the splits, the H-field is oriented parallel to the rings' axis, and propagation direction is perpendicular to the axis of the SRR. The second one is the $H_{\perp}$ excitation in which the external E-field is oriented parallel to the splits, and H-field is perpendicular to the SRR axis with the direction of propagation along the axis of the SRR. These two excitation scenarios and the corresponding dispersion characteristics are plotted in Fig. 22 (Smith et al. 2000).

For both these excitations, a propagation region is identified up to the lower band edge, followed by a bandgap. For $H_{\|}$ excitation, the bandgap is identified with negative permeability resonance in which the effective permeability $\mu_{\text{eff}}(\omega)$ is negative and $\varepsilon_{\text{eff}}(\omega)$ is positive and slowly varying. Similarly, for $H_{\perp}$ excitation, the bandgap is characterized by permittivity resonance in which the effective permittivity $\varepsilon_{\text{eff}}(\omega)$ is negative. A split-ring resonator array working under $H_{\|}$ excitation can be approximated as an anisotropic medium characterized by an effective permeability given by [Pendry and Holden 1999]

$$\mu = \mu_0 \left(1 - \frac{f^2{}_p}{f^2 - f^2{}_0} \right) \tag{17}$$

where f_p is the plasma frequency, the resonant frequency is f_0, and μ_0 is the permeability of free space. It is evident from the equation that the medium's effective permeability is negative above resonant frequency f_0. A typical plot of permeability response of the SRR array is illustrated in Fig. 23.

Miniaturized Patch Antenna Using MNG Metamaterial Loading

Microstrip patch antennas are highly desirable for wireless communications due to their relative ease of fabrication, lightweight, etc. Even though they are thin, their

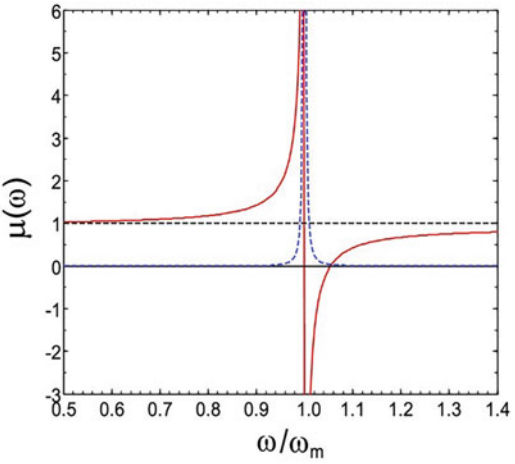

Fig. 23 Permeability response of an SRR array under $H_{||}$ excitation (red Line indicates the real part and the blue dotted line indicates the imaginary part)

transverse dimensions are of the order of half wavelengths. So many techniques are normally employed for squeezing the aperture area of the patch radiator. The conventional methods use high dielectric constant materials or magneto-dielectric materials as the patch substrate (Ikonen et al. 2006). But the miniaturization process significantly deteriorates the radiating performance of the antenna.

A sub-wavelength patch antenna is weakly coupled to free space and shows a significantly reduced radiation efficiency. To overcome this problem, Alù and Engheta proposed paired combinations of DPS/SNG layers as the patch substrate (Alù et al. 2007). Their study showed that rectangular or square patch antenna loaded with DPS/SNG combination is not the right candidate for radiation applications. The designed antenna offers good impedance matching performance, but the radiation patterns are distorted. They showed that circular patch antennas loaded with DPS/MNG materials are good candidates for sub-wavelength miniaturized and efficient radiators.

Filiberto Bilotti showed that the sub-wavelength patch antenna would work even when the MNG sample is not isotropic (Bilotti et al. 2008). In their design, a single ring of SRR is placed beneath the patch antenna where the magnetic field is maximum. The problem is that the substrate material used has a thickness of about $\lambda/80$ and the dimensions of the standard SRR is of the order $\lambda/20$. Hence, we cannot use standard SRR in this design. The authors used multiple split-ring resonators (MSRR) or spiral resonators (SR) to overcome the size problem. The SR can reduce the dimension of the MNG inclusions to the order of $\lambda/100$. Figure 24 illustrates the geometry of the SR and SR incorporated patch antenna. A continuous SR is placed underneath the patch to obtain an MNG response. The number of turns and the SR dimensions is adjusted so that the inclusion resonates at 0.4 GHz. The MNG region is found to be at 0.47 GHz.

The magnetic field distributions are maximum at the MNG and the air interface and are minimum at the patch's center. Hence, perfect TM_{11} mode is excited on the

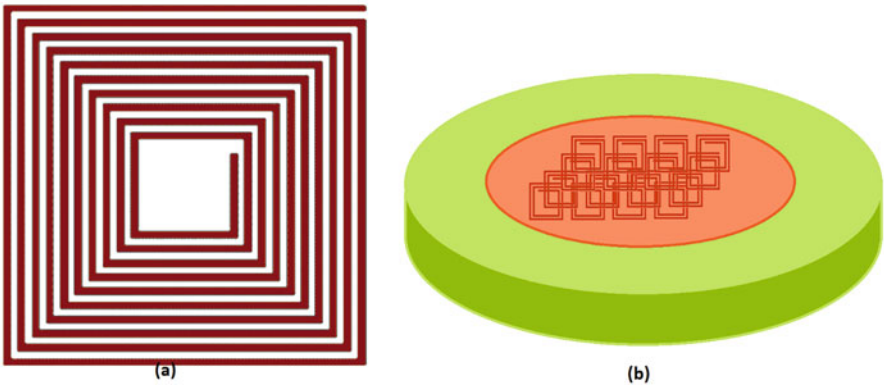

Fig. 24 Development of MNG loaded patch antennas: (**a**) geometry of SR and (**b**) SR ring loaded patch antenna. (With permission from Bilotti et al. 2008)

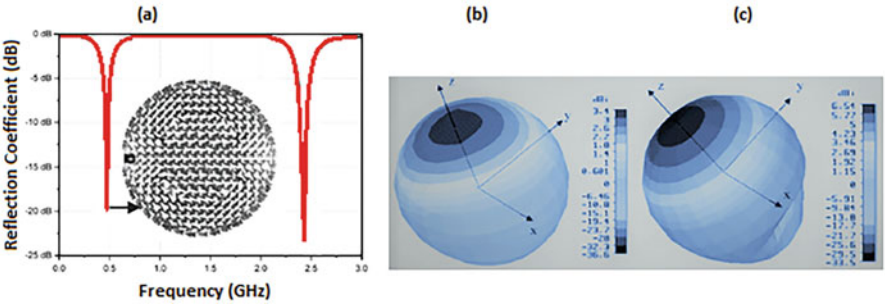

Fig. 25 Characterization of the MNG loaded circular patch antenna: (**a**) reflection coefficient, (**b**) radiation pattern at 0.47 GHz, and (**c**) radiation pattern at 2.44 GHz. (With permission from Bilotti et al. 2008)

patch for the fundamental resonance and is evident from the current patterns shown in Fig. 25a. The MNG incorporated design shows two well-matched resonances at 0.47 GHz and 2.44 GHz, as depicted in Fig. 25. The gain patterns in Fig. 25b, c are comparable with the theoretical models. The fundamental mode's radiation efficiency is 67%, and for the higher mode, it is 92%. The practical realization of such sub-wavelength antenna designs and technological advancements in the fabrication process will surely pave ways to develop deep subwavelength and efficient patch antennas in the future.

SRR Metasurface Based Evanescent Amplification and Radiation Enhancement of an ESA

The metasurface is the two-dimensional compact counterpart of metamaterials. Their characterization is purely based on the induced electric and magnetic polarizabilities

rather than effective material parameters. Since they are planar, a simple two-dimensional photolithographic technique is only required for their fabrication. An essential application of metamaterials is near-field imaging. J.B. Pendry showed that a DNG metamaterial slab with refractive index $n = -1$ could amplify the evanescent spectrum emitted from a source (Pendry 2000). An electromagnetic source emits both propagating and evanescent spectra. In a conventional lens, shown in Fig. 26a, the propagating spectrum (blue line) reaches the far-field, and the evanescent spectrum (red line) undergoes exponential decay within the near-field of the object (Zhang and Liu 2008). Since free space acts as a low-pass filter, the evanescent high spatial spectrum containing the objects' fine details is prohibited. The specialty of this lens is that it amplifies the evanescent spectrum emitted by the electromagnetic source, and it can be said as a superlens.

The practical implementation of a superlens is a challenging task. J.B. Pendry also proved that SNG metamaterials could be employed for evanescent amplification. The description of the problem is shown in Fig. 27a. A silver slab of thickness 40 nm is

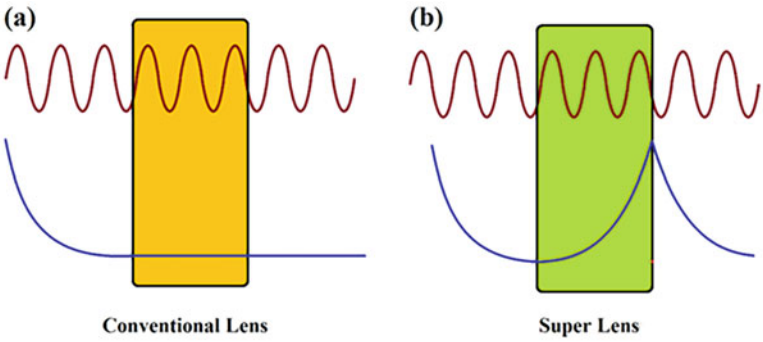

Fig. 26 Operation of lenses: (**a**) conventional lens and (**b**) superlens. (With permission from Zhang and Liu 2008)

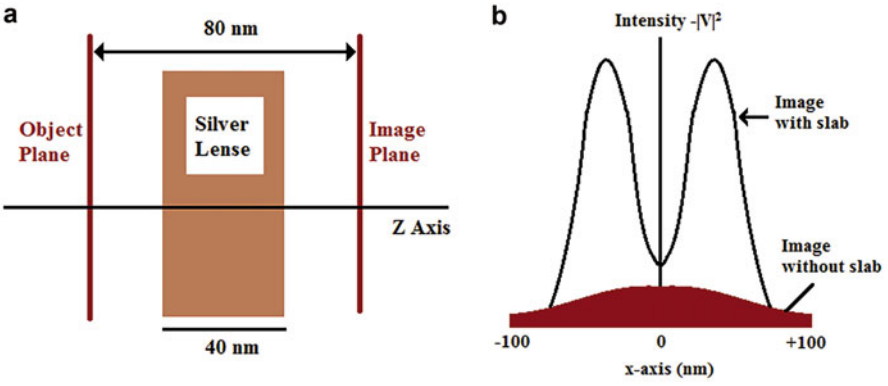

Fig. 27 Near-field focusing by a silver slab: (**a**) geometry of the problem and (**b**) reconstruction of the image with the silver lens. (With permission from Pendry 2000)

used as the plasmonic slab, and it can be represented using the Drude-dispersion model, and the dielectric function can be written as $\varepsilon = \varepsilon_0\left(1 - \frac{\omega_p^2}{\omega^2}\right)$ and below the plasma frequency ω_p; the silver slab shows a negative dielectric constant. The $\varepsilon = -1$ is the essential condition to excite surface plasmon. This condition is applied here for the amplification of evanescent waves. Figure 27b shows the amplification scheme using the plasmonic silver slab meeting the condition $\varepsilon = -1$. In the absence of the silver slab, the image at a distance $z = 2d = 80$ nm is blurred because free space cannot transport high spatial components of the object. When the silver slab is present, considerable focusing is achieved due to evanescent amplification.

However, in the microwave range, we cannot excite surface plasmon modes in a silver slab. At these frequencies, the slab shows pure metallic nature. So, J.B. Pendry proposed that in the microwave range, it is better to use magnetic plasma having $\mu = -1$ in the form of SRR arrays for perfect imaging. This method could also be used for enhancing radiation from electrically small radiators. A reactive near-field characterizes an electrically small dipole source, and the far-field radiated power is significantly low. A magnetic metasurface could be used for amplifying the evanescent spectrum radiated by an ESA. For that, an electrically small radiator in the form of a chip inductor loaded coplanar antenna is used (Deepak et al. 2014). The specialty of this antenna is that it is a well-matched monopole type radiator with $ka = 0.2924$. Figure 28 illustrates the geometry and characterization of this ESA (Sarin et al. 2019a, b). The antenna is formed by soldering a coil craft inductor having inductance $L = 23$ nH to the center conductor of a coplanar waveguide feed. The feed dimensions are $P_g = 0.8$ mm, $P_w = 1$ mm, $P_h = 0.8$ mm, $L_g = 4$ mm, and $W_g = 3$ mm, and is printed on a low-cost epoxy substrate with dielectric constant $\varepsilon_r = 4.4$ and height 1.6 mm. The resonance of the antenna is at 2.38 GHz, and the reflection coefficient value is -12 dB as shown in Fig. 28b. The antenna shows monopole type resonant behavior and radiation behavior, as depicted in Fig. 28c, but the radiated gain is significantly low in the order of -9.74 dBi, and the radiation efficiency is only 13%. The reason behind low radiation efficiency is studied by taking the spatial Fourier transform of the near-field electric field distributions. The computed radiated spectrum of the antenna at resonance is shown in Fig. 28d. Here, the X-axis represents the wavenumber normalized to free space, and Y-axis represents the normalized amplitude. The spatial spectrum with wavenumber $|K_x/K_0| < 1$ is designated as the propagating spectrum resulting in far-field radiation. The spatial spectrum with $|K_x/K_0| > 1$ is called the evanescent spectrum, and the information contained in this spectrum decays exponentially within the near-field. It is clear from the figure that the evanescent components dominate the antennas near-field. The resulting far-field radiated spectrum is weak in magnitude resulting in low radiation efficiency.

This evanescent near-field could be amplified using an SRR array working under the $H_{\|}$ excitation scenario. For that, we need to select the dimensions of the SRR. Its resonant frequency is just below the antenna's resonance so that the negative magnetic polarizability region is allowed to interact with the evanescent spectrum emitted by the source antenna. The dimensions of the SRR is selected such that its resonance is at 2.2 GHz. Hence the negative polarizability region of the SRR coalesces with the

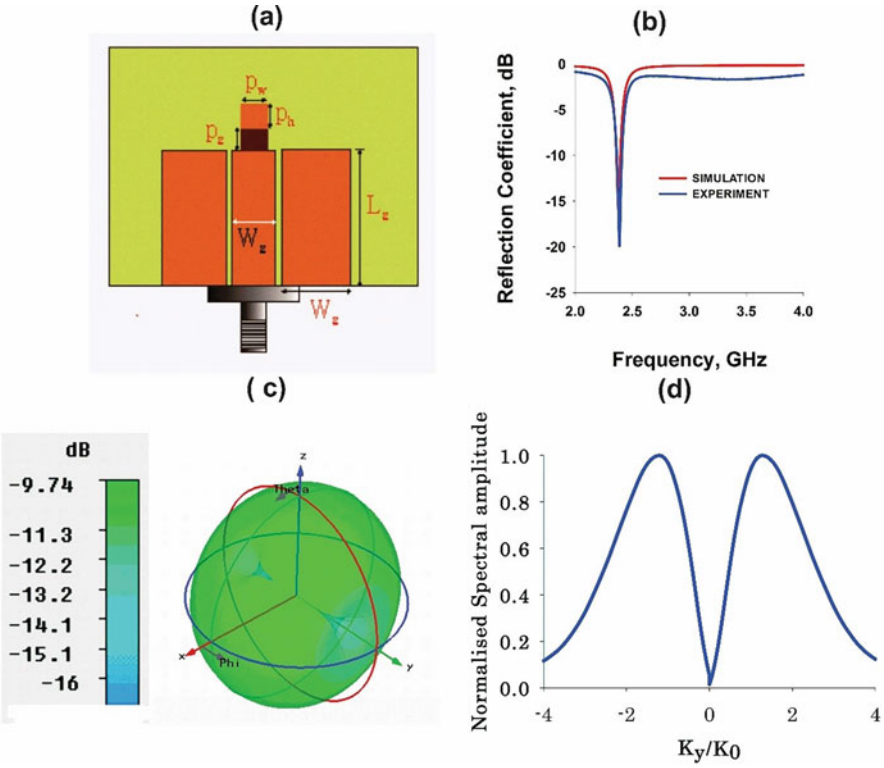

Fig. 28 Characterization of the ESA used for evanescent amplification: (**a**) geometry of the antenna, (**b**) reflection characteristics, (**c**) 3D radiated gain pattern, and (**d**) normalized radiated spectrum. (With permission from Sarin et al. 2019a, b)

antenna resonance. A 3 × 8 SRR array, having size $L_h = 60$ mm and $L_v = 160$ mm, are placed on both sides of the ESA with a small vertical offset. Figure 29 shows the geometry of the designed configuration. The offset parameter $L_d = 2$ mm is selected to achieve maximum radiated power from the antenna.

Since the resonance of the antenna lies within the negative polarizability spectrum of the SRR array, the magnetic resonance of the SRR is excited. The bandgap does not support the propagation of plane waves. Hence, the only possibility of waves supported by the array is the evanescent wave. Figure 30a shows the computed magnetic field distributions taken along a line drawn along the Y-axis for single-, double-, and three-column designs of the antennas. The magnitude of magnetic field intensity shows exponential growth from the ESA location to the SRR location. The amplitude then decays and again grows up to the next SRR location. The peaks are associated with the presence of an SRR. To detect evanescent waves, the authors used near-field probing in full-wave simulations (Sarin et al. 2020). The probe detects the E_y component of electric field and H_x component of the magnetic field in the selected point. It indicates that the electric and magnetic fields are temporally

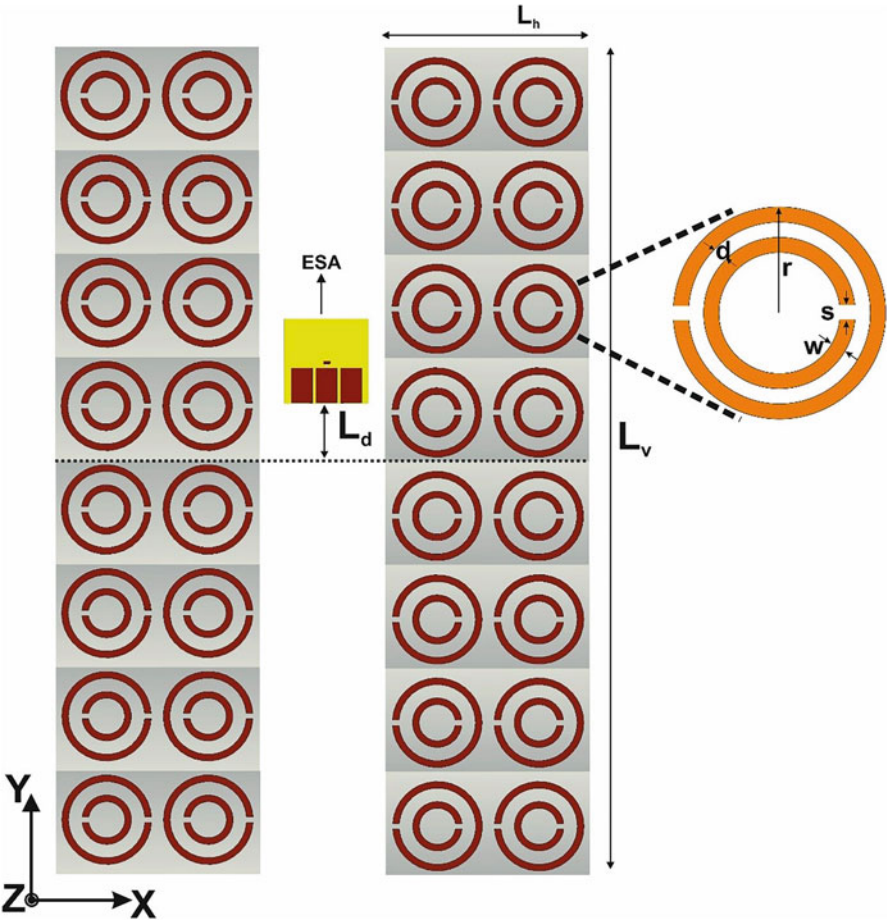

Fig. 29 Configuration of SRR array loaded ESA. (With permission from Sarin et al. 2019a)

shifted in phase by 90°. It shows the presence of pure reactive power within the near-field of the antenna. A comparison of the antenna's radiated spectra with and without the SRR layer is also studied, and Fig. 30b illustrates the radiated spectra of these antennas. The loaded antenna's radiated spectrum is narrower than that of the unloaded case. However, the propagating spectrum's amplitude is enhanced in the loaded design. It means that the SRR layer is responsible for evanescent amplification and converts the high spatial reactive spectrum into radiation.

Figure 31 illustrates the comparison between the radiation patterns of the loaded and unloaded designs. The single-column, two-column, and three-column designs significantly improve radiation from the ESA. The optimized three-column case's gain is enhanced to 2.5 dBi. Nearly tenfold enhancement in radiated power is achieved compared to the unloaded ESA.

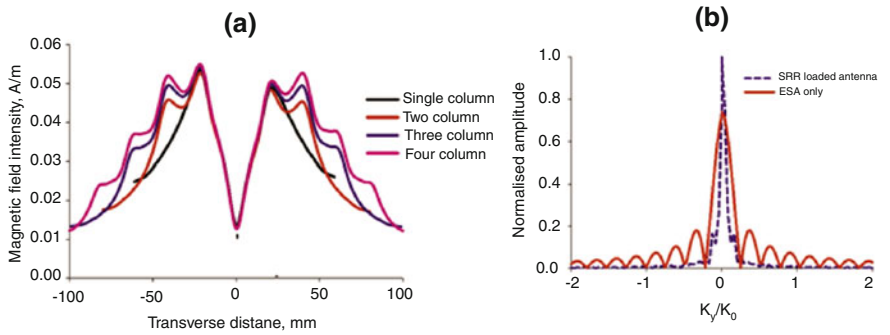

Fig. 30 Field characterization of the antennas: (**a**)magnetic field distributions for the loaded antennas and (**b**) radiated spectra of the loaded and unloaded antennas. (With permission from Sarin et al. 2019a)

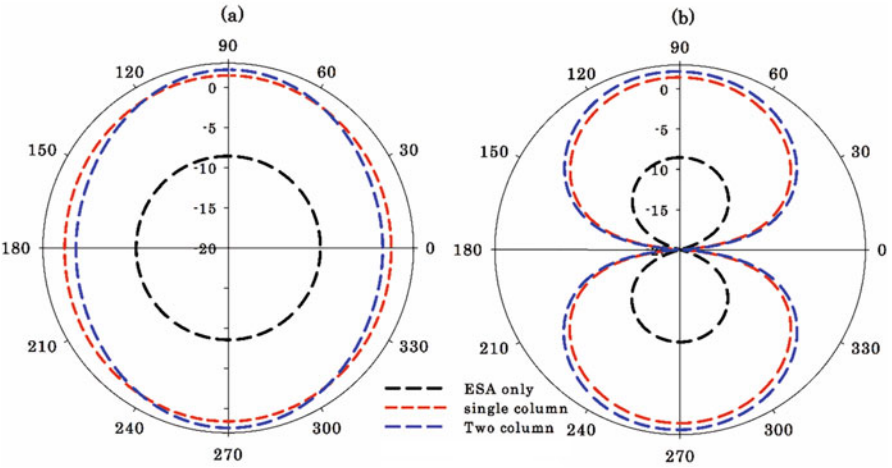

Fig. 31 Comparison of radiation patterns of the loaded and unloaded designs. (With permission from Sarin et al. 2019a)

SRR Coupled Enhanced Transmission for Front to Back Ratio Improvement

The theory of light transmission through a sub-wavelength hole in a metallic hole is treated as a fundamental electromagnetic problem. It is a direct consequence of the wave nature of light. Hans Bethe proposed the first successful theory behind this problem (Bethe 1944). The subwavelength hole is equivalent to a magnetic dipole $M = -\frac{2}{3\pi}a^3 H_0$ and an electric dipole $P = \frac{1}{3\pi}a^3 E_0$ where E_0 and H_0 are the electric and magnetic fields, respectively, at the hole's position when the hole is absent and "a" is its radius. The magnetic dipole is parallel to the hole, and the electric dipole is normal to the hole.

Many techniques are used to enhance electromagnetic transmission through a sub-wavelength hole beyond the diffraction limit. Among them, an important class belongs to the resonant transmission enhancement using split-ring resonators (Bilotti et al. 2009). They showed that covering the hole with ENG materials does not favor enhanced transmission, and hence the solution is the use of permeability resonance. Consider a sub-wavelength hole covered with a single SRR element with two orientations as shown in Fig. 32 (Aydin et al. 2009). The radius of both the SRR is selected to be 3.6 mm and the circular aperture has a radius 4 mm.

The transmission enhancement observed for these two configurations are depicted in Fig. 33. It is observed that a resonant transmission enhancement is observed for the SRR arrangement shown in Fig. 32a. It is interesting to observe that resonant nature is absent for the configuration shown in Fig. 32b.

Enhanced transmission could also be used for microwave antenna applications for reducing the backward radiated power. The typical antenna configuration employing SRR based enhanced transmission is depicted in Fig. 34 (Pushpakaran et al. 2014).

The configuration is a stacked metal slab antenna. A slot line fed metal plate dipole is stacked with another rectangular plate having the same dimension. The microstrip line to slot line transition provides the microwave signal to the antenna

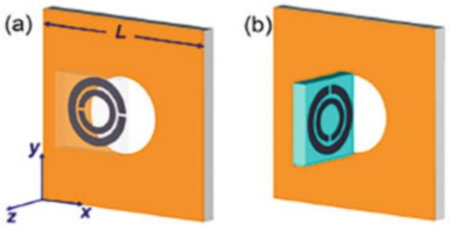

Fig. 32 Normalized power transmission enhancement corresponding to the permeability resonance of MNG metamaterial. (With permission from Aydin et al. 2009)

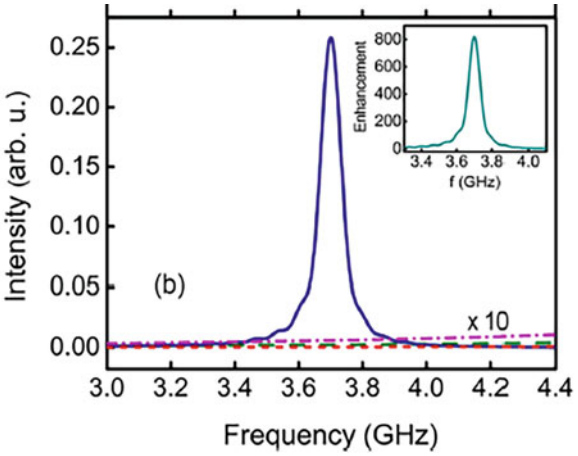

Fig. 33 Transmission characteristics of the prototypes. (With permission from Aydin et al. 2009)

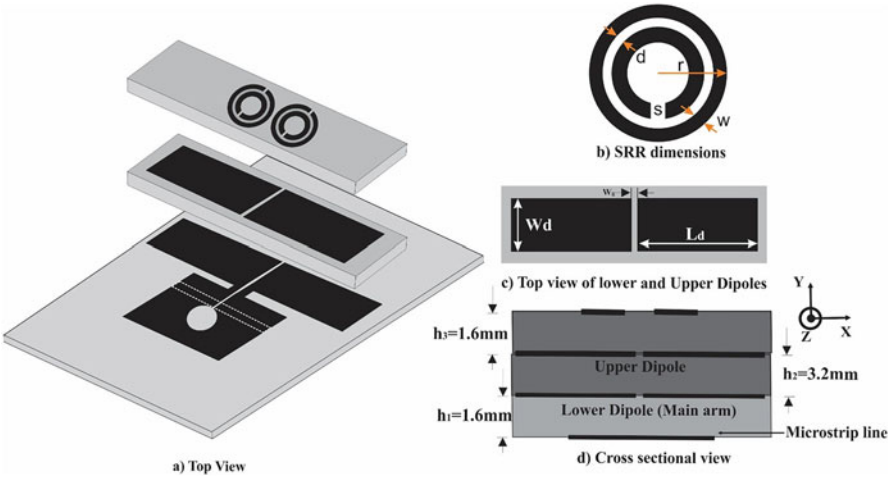

Fig. 34 Enhanced transmission phenomenon for antennas using metamaterials: (**a**) geometry of the SRR incorporated antenna and (**b**) geometrical specifications. (With permission from Pushpakaran et al. 2014)

using a network analyzer's coaxial cable. This antenna could be termed as a magneto-dipole. It uses the coupling between the electric and magnetic dipole moments to achieve broadside radiation coverage without using conventional ground plane backing (Pushpakaran et al. 2013). The resonant frequency of this stacked slab antenna is at 3.08 GHz. Enhanced transmission is achieved by stacking two SRR elements over these stacked dipole plates, as shown in Fig. 34a. Figure 34b illustrates the geometrical specifications of the stacked dipole and the SRR. The SRR dimensions are selected to be $r = 5.2$ mm, $s = 1$ mm, $w = 1$ mm, and $d = 1$ mm, such that the magnetic resonance of the SRR occurs at 3 GHz. The dimensions of the antenna are $L_d = 20$ mm, $W_d = 8$ mm, $L_g = W_g = 14$ mm, $W_s = 0.35$ mm, $L_0 = 0$ mm, $D = 4$ mm, $W_m = 3$ mm, $h_1 = 1.6$ mm, $h_2 = 3.2$ mm, and $h_3 = 1.6$ mm.

The specialty of this magneto-dipole antenna, i.e., the stacked antenna without SRR, is that this structure uses both electric and magnetic dipole moments to achieve directional radiation characteristics. The electric dipole moment (P_z) is created due to the electric charge separation in the dipole plates, and the magnetic dipole moment (M_y) is created due to the antiparallel currents circulating between the lower and upper stacked plates. These moments are oriented orthogonal to each other. So this problem can be treated as equivalent to the electromagnetic transmission problem through a sub-wavelength hole. So, transmission enhancement is achieved by using additional resonant magnetic dipole creation over the plate. So, the function of the SRR is to enhance magnetic dipole excitation over the antenna aperture, and hence enhanced transmission is thus achieved. Therefore, the antenna couples more power along the forward direction. Hence, a significant reduction in radiation is observed along the unwanted backward direction. Figure 35 compares the radiation performance of the antennas with and without the SRR layer. The

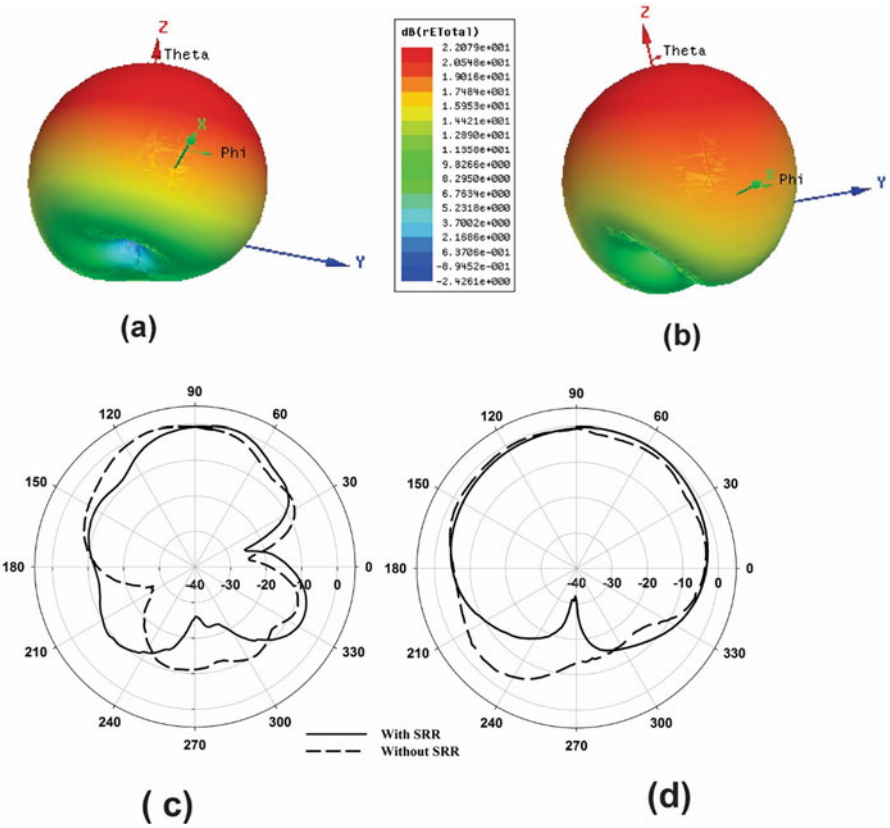

Fig. 35 Radiation comparison between the antennas with and without the SRR: (**a**) 3D radiation pattern of the SRR loaded dipole, (**b**) 3D pattern of the antenna without SRR, (**c**) E-plane radiation patterns of the two antennas, and (**d**) H-plane radiation patterns of the antennas. (With permission from Pushpakaran et al. 2014)

antenna with the SRR reduces the backscattered power significantly in comparison with the unloaded case. The front to back ratio is enhanced from 16 dB to 25 dB for the SRR loaded design.

Artificial Magnetic Conductor Realization for Antenna Applications

Recently, artificial sub-wavelength magnetic inclusions, made in the form of periodic conducting elements printed over a substrate, each of them is connected to the underneath ground plane with vias has attracted significant research interest. These inclusions are collectively termed artificial magnetic conductors (AMC). The AMC nature arises from the formation of small LC tank circuits in the domain. The inductance is offered by the current circulating between the upper patch and the

ground plane. The spacing between the edges of successive patches offers the capacitance. When an electromagnetic plane wave impinges on such a surface, the magnetic resonance will be excited. It creates circulating current loops in the unit cell. This parallel resonance offers infinite resistance at the resonant frequency. Hence, these structures are also termed as high impedance surfaces (HIS). This structure was first proposed by Daniel Sievenpiper and colleagues from the University of California (Sievenpiper et al. 1999). Figure 36 shows the geometry of a typical HIS (Kumar et al. 2020b).

In a conventional plane metal, the impedance offered by the surface is zero due to the anti-phase reflection provided by the plate. An in-phase reflection characterizes the HIS. Hence, it could be used to enhance the radiation performance of conventional antennas. When a dipole antenna is backed by a metallic ground plane, image currents will be induced on the ground plane. The field radiated by the dipole and the image destructively interfere at the far-field (Azad and Ali 2008). Hence both the matching and radiation performance of the dipole source severely deteriorate. This problem can be avoided by replacing the ground plane with an HIS. The advantage is that the in-phase reflection offered by the HIS ground plane, as shown in Fig. 37a, b, offers constructive interference at the far-field, and hence the radiated power at the far-field is improved. The other advantage of using HIS underneath the antenna is that the HIS efficiently suppresses the surface waves that are guided in between the antenna and the ground plane, and a stable radiation pattern is thus obtained.

Richard Ziolkowski proposed an alternate scheme in which the AMC behavior is implemented using capacitive loaded loop resonators. The design is devoid of conventional ground plane backing used in Sievenpiper's HIS (Erentok et al. 2005). The geometry of the AMC unit cell is shown in Fig. 38a. The final arrangement, depicted in Fig. 38b, could be excited in two ways. The reflection phase from the structure when illuminated with a plane wave moving along +Z direction and

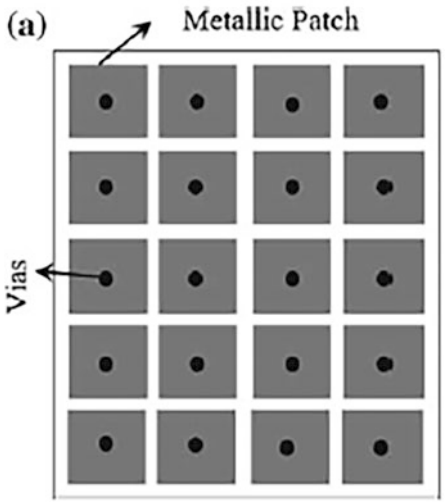

Fig. 36 Geometry of the HIS: (**a**) side view and (**b**) top view antennas. (With permission from Kumar et al. 2020b)

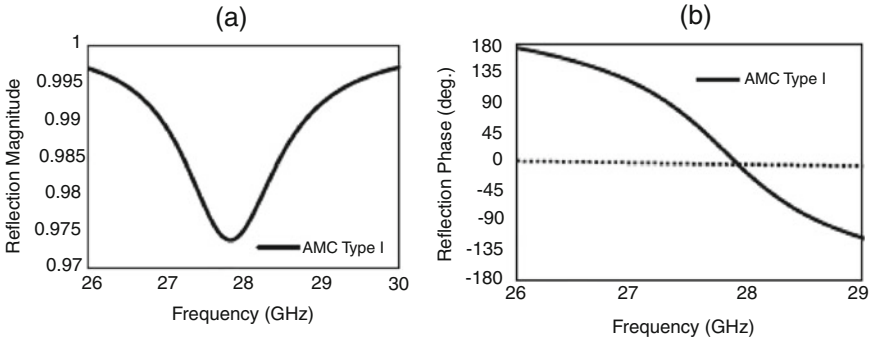

Fig. 37 (**a**) Reflection coefficient of the HIS and (**b**) reflection phase. (With permission from Kumar et al. 2020b)

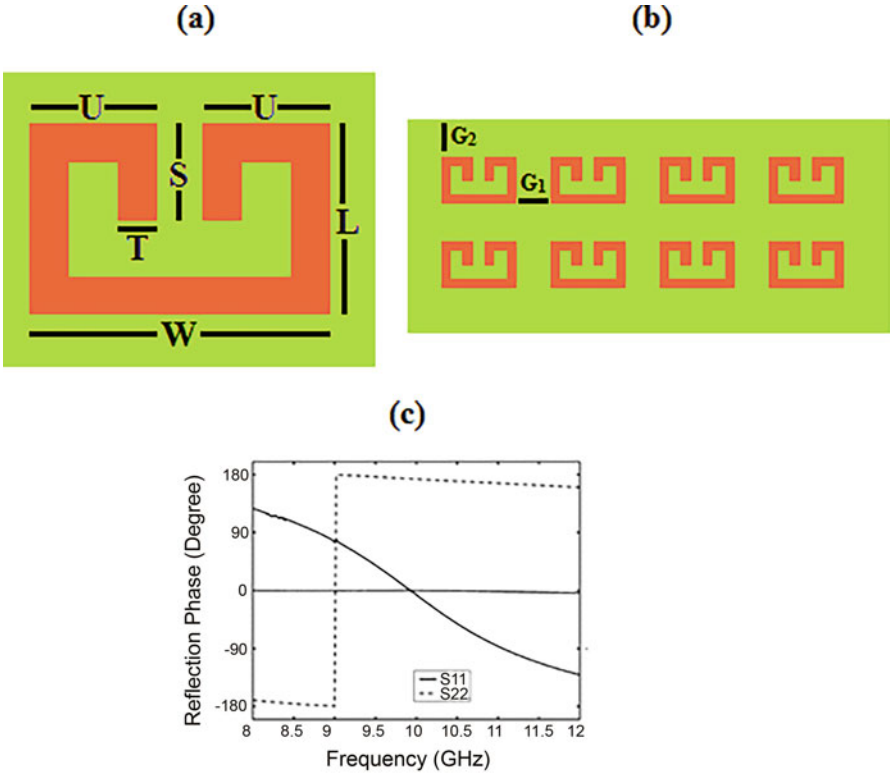

Fig. 38 Characterization of the AMC: (**a**) unit cell dimensions, (**b**) constructed AMC array and, (**c**) phase of reflection coefficients. (With permission from Erentok et al. 2005)

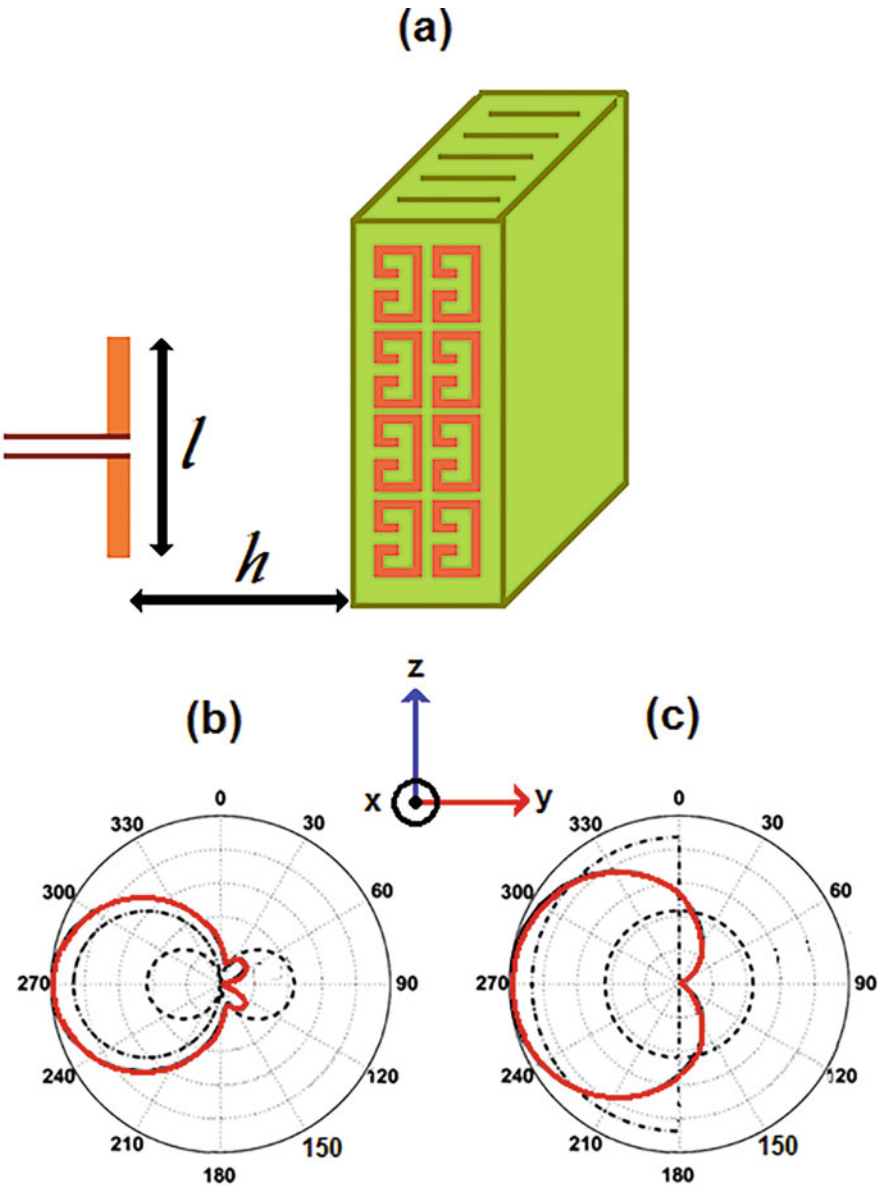

Fig. 39 Characterization of AMC backed dipole: (**a**) geometry of the AMC backed dipole, (**b**) E-plane radiation patterns, and (**c**) H-plane radiation patterns. (With permission from Erentok et al. 2005)

polarization parallel to the slit is depicted in solid black lines in Fig. 39c. In this case, the structure acts like an AMC. The reflection coefficient phase gradually crosses zero around the resonance at 10 GHz. When the incident electromagnetic wave is

directed in the opposite direction, the structure acts like an ordinary perfect electric conductor (PEC). Its reflection phase is represented using the dotted lines in Fig. 38c.

Based on this characterization, an electric dipole working around the AMC's resonance is placed symmetrically in front of the AMC block, as shown in Fig. 39a. The fed dipole's main beam propagates along the +Y direction towards the center of the AMC block. The reflected waves from the AMC block travels along the −Y direction. The E-plane and H-plane radiation patterns of the bare dipole in free space and the AMC backed dipole are shown in Fig. 39b, c, respectively. The deep null at the backside quantifies the front to back ratio improvement offered by the AMC block. More than twofold enhancement in radiated fields is observed at the far-field for the AMC backed dipole.

DNG Metamaterials

DNG metamaterials exhibit simultaneous negative values of constitutive parameters. DNG behavior is achieved by combining the electric plasma behavior of metal inclusions and the magnetic plasma nature of SRR arrays. D.R. Smith from the University of California first demonstrated the existence of a DNG medium by cascading the wire medium and SRR array (Smith et al. 2000). The dispersion diagram of the proposed SRR-wire pair array under the $H_{\|}$ excitation scenario is illustrated in Fig. 40a. Under $H_{\|}$ excitation, a passband is introduced in the bandgap region denoted by the dashed lines. This passband is characterized by negative group velocity. Transmission measurements performed on the fabricated SRR array and the SRR-wire pair array are illustrated in Fig. 40b. The solid black line indicates the transmission coefficient of the SRR array. The SRR array is characterized by a stopband around 5 GHz, as indicated by the transmission dip. The wire array is selected to have a plasma frequency of 12 GHz. Below the plasma frequency, the

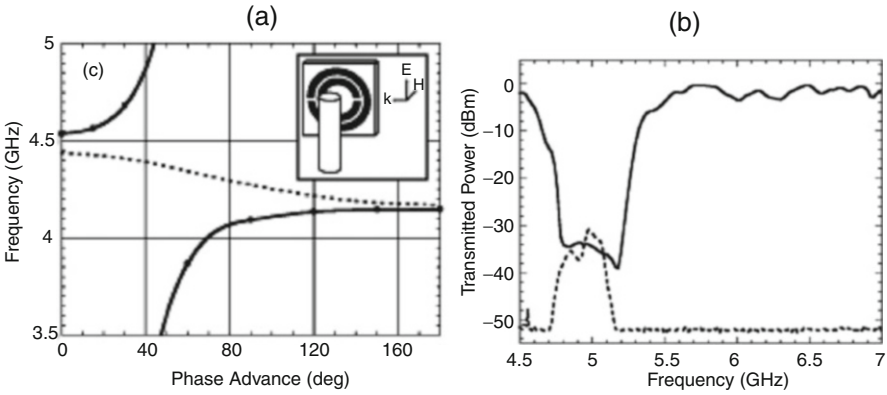

Fig. 40 Characterization of a DNG medium composed of SRR-wire pair arrays: (**a**) dispersion diagram of the unit cell under $H_{\|}$ excitation and (**b**) corresponding transmission coefficient. (With permission from Smith et al. 2000)

transmission is prohibited and shows a transmission coefficient of the order of −52 dBm. Below the plasma frequency, the effective ε of the medium is negative. A frequency bandgap around 5 GHz characterizes the SRR array. Hence combining these two media creates a transmission tunnel band indicated by the dotted lines in Fig. 40b. The effective refractive index of the medium $n_{\text{eff}}(\omega) = \sqrt{\varepsilon_{\text{eff}}(\omega)\mu_{\text{eff}}(\omega)}$ is negative at the tunnel band.

Efficient Radiation from Electrically Small Antennas Using DNG Shell

The complex power at a distance r from an electrically small dipole source in free space is given by (Ziolkowski and Kipple 2005)

$$P_{\text{DPS}} = P_{\text{rad}} + jP_{\text{reac}} = \eta\left(\frac{\pi}{3}\right)\left|\frac{I_0}{\lambda}l\right|\left[1 - j\frac{1}{(kr)^3}\right] \tag{18}$$

where η is the wave impedance, I_0 is the driving current, λ is the operating wavelength, l is the length, and k is the free space wavenumber. Its reactance ratio $P_{\text{reac}}/P_{\text{rad}} > 1$ indicates that the dipole's reactive power is dominated by electric field energy, and this capacitive reactance could be converted into inductive reactance using a DNG shell. For a DNG medium, the substitution of $k \rightarrow -|k|$ in the above equation yields the radiated power when the electrically small dipole source is immersed in a DNG media and is given as

$$P_{\text{DNG}} = \eta\left(\frac{\pi}{3}\right)\left|\frac{I_0}{\lambda}1\right|\left[1 + j\frac{1}{(kr)^3}\right] \tag{19}$$

Here the real and reactive components are both positive; hence the reactive energy is inductive. The reactive power in the DNG case is the complex conjugate of that in a DPS case. The electrically small dipole antenna can be covered with a DNG shell, as shown in Fig. 41, to achieve resonant impedance matching, and the

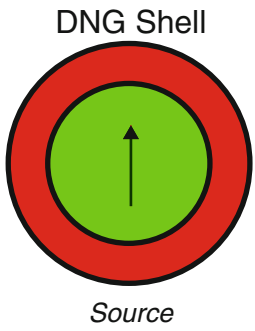

Fig. 41 Application of DNG shell to enhance radiation from an electrically small radiator (With permission from Ziolkowski and Kipple 2005)

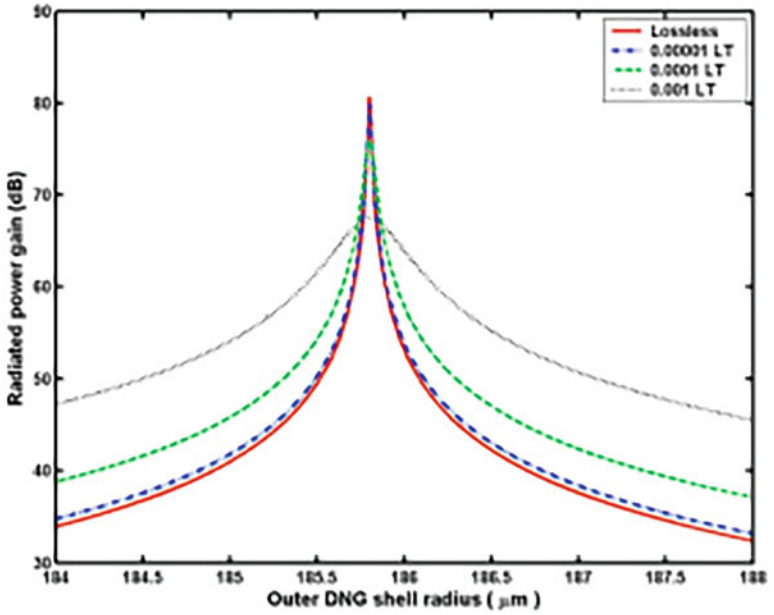

Fig. 42 Radiated power gain of the DNG-dipole configuration for various DNG outer shell thickness. (With permission from Ziolkowski and Kipple 2005)

radiated power is increased significantly (Ziolkowski and Kipple 2005). The DPS medium acts like a high-pass filter, and hence the electrically small antenna in a DPS medium is not an efficient radiator. The DND shell covering the antenna acts like a low-pass filter and provides resonance, reducing the system's total reactance to zero.

The radiated power gain of such a system having the constitutive parameters $(\varepsilon_{\mathrm{DNG}}, \mu_{\mathrm{DNG}}) = (-3\varepsilon_0, -3\mu_0)$ with an inner radius of $a = 100$ μm is shown in Fig. 42. The radiated power gain is the ratio of the radiated power from the dipole source with the cover to that by the bare dipole in free space whose half-length is made equal to the outer radius of the DNG shell. The maximum radiated power gain is observed when the outer radius =185.8 μm. It is observed that a significant increase in loss tangent of the DNG layer decreases the radiated power gain.

Double Positive Materials

Now let us focus on the first quadrant of the Re(ε)-Re(μ) graph. This quadrant represents materials with positive values of the real part of ε and μ. All-natural dielectrics and magnetic materials belong to this category. The propagation in this regime is identified as the right-handed in which the Poynting vector and the energy flow are in the same direction. This section describes the available techniques to enhance the radiation performance of antennas using the DPS materials.

DPS Shell for Radiation Enhancement from an Electrically Small Antenna

Richard Ziolkowski showed the possibility of enhancing the radiated power from an electrically small radiator using SNG (Ziolkowski and Erentok 2006) and DNG metamaterials (Ziolkowski and Kipple 2003). The major problem associated with metamaterials is the complexity associated with their fabrication. B. Ghosh et al. from the Indian Institute of Technology, Kharagpur, proved that DPS materials could also be used to enhance radiation from electrically small radiators (Ghosh et al. 2008). The advantage of using DPS materials is their ease of manufacturing and is nondispersive. The geometry of the computational problem is illustrated in Fig. 43. Here a small magnetic loop of radius "a" is surrounded by a DPS shell with an inner radius "r_1" and outer radius "r_2." The constitutive parameters of the DPS shell are represented by ε_2 and μ_2.

Then the average power radiated by the magnetic loop is

$$P_{\text{rad}}^{\text{DPS}} = \eta\left(\frac{\pi}{12}\right)(ka)^4|I_0|^2 \tag{20}$$

where η is the intrinsic impedance, k is the wavenumber, and I_0 is the driving current. It is evident that the power radiated by the loop decreases as the quadratic power of the radius of the loop. A highly inductive reactance dominates the near-field of the loop, and hence it cannot radiate efficiently, similar to an electrically small electric dipole source. An ENG shell surrounds an electrically small dipole, offers inductive reactance that nullifies the high capacitive reactance. However, for an infinitely small magnetic source, a DPS cover is sufficient to match the antenna with free space.

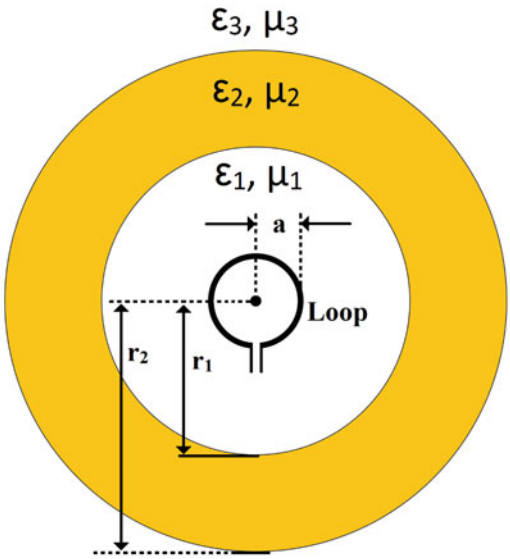

Fig. 43 An electrically small magnetic loop antenna covered with a DPS shell. (With permission from Ghosh et al. 2008)

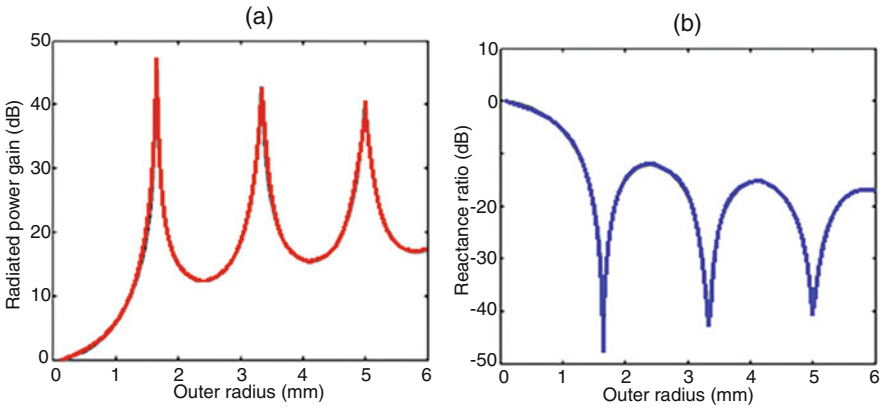

Fig. 44 Characterization of loop antenna covered with DPS shell: (**a**) radiated power gain and (**b**) reactance ratio. (With permission from Ghosh et al. 2008)

For computational studies, the parameters of region 1 are selected as $\varepsilon_1 = 1$ and $\mu_1 = 1$. The DPS shell is characterized to have parameters $\varepsilon_2 = 80$ and $\mu_2 = 1$. Figure 44a shows the system's radiated power gain for different values of the DPS shell's outer radius. Three different power maximum points are observed while varying the outer shell thickness. The highest gain peak is observed at a level of 46.7 dB when the outer radius is made at $r_2 = 1.655$ mm. The magnitude of power gain is observed to be lower for the two other peaks. All the peaks are associated with corresponding capacitive dips indicated in Fig. 44b. These high capacitive dips compensate for the inductive reactance of the loop antenna, thereby creating resonance. The power gain can be improved by using a high dielectric constant material as the shell.

Graded Index Dielectric Loaded Horn Antenna for Side Lobe Suppression

Graded refractive index (GRIN) metamaterial offers gradient spatial permittivity or permeability. GRIN metamaterials are commonly used to tailor the amplitude and phase distribution over the apertures of radiating systems and control the antenna's directional radiation performance. The graded refractive index is achieved by varying the unit cell's geometry and dimensions of the metamaterial. By controlling the aperture field distributions, GRIN metamaterials could be used for various technological applications. One such application is converting the spherical wavefronts emitted by a radiator into a plane wavefront. Another important application is the reduction in side-lobe power radiated by a wideband horn antenna. Horn antennas are widely used in wireless applications because it offers wideband frequency response and low loss. But the major disadvantage is the increased side-lobe level due to nonideal aperture field distributions. GRIN metamaterials could be used to

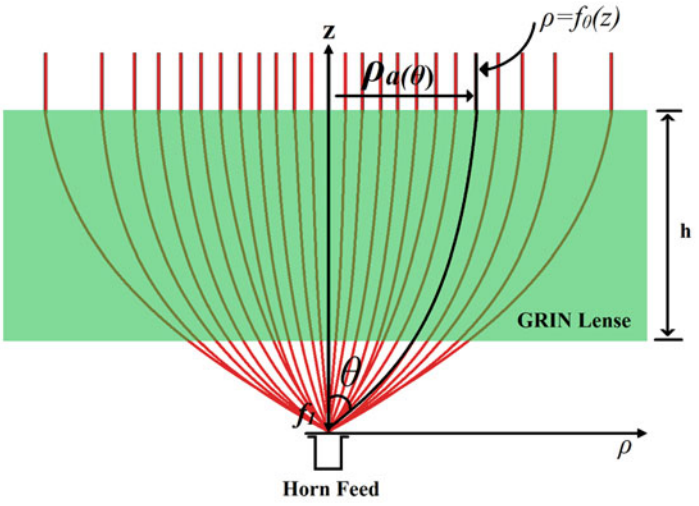

Fig. 45 Schematic of the GRIN lens. (With permission from Qi et al. 2015)

tailor the aperture field distributions of the horn without increasing the overall antenna size.

Mei Qing Qi et al. used the refractive index profile inside the horn's aperture to control the aperture field distributions (Qi et al. 2015). The GRIN lens considered here is shown in Fig. 45. The refractive index $n(\rho)$ is graded along the transverse direction of electromagnetic energy flow within the horn. For achieving a highly directional radiation pattern, the aperture phase distribution should be made uniform. If this condition is achieved, then the spherical waves coming from the horn will be transformed into plane wavefronts.

The GRIN lens is constructed using metallic strips of various length printed on Rogers RO4350B substrate. The GRIN lens's refractive index profile for different metal strip lengths is illustrated in Fig. 46a. For the GRIN layer, the refractive index varies from 1.16 to 1.7 when the strip length varies from 0 to 5.9 mm. To minimize reflections from the GRIN-air interface, a matching layer is also utilized using the same printed metallic strip. The refractive index profile of this impedance matching layer (IML) layer varies from 1.11 to 1.29 and is also shown in Fig. 46a. The front view of the GRIN lens incorporated horn antenna is shown in Fig. 46b.

Simulation studies are performed on the horn antennas with and without the GRIN layer. Figure 47a, b represent the electric field distributions with and without the GRIN lens. It is observed that the diffractive spreading present in the air horn in the E-plane is absent in the GRIN filled horn antenna. Radiation is found to be more directive for the GRIN filled horn. The measured radiation patterns at two distinct frequencies are illustrated in Fig. 47c, d. It is noted that the GRIN lens reduces the side lobes in the E-plane of the horn antenna. In the H-plane, no significant variation in radiation pattern is observed.

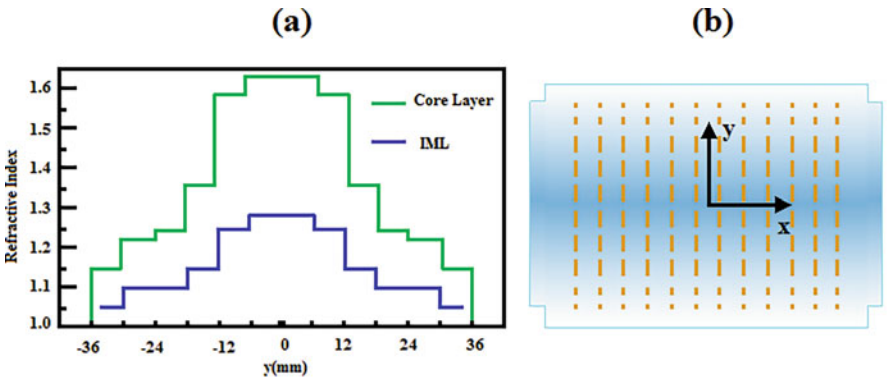

Fig. 46 (**a**) Refractive index variation of GRIN lens and IML layer and (**b**) front view of the GRIN incorporated horn antenna. (With permission from Qi et al. 2015)

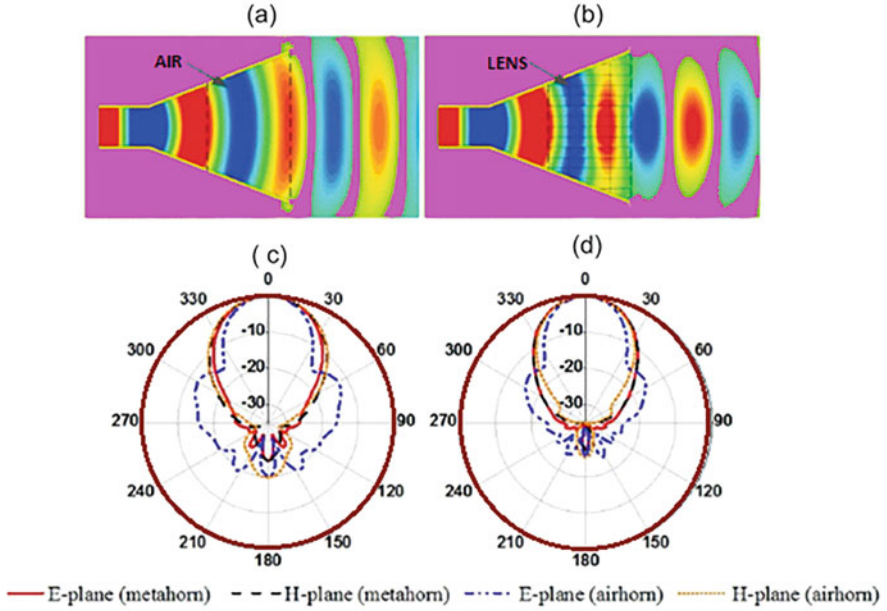

Fig. 47 Characterization of the horn antennas: (**a**) electric field distribution of the air horn, (**b**) electric field distribution of the GRIN horn, radiation patterns at (**c**) 5 GHz and (**d**) 6 GHz. (With permission from Qi et al. 2015)

Near-Zero-Index Metamaterials for Antenna Applications

Now let us focus our attention on the origin of the Re(ε)-Re(μ) graph. This region is identified as the zero-index region constituted by either epsilon-near-zero (ENZ), or mu-near-zero (MNZ), or both ε- and μ-near-zero materials. These three classes exhibit

near-zero refractive index since $n = \sqrt{\varepsilon\mu}$. By inspecting Maxwell's equation in time-harmonic form, $\nabla XE = j\omega\mu H \; and \; \nabla XH = -j\omega\varepsilon H$, substituting $\varepsilon = \mu = 0$ implies that $\nabla XE = 0$ and $\nabla XH = 0$. It means that the electric and magnetic fields are decoupled in a zero-index medium even though they are time-varying. A quasi-static field distribution is observed inside a zero-index material.

It is well known that the electromagnetic response of noble metals obeys the Drude-Lorentz dispersion model. At higher frequencies, especially in the visible and IR ranges, their plasma frequency f_p is characterized by an effective permittivity value approaching zero. This frequency window is located within the THz frequency range for most of the semiconductors and polar dielectrics. The dielectric plasma frequency can be dressed down to the lower microwave ranges using a periodic arrangement of metamaterial inclusions. A simple way to implement ENG metamaterials is to use a standard rectangular waveguide below its cut-off frequency (Rotman 1962). The propagation constant of a traditional rectangular waveguide for the fundamental dominant transverse electric mode TE_{10} is given by $\beta = \sqrt{(2\pi nf/c)^2 - (\pi/w)^2}$, where f is the operating frequency, n is the refractive index of the medium filling the waveguide, w is the width of the waveguide, and c the velocity of light in freespace. In this regard, the effective permittivity of this dominant TE_{10} is represented as $\varepsilon_{eff}/\varepsilon_0 = n^2 - c^2/(4f^2w^2)$ and it means that the waveguide offers structural dispersion. At the cut-off frequency, the effective permittivity offered by the rectangular waveguide is zero, and the propagation constant at this frequency is $\beta = 0$. The wavelength at the cut-off frequency is stretched to infinity, and the phase velocity of propagation reaches infinity.

We simulated a WR284 waveguide in CST Microwave Studio to understand the nature of electromagnetic fields at cut-off. This standard rectangular waveguide's dimensions are 2.84 in × 1.34 in, and its cut-off frequency is around 2.078 GHz. The usable bandwidth of this waveguide is 2.60–3.95 GHz for the dominant mode. Figure 48 shows the computed electric and magnetic field distributions within the waveguide at the cut-off frequency. Since the phase constant $\beta = 0$, a uniform phase of the fields is observed inside the waveguide.

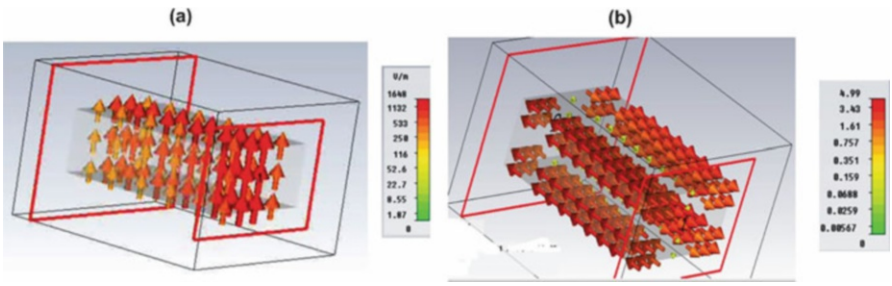

Fig. 48 Field distributions inside the waveguide at cut-off: (**a**) electric field distributions and (**b**) magnetic field distributions

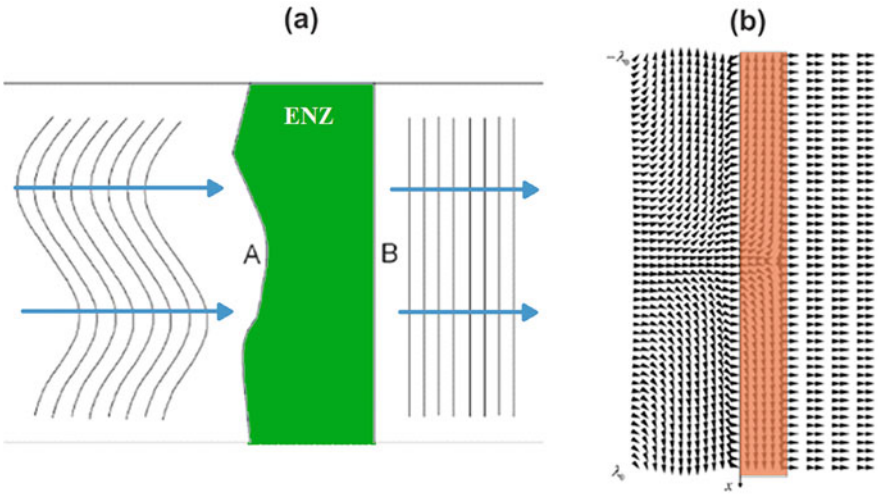

Fig. 49 Characterization of an ENZ metamaterial: (**a**) conversion of a curved wavefront impinges on the entrance side to a planar wavefront and (**b**) corresponding Poynting vector distribution. (With permission from Alù et al. 2007c)

In ENZ materials, the phase variation inside this material is negligibly small compared to that in free space. A typical application of the ENZ slab is illustrated in Fig. 49a (Alù et al. 2007c). Here, an ENZ metamaterial is illuminated with an electromagnetic field exhibiting nonuniform phase distribution on the entrance side designated as side A. Suppose that the ENZ material is perfectly matched with free space such that no reflection occurs from the entrance face. But practically an impedance mismatch is present and a portion of the incident wave will be reflected back. The coupled electromagnetic field into the ENZ material should exhibit uniform phase distribution at the entrance side B. Inside the ENZ slab, the wavelength is stretched to infinity. So, the emitted wave will have uniform phase distribution, and hence a directive beam is formed. The Poynting vector distributions of the same problem is shown in Fig. 49b. In the entrance face, an irregular Poynting vector distribution is observed. The ENZ slab converts this irregular power flow into a uniform one directed perpendicular to the exit face.

This phase conversion property is used for generating highly directional radiation from standard electromagnetic antennas. Stefan Enoch used the near-zero-index behavior of thin metallic mesh wires to control the direction of emission from a source located inside the metamaterial (Enoch et al. 2002). For a metallic mesh structure, effective permittivity is less than one just above its plasma frequency, and its refractive index is near zero. When a point source is located inside such an isotropic near-zero-index metamaterial, then by Snell's law, the ray just above the interface will be refracted in a direction very much closer to its normal. All the refracted rays will be in the same direction just above the interface, thus achieving constant phase distributions. For the practical realization of near-zero-index

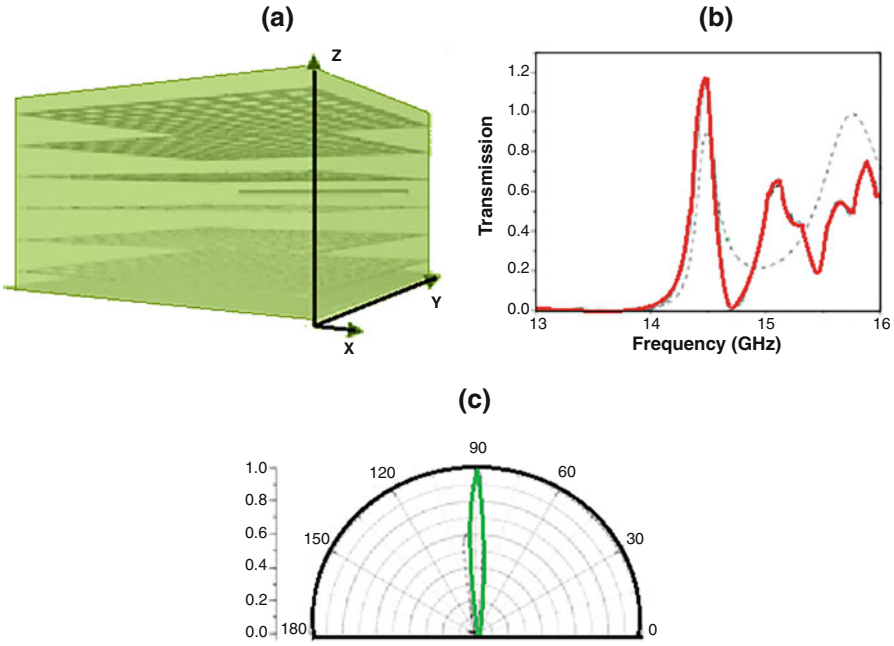

Fig. 50 Directive emission from near-zero-index metamaterial: (**a**) geometry of the source incorporated within the metamaterial, (**b**) transmission coefficient, and (**c**) emission patterns. (With permission from Enoch et al. 2002)

metamaterial, copper grids printed on a planar substrate having a relative permittivity $\varepsilon_r = 1.08$ is used. The metamaterial exhibits a near-zero index at 14.6 GHz. Six such layers are stacked, and a monopole antenna working at near-zero-index frequency is placed inside the material, as shown in Fig. 50a. The transmission coefficient of the structure measured at far-field is shown in Fig. 50b. The extremely low transmission coefficient below 12 GHz indicates the plasma behavior of the metallic mesh array. The radiation patterns of the antenna shown in Fig. 50c indicates that a highly directional emission is attained from the monopole source. The half-power beamwidth is 8.9° in the H-plane and 12.5° in the E-plane.

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Part IV

Metamaterial-FSS for Industrial Applications

Metamaterial-FSS for Energy Harvesting in the Railway Environment

20

Mohammed Kalaagi and Divitha Seetharamdoo

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Abstract

In this chapter, a brief overview on energy harvesting systems in the railway environment and recent advancements is presented. Energy harvesting systems in the railway environment vary from piezoelectric energy harvesters to electromagnetic energy harvesting systems (mostly mechanical vibrations to electric energy) which has been an interesting topic for researchers especially for applications to supply power to wireless sensor networks to insure safety on the railway track. Recently, experimental measurements have been observed for electromagnetic fields in the railway environment at a wideband range of frequency from 10 kHz to 1 GHz where unintentional signals or ambient energy has shown to exist in the

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railway infrastructure. The design of a high efficient and compact energy harvesting system based on metamaterials has been considered. It is based on a metamaterial structure designed to operate at low frequencies, which can miniaturized and compatible for railway applications in the microwave region between 300 MHz and 1 GHz. The metamaterial design given is based on a spiral structure and has shown high efficiency in terms of absorption at the given operating frequencies up to 98% with a subwavelength structure. Unlike metamaterial absorbers, the energy is dissipated through a via where 84% of the energy was delivered through a resistive load of and the power efficiency is calculated at that point, in which a full wave rectifying system can be given with a miniaturized, high efficiency and compatible for the given application.

Keywords

Metasurfaces · Metamaterial · Absorption · Energy harvesting · High efficiency · Spiral structure · Railway application

Introduction

Nowadays, energy harvesting in the railway environment has become an interesting topic for researchers (Zhang et al. (2017); Wang et al. (2012); Jiang et al. (2014); Duarte and Ferreira (2016); Wang et al. (2013); Gao et al. (2016); Nelson et al. (2011); Cahill et al. (2014); Ruscelli et al. (2017); Bradai et al. (2018); Park and Kim (2016); Nagode et al. (2010)) with a gain of popularity for railway transportation systems. Several low-power energy harvesting systems have been developed and deployed in the railway environment (Beeby et al. (2007); Zuo and Tang (2013); Yang et al. (2009)) based on the existing sources in the railway infrastructure from mechanical vibrations De Pasquale et al. (2012), disturbances and electromagnetic fields Lin et al. (2018). The interest is supplying low power to feed sensors in wireless sensor networks (Rakshit et al. (2015); Macucci et al. (2016)) deployed for instance for rail-way health monitoring, thus reducing maintenance, inspection costs, and constraints (Tan et al. (2009); Lynch and Loh (2006); Shafiullah and Gyasi-Agyei (2007)).

Several types of energy harvesting technologies have been deployed; they are popular in railway applications due to the fact that they are based on mature technologies and that the sources present have a sufficiently high amplitude to yield the required power for feeding sensors or devices. Some of the harvesters that we can mention can vary from piezoelectric energy harvesting systems Li et al. (2014) to electromagnetic Guo et al. (2017) and solar-based energy harvesting technologies for railways Darshana et al. (2015). Piezoelectric harvesting systems have been developed to harvest mechanical vibrations and pressure induced by the moving trains to the rail and convert it to electrical power (Nelson and Platt (2007); Mallouli and Chouchane (2017); De Pasquale et al. (2012); Nagode and Ahmadian

(2011)). In the field of electromagnetic energy harvesting systems in the railway environment, different systems have been deployed which are mostly based on mechanical vibrations at very low frequencies (Lin et al. (2018); Wang et al. (2013), (2012); Bradai et al. (2018); Park and Kim (2016); Yang et al. (2009)).

However, in electromagnetics a high level of electromagnetic energy is present; it has been experimentally observed recently by Heddebaut et al. (2018) for frequencies as high as the microwave region. Moreover according to the EMC standard EN 50121–3-1 Standard (2017) for EM emission limits in the railway environment, a high level of electromagnetic energy can exist at a wideband of frequency from 10 kHz up to 1 GHz. To propose the design of a rectenna system (Georgiadis et al. (2010); Marian et al. (2011); Sun et al. (2012)) for such application can be a challenge due to its low efficiency and incompatibility at low frequencies.

To date no mature technology has been proposed for such applications and frequency range. We are interested in proposing the design of a novel electromagnetic energy harvesting system based on metamaterials, and investigate their potential in terms of efficiency and miniaturization at low MHz frequencies. Metamaterials Smith et al. (2004) have become a subject of interest for researchers nowadays, especially of their exceptional and unprecedented physical properties that can be applied for different applications, due to the generality of their design techniques and methods Jahani and Jacob (2016); Feng et al. (2018). They are characterized by their sub-wavelength and periodic structures, as well as the ability to control an electromagnetic wave in different manners such as abnormal reflection Yu et al. (2011) Kalaagi and Seetharamdoo (2019), cloaking Hao et al. (2010), enhancing the performance of antennas Rabah et al. (2016), collimating orbital angular momentum vortex waves at microwave frequencies Yi et al. (2019), absorption Bowen et al. (2016); Wu and Shvets (2012), and energy harvesting Almoneef and Ramahi (2015).

Particularly, for metamaterial absorbers, different techniques have been presented to achieve high efficient absorption Khuyen et al. (2015), such as using bending shapes Bui et al. (2014), high resonance metamaterial structures with narrow band absorption Tao et al. (2008), as well as other techniques for multi-band Park et al. (2013) and ultra-wideband absorbing metamaterial designs using multi-layer structures Xiong et al. (2013) and designs with a low quality factor Zuo et al. (2016). On the other hand, metamaterial energy harvesters do not just require full absorption, but also the dissipation of the absorbed energy across a load resistance Almoneef and Ramahi (2015). Therefore, it is interesting to investigate the potential of metamaterial energy harvesters for the range of frequencies between 300 MHz and 1 GHz, especially for the application of railway environment.

In this chapter, we introduce other concept that can be interesting and compatible for our application, especially at low microwave frequencies, to overcome the challenges of low efficiency and miniaturization. Thus, we investigate the potential of metamaterial energy harvesters based on absorption, with designs at operating at a single frequency and multi-band frequencies. The results show high efficiency at the given frequencies especially for power dissipation to the load resistance.

Summary on Energy Harvesting Systems and Sources in the Railway Environment

Overview of Existing Energy Harvesting Systems in the Railway Environment

Among most common energy harvesting systems that have been developed and deployed, one that we can mention is Piezoelectric energy harvesters. These technologies focus on the conversion of ambient sources such as vibrations and pressure Zhang et al. (2017); Gao et al. (2016); Li et al. (2014); Mallouli and Chouchane (2017) into electricity. Nelson and Platt (2007) introduced a piezoelectric system based on crystalline material where mechanical energy in the form of vibration and pressure provided by the train on the rail can be harvested and converted into electrical energy. They concluded that 0.05 mW can be generated by every train passage, which is sufficient for supplying wireless power sensors. Other types of energy harvesting systems for railways have been also practically implemented such as railway solar panels for solar energy harvesting Darshana et al. (2015).

In the field of electromagnetic energy harvesting less systems have been developed for the railway application, where most are based on converting mechanical vibrations into electricity at very low frequencies. Wang et al. (2013) have designed a mechanical motion rectifier to harvest vibrational power from rail-road track turnings done by passing trains. The details of the prototype are shown in Fig. 1 (a and b).

The system was tested for moving trains at 40 mph and for vibration frequencies between 1 Hz and 4 Hz. They concluded that the overall system efficiency ranged between 49% and 71% which depends on the resistor used where it was capable of harvesting 10 to 50 Watts.

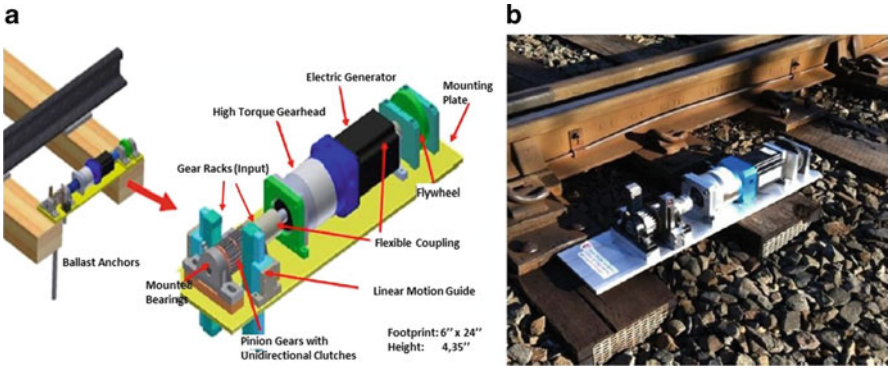

Fig. 1 (**a**) 3D design description of the mechanical motion rectifier; (**b**) Prototype design on field test. Wang et al. (2013)

Existing Electromagnetic Sources up to the Microwave Region in the Railway Environment

Recently, Heddebaut et al. (2018) have performed experimental investigations and analysis of the railway catenary line radiation and unintentional emitted signals found in the railway environment.

Moreover, it can also be inferred from the electromagnetic compatibility (EMC) standard EN 50121–3-1 Standard (2017) limiting the level of EM emissions in the railway environment that a high level of electromagnetic energy can exist at wide band range of frequencies from 10 kHz to 1 GHz. Figure 2 shows the EM emission energy levels limits in the railway environment for the electric field region and magnetic field region with respect to frequency from 10 kHz to 1 GHz which has been extracted from the standard EN 50121–3-1 Standard (2017). It should be noted that the gap in field levels in the frequency bands are due to the fact that the field is measured using detector with different integration times in each band. The integration time of the detectors is defined based on those present in victim EM devices available at those frequencies. For instance, for higher frequency range, devices considered are telecommunication systems, and the integration time is higher resulting in a higher E-field level.

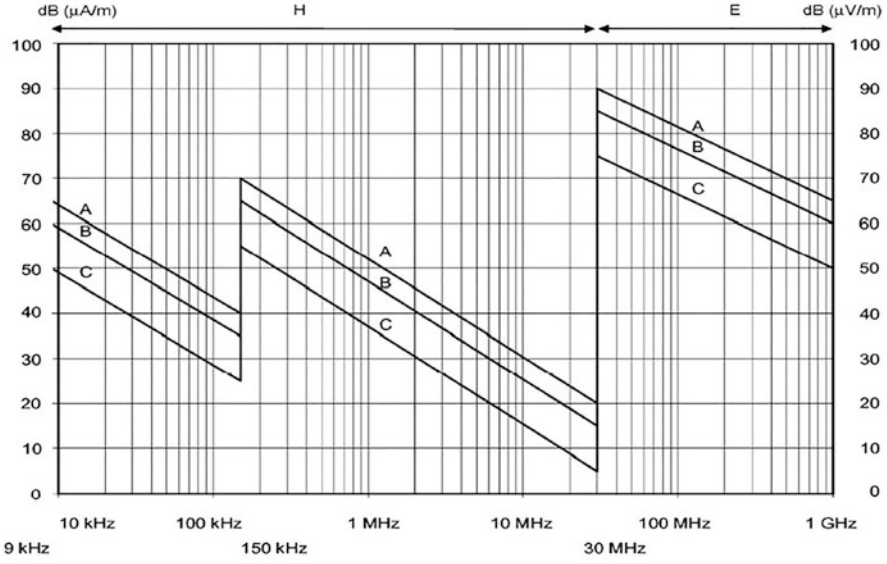

Fig. 2 Measurements of the field radiation in the railway environment, where the curves A, B and C show the radiation emission limits. For more details refer to the standard EN 50121–3-1. Standard (2017)

Proposal of High Efficient Metamaterial Energy Harvesters at Low MHz Frequencies

In this part, we propose an EM energy harvester based on metamaterials at low MHz frequencies, especially for that which is compatible with our application and measurements for the railway application and according to the standard in the microwave region shown in Fig. 2. The aim is to overcome the challenges of low efficiency and bulkiness that existing EM energy harvesting systems can have. This work has been presented at the 13th European Conference on Antennas and Propagation (EuCAP2019) Kalaagi and Seetharamdoo (2019a) and the International Workshop on Antenna Technology (iWAT2020) Kalaagi and Seetharamdoo (2020).

Challenges

Rectenna systems are well-known existing technologies used for EM energy harvesting and wireless power transfer. A simple rectenna system consists of a dipole antenna with an RF diode connected across the dipole elements. In terms of railway applications, where electromagnetic sources are proved to exist at low kHz and MHz frequency bands, the aim is to design an electromagnetic energy harvesting system which is compact and highly efficient. The implementation of a rectenna energy harvesting system Sun et al. (2012) can be a challenge where antennas can have a large size as well as poor efficiency and hard to implement for such applications. We propose and investigate the potential of metamaterial energy harvesters at low MHz frequencies in terms of efficiency and miniaturization which can be compatible for the railway applications.

Principle of Operation and Concept

Metamaterials Cui et al. (2010) and metasurfaces have become a remarkable subject of interest nowadays for many researchers, and this is specially due to their unprecedented physical properties in terms of wave control Lee et al. (2020); Kalaagi and Seetharamdoo (2019), cloaking Cai et al. (2007), focusing Salama et al. (2020) and high efficiency absorption Jiang et al. (2014); Lee et al. (2016); Amiri et al. (2020).

Particularly, an important feature for metamaterials is perfect absorption Landy et al. (2008), in which a media is designed neither to transmit nor reflect an incoming EM wave. A metamaterial absorber is usually designed in a way composed of a metamaterial inclusion printed on top of a substrate with a metallic ground plane on the bottom of the design to prevent transmission. Since transmission is neglected with the presence of the ground plane, for perfect absorption and zero reflection, the metamaterial inclusion should be designed and optimized where the surface impedance is equal to that of the impedance of air $\eta = 377\ \Omega$.

However, for EM energy harvesting, starting from the design of a perfect metamaterial absorber, the energy should be dissipated and collected for energy

conversion. A concept has been introduced by Almoneef and Ramahi (2015), where a via is inserted connecting the metamaterial inclusion to a load resistance placed at the ground plane, thus dissipating the energy absorbed by the metamaterial in which a rectenna system can be placed at the bottom part of the metamaterial collector.

In our application, for railways, we want to investigate the potential of metamaterial energy harvesters for low MHz frequencies in terms of efficiency and miniaturization. The choosing of the right metamaterial inclusion and its optimization can be a challenge. To be able to have a design operating at low MHz frequencies, this is directly related to the inductance and capacitance of our design as defined in eq. (1).

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{1}$$

where f represents the resonance frequency, L the inductance, and C the capacitance of the metamaterial structure. We can see that with structures with high inductance and capacitance we can achieve a metamaterial design operating at low frequencies.

Design of a Single Band Metamaterial Energy Harvester

A spiral inclusion is designed and optimized in this case as shown in Fig. 3. The spiral metamaterial design is chosen in this case, this is due to its loop structure having a high effective inductance resulting in resonance at low frequencies for the given dimensions which can be suitable for our application.

The spiral has been printed on a Rogers TMM4 grounded substrate of relative permittivity $\varepsilon = 4.4$, loss tangent 0.002, dimensions 6×6 cm ($\lambda/14$), and thickness $t = 10$ mm ($\lambda/100$). To ensure a highly efficient absorption at the resonance

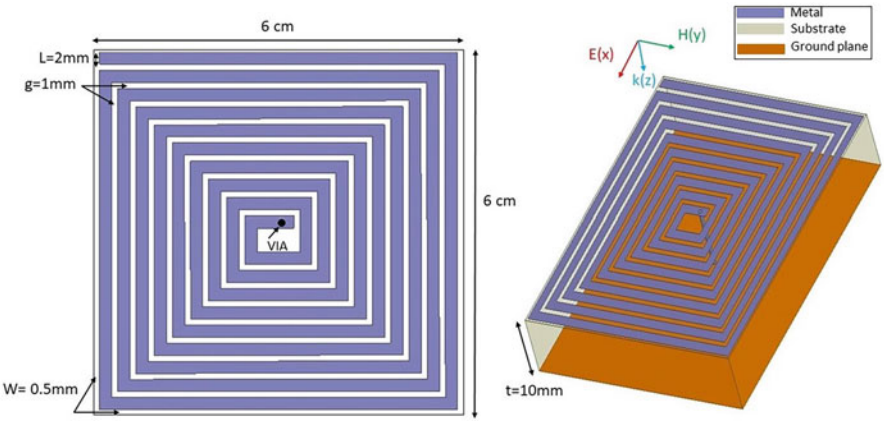

Fig. 3 Spiral structure design with the given dimensions printed on top of a substrate with a via connecting the inclusion to the ground plane

frequency, the surface impedance of the metamaterial structure should be optimized to be equal to that of air $\eta = 377\ \Omega$. The structure has been designed to the following dimensions given in Fig. 3 where $w = 0.5$ mm refers to the gap between the spiral and the substrate, $g = 1$ mm is the gap between the spiral lines and L = 2 mm is the width of the spiral. A via is inserted, in which it connects the spiral structure to the ground plane. The resonant frequency and the surface impedance have been calculated using Ansys HFSS as shown in Figs. 4 and 5 where the structure has shown to resonate at 350 MHz having a surface impedance which has been optimized to a value approximately equal to 377 Ω.

Figure 4 shows a resonance of −38 dB at 350 MHz, this indicated that the energy is perfectly absorbed at the resonant frequency. Figure 5 shows the surface

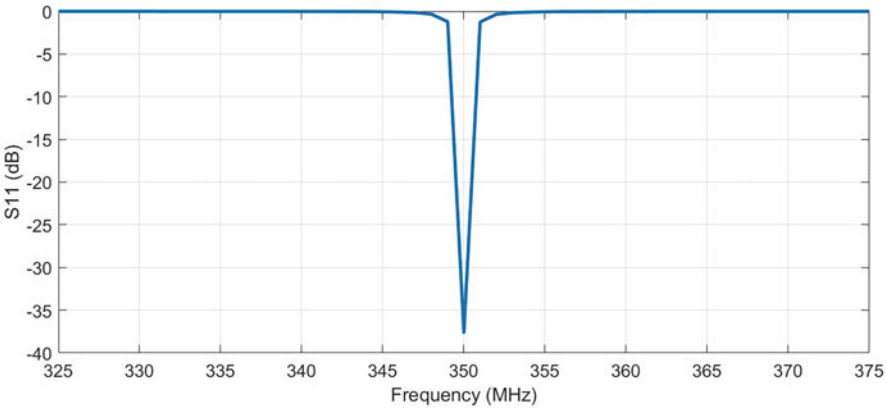

Fig. 4 Simulation results of the reflection coefficient S_{11}(dB) with respect to frequency MHz) using Ansys HFSS

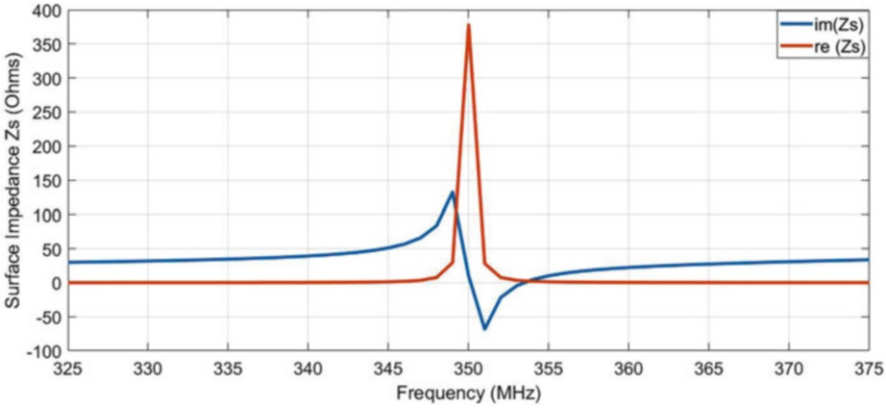

Fig. 5 Real Re(Zs) and Imaginary Im(Zs) parts of the surface impedance Zs of the spiral metamaterial design with respect to frequency (MHz)

impedance of the metamaterial structure which is approximately equal to 377 which is that of the impedance of air.

However, to obtain the amount of energy absorbed by the metamaterial design at the resonant frequency, the absorption rate is calculated according to the following eq. (2):

$$A(w) = 1 - |S_{11}|^2 - |S_{21}|^2 \tag{2}$$

In Eq. (2) S_{11} presents the reflection coefficient of the metamaterial slab, whereas S_{21} is the transmission coefficient. In this case, the structure is printed on top of a grounded substrate thus S_{21} is said to be zero. The absorption rate of the design has been calculated at the resonant frequency as shown in Fig. 6.

The results in Fig. 6 show a level of absorption up to 98%, this indicates that our design is a perfect absorber at the resonant frequency 350 MHz. The second step is to have a design which can be used for energy harvesting, thus we connect a load resistor (lumped port) to the end side of the via, this is in order to calculate the amount of dissipated energy at the load with respect to the energy absorbed by the metamaterial. To ensure total power delivery to the load resistors, the impedance of the load should be matched to that of the spiral structure design. An impedance study has been done from the load side in order to calculate the appropriate value of the load resistor as shown in Fig. 7.

Figure 7 shows the impedance from the load side of the design. The results show a real value to match the metamaterial structure of 8.5 kΩ at a frequency of 382 MHz. The shift in frequency is expected due to the presence of the load resistance. Moreover, the resonant frequency from the load side has been calculated for a 4 × 4 array of metamaterial unit cells as shown in Fig. 8.

The results in Fig. 8 indicate that the design with the load resistance is well matched at the resonant frequency 382 MHz. However, it is important to calculate the power efficiency of the energy dissipated to the load from the absorbed energy by the metamaterial design. The amount of power dissipated to the load from the absorbed energy is shown in Fig. 9.

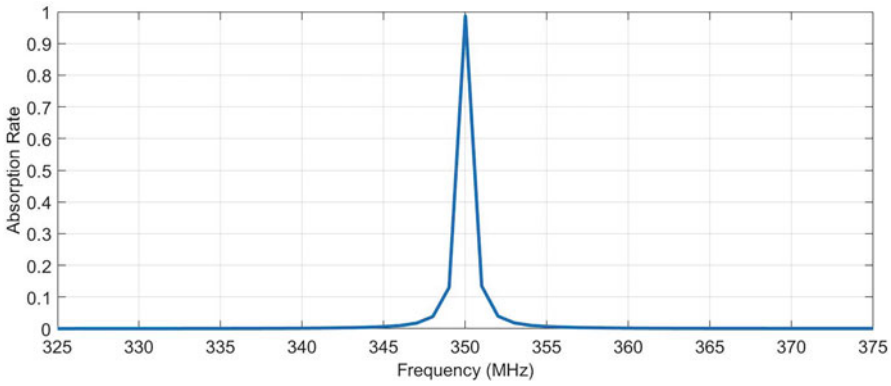

Fig. 6 Absorption rate of the spiral design with respect to frequency (MHz)

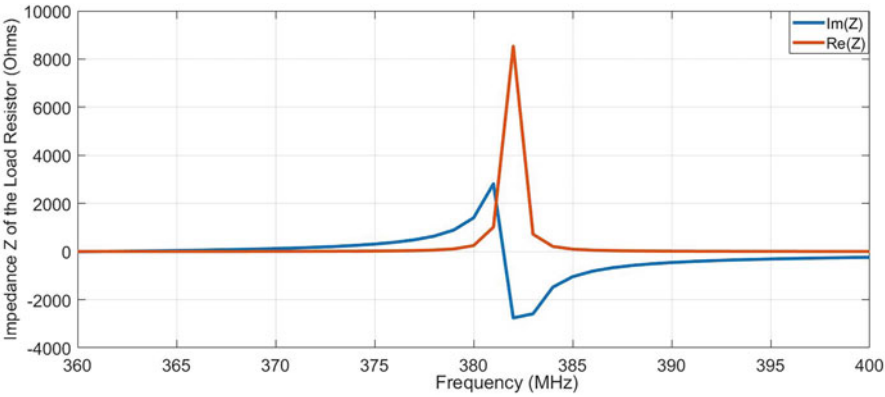

Fig. 7 Impedance Z of the load resistance real and imaginary parts in Ohms with respect to frequency(MHz)

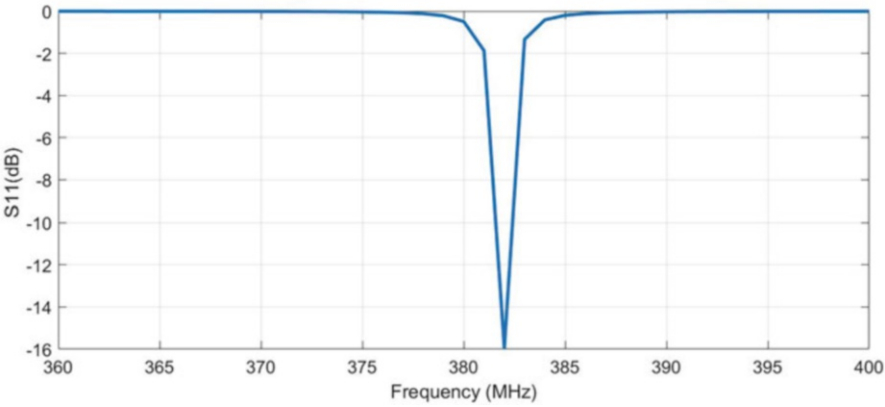

Fig. 8 Reflection coefficient S_{11} (dB) from the load side of the design structure

The results in Fig. 9 show a power efficiency of 84% dissipated to the load resistance. This indicates that most of the energy absorbed by the metamaterial is delivered to the load.

In this part, we have given a first proposal and design of a high efficient energy harvesting system based on metamaterials at a given frequency where 84% of the absorbed energy was dissipated across an 8.5 kΩ load resistor. The design has shown to be high efficient and of low profile of unit cell dimensions $\lambda/14 \times \lambda/14$ and thickness $\lambda/100$ compared to other existing energy harvesting systems Georgiadis et al. (2010); Almoneef and Ramahi (2015); Marian et al. (2011) which can be an important feature specially at low frequencies. The frequency at 350 MHz has been chosen as an example to prove the idea that metamaterial energy harvesters can be highly efficient and of subwavelength structure at low frequencies which can be a solution for applications such as railways.

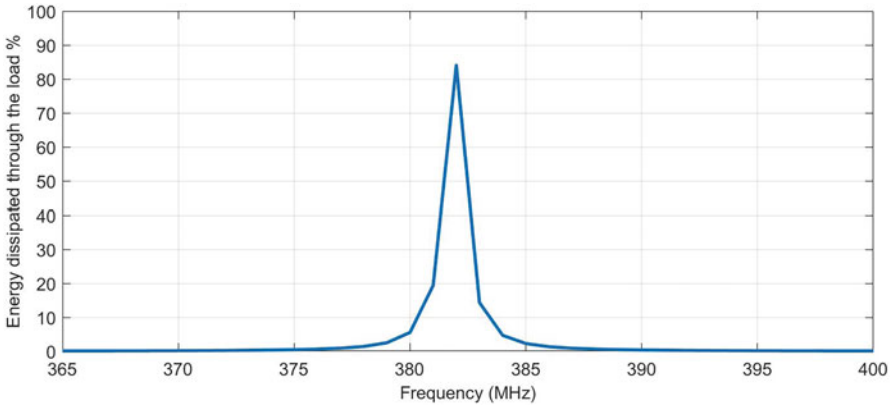

Fig. 9 Power efficiency of the energy dissipated through the given load resistor with respect to frequency (MHz)

Design of a Multi-Band Metamaterial Energy Harvester

In this part, we introduce another type of metamaterial energy harvesting design based on absorption for multi-bandof frequencies. We follow the same concept we introduced before for the energy harvester at a single band. Researchers have introduced different design types and concepts for multi band absorbers, where different methods have been introduced from multi-layer Xiong et al. (2013), to bending shapes Bui et al. (2014) and low quality factor designs Zuo et al. (2016).

However, in terms of energy harvesting, we introduce a coil spiral unit cell design given in Fig. 10a. This type has been chosen specially due to the metamaterial structure's high inductance and ability to resonate at low MHz frequencies, which can be suitable for our application for railways. Moreover, an air gap is given between the metamaterial unit cell and the ground plane, to ensure a low quality factor design structure.

The metamaterial design is printed on top of an FR4 substrate of permittivity $\varepsilon = 4.4$, dimensions 80×80 mm, and thickness 1.6 mm. An air layer gap separates the metamaterial structure from a ground plane with thickness 60 mm. This is to ensure a low quality factor of the design and multi-band resonances. On the other hand, for the dissipation of the absorbed energy, a VIA is connected from the top layer to the ground plane with a load resistance of 282 Ω at the end part of the VIA as shown in Fig. 10b. The coil spiral structure shown in Fig. 10a consists of 12 turns, width $w = 2$ mm, the gap between the turns $g = 0.75$ mm, and the inner radius is equal to 1.5 mm. The unit cell design is shown in Fig. 11.

After introducing the concept and optimization of our design structure, the resonant frequency S_{11} (dB) of the given design is calculated using Ansys HFSS. The results are given in Fig. 12.

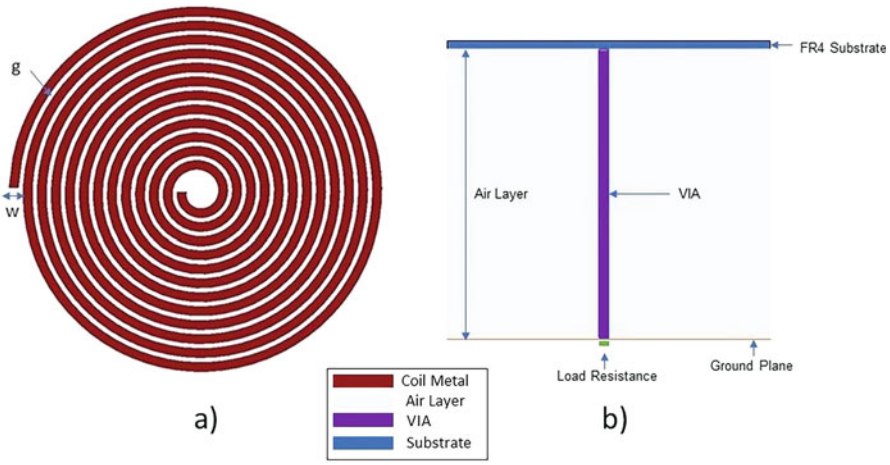

Fig. 10 (**a**) Coil spiral structure of 12 turns, width w, and gap g; (**b**) side view of the proposed design

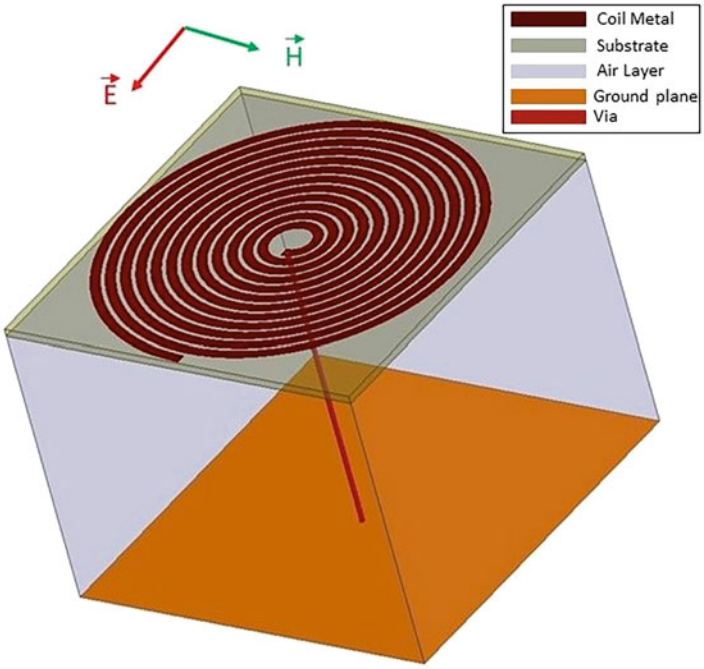

Fig. 11 Unit cell design of the metamaterial structure with the direction of the electric field E and magnetic field H

The results in Fig. 12 show that our metamaterial is well matched and resonates at eight multiple frequencies simultaneously. The frequencies and the S_{11} in dB are given in Table 1.

The next step is to calculate the absorption rate of our design at the resonant frequencies following eq. (2). Since our design is with a ground plane, transmission is neglected. The calculated absorption rates at each frequency of our design are shown in Fig. 13. The absorption rate percentages at each frequency of the metamaterial design are given in Table 2.

The results in Table 2 show that our design is perfect absorber at multiple frequencies simultaneously and well optimized and matched. Moreover, in order to design an energy harvesting metamaterial system, a load resistance has been inserted at the end part of the VIA. To ensure total power dissipation across the load, the resistance should match the impedance of the metamaterial design. Thus, a 282 Ω was given and connected at the end part of the VIA. A 6 $\times$ 6 of unit cells metasurface structure has been designed as shown in Fig. 14.

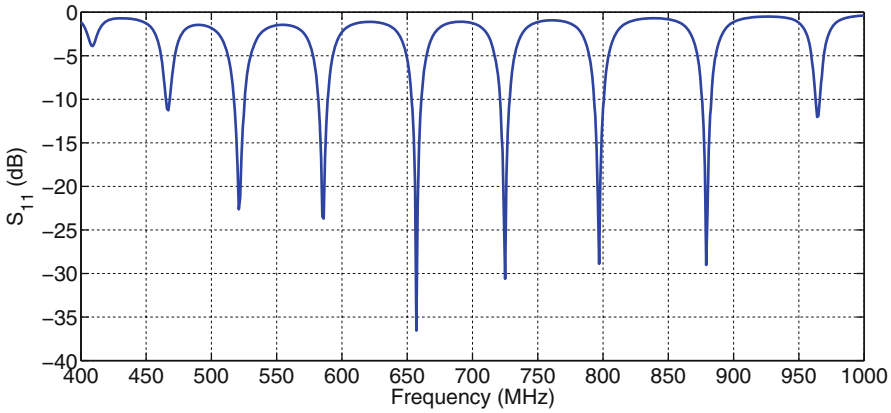

Fig. 12 Reflection coefficient of the proposed design S_{11} in dB with respect to frequency (MHz)

Table 1 The resonant frequencies in MHz and the value of the reflection coefficient S_{11} (dB)

Frequency (MHz)	S_{11} (dB)
462	−12
520	−23
582	−24
652	−37
725	−31
800	−28
875	−28
970	−12

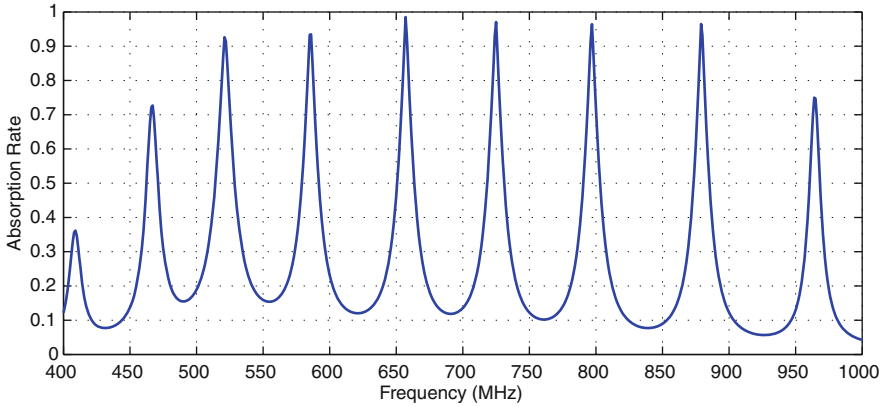

Fig. 13 Absorption rate of the proposed design with respect to frequency (MHz)

Table 2 The resonant frequencies in MHz and the value of the absorption rate (%)

Frequency (MHz)	Absorption rate (%)
410	36
462	73
520	93
582	93
652	98
725	98
800	98
875	98
970	73

Figure 14 shows the metasurface of dimensions 48 × 48 cm and thickness 61.6 mm. A load resistance of 282 Ω is connected at each end part of the VIAs, and a plane wave has been illuminated with the given polarization. Moreover, the power dissipated from the absorbed energy by the metasurface across each given load 282 Ω has been calculated as shown in Fig. 15. The results of the power dissipated at the load from the absorbed energy at each frequency are given in Table 3

The results in Table 3 indicate that the energy absorbed by the metasurface design is mostly dissipated at the load resistance given. This makes us conclude that our design can work as perfect EM energy harvester with high efficiency. Furthermore, the full setup of an EM energy harvester that can be introduced along with the metasurface design is shown in Fig. 16. The concept is a full wave rectification system that has been introduced by Erkmen et al. (2017), where one rectifying system can be used for an array of unit cells.

In this chapter, we have introduced a high efficient and miniaturized metasurface energy harvesting design, operating at multiple frequencies. This can be suitable and compatible for railway application, especially for lower MHz frequencies in the microwave region following the measurements carried on in Heddebaut et al. (2018) and the standard for electromagnetic compatibility for the railway application standard (2017).

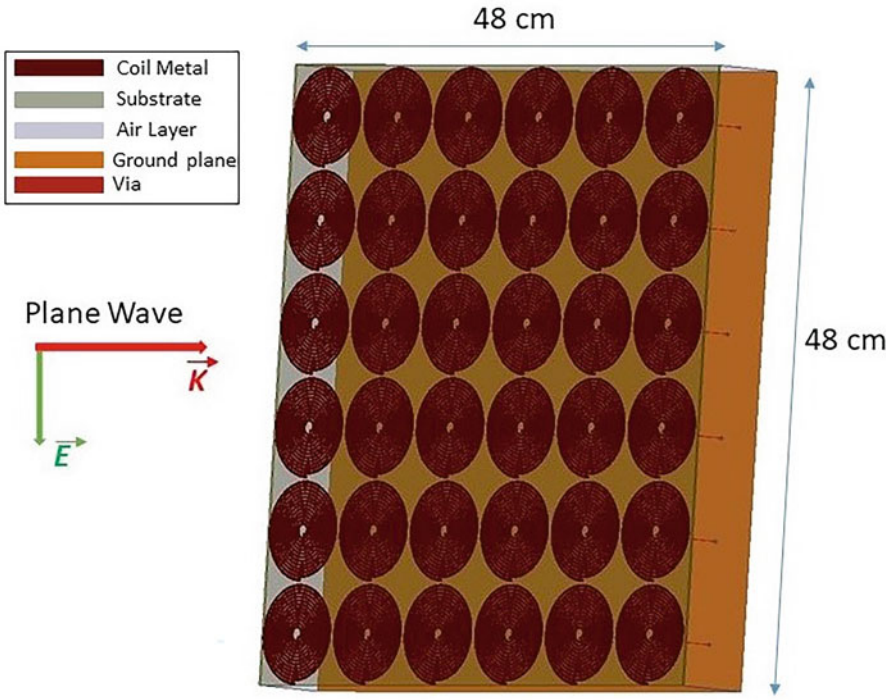

Fig. 14 Metasurface design of 6 × 6 unit cell structures of dimensions 48 × 48 cm and thickness 61.6 mm consisting of an FR4 substrate of thickness 1.6 mm and $\varepsilon = 4.4$ and an air of thickness 60 mm, as well as load resistance of 282 Ω connected to each unit cell end part

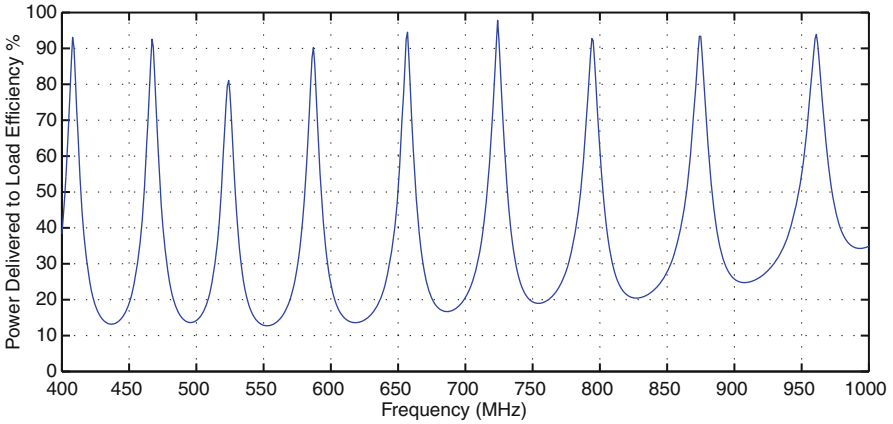

Fig. 15 Power efficiency % of the absorbed energy across the 282 Ω load resistance

Table 3 The resonant frequencies in MHz and the value of the power efficiency dissipated at the load from the absorbed energy (%)

Frequency (MHz)	Power efficiency dissipated at the load from the absorbed energy (%)
410	93
462	93
520	80
582	93
652	95
725	98
800	93
875	93
970	93

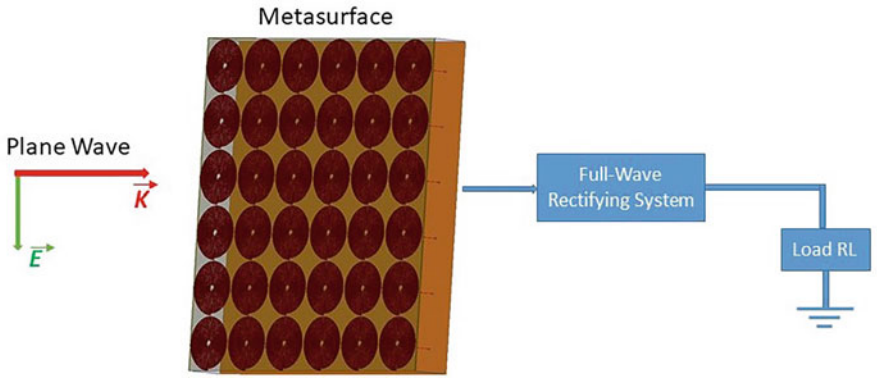

Fig. 16 Full wave rectification energy harvesting system alongside the metasurface design

Conclusion and Future Work

Different challenges are faced from low efficiency technologies to bulkiness for EM energy harvesting in the railway environment. We proposed other interesting solutions with metasurfaces, where we investigated the potential of metamaterial energy harvesting systems based on the concept of absorption. Different designs have been given for a single band of frequency and also multiple bands. The metamaterial energy harvesting designs have shown remarkable results in terms of efficiency and miniaturization at low MHz frequencies, which can be suitable and reliable especially for railway applications. However, different designs have been introduced in the literature based on this concept Almoneef and Ramahi (2014, 2015); Almoneef et al. (2017). Moreover, the fabrication and design of such metasurfaces can be complex especially with the need of multiple rectenna designs at multiple frequencies.

In future work, we can investigate wide-band metamaterial absorbers that have been are introduced by Ranjan et al. (2019) following different concepts in the literature. We can study the potential to design wide-band metamaterial energy harvesters following the same concept we introduced for our designs in this chapter. Moreover, fabrication of a prototype and testing will be done for validation of our results.

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Metamaterials in Medicine

21

S. Raghavan

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Abstract

In the world of tomorrow's Medicine, the Patient and the Doctor are going to be the only non-technological entities. The amount of impact of Engineering in Medicine cannot be described in words. Therapeutics becomes easy once an accurate diagnosis is done. Therapeutics also has developed sky-high. All these Medical miracles happen due to the unimaginable research and contribution of technology in health care. This chapter is an overview of what "metamaterials" (MTM) can do in Medicine both in diagnosis as well as therapeutics. "Ordered lattices of nanostructures with unequal electromagnetic properties and found in nature offer novel sensing possibilities for diagnosis, treatment and management of disease." Negative dielectric constant and Negative permeability are the

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notable and peculiar properties of Metamaterial that are utilized for the applications of "Metamaterials in Medicine." It will not be an exaggeration if it is pointed out that Metamaterial has found importance in most of the medical specialties of health care. Starting from the mega diagnostic instrument Magnetic Resonance Imaging (MRI) to Neuroscience, the usage of MTM is enormous. For example, MRI diagnoses certain abnormalities. But which technology can shield both the medical fraternities and the instrument? Yes. MTM acts as a perfect shielding from Electromagnetic Interference (EMI) and Electromagnetic Compatibility (EMC). Usage of MTM opens up many more advantages. This chapter gives an overview of "Metamaterials in Medicine."

Keywords

Metamaterials · Magnetic Resonance Imaging (MRI) · Neurology · Orthopedics · Dentistry · Gynecology · Surgery Antennas · Lens · Ear Nose and Throat (ENT) · Resonators · High frequency · Unit cells · MEMS · Hyperthermia · Split Ring · Radio waves · Cancer cells · Electromagnetic · Absorbers · Microwaves · Specific Absorption Rate · Cellular phones · Heating · Sensing · Fracture · Graphene · Mammogram

In the world of tomorrow's medicine, doctors and patients alone will be non-engineering entities. Let us mean it literally. "Silent Medical Engineering Revolution," "Breakthrough in Lifesaving Technology," "Point of Care," "Affordable," "reaching the unreachable," and "Medical tourism" are the buzzwords that keep mounting. All mean the same "State of the Health Care for all at affordable prices, unbelievable medical diagnostics art, and therapeutics magic at the time of GOLDEN HOUR emergencies." If only the doctors and engineers join, a miracle will happen in Engineering Medicine and Medical Engineering. The time is ripe for the interactive multidisciplinary medical engineering health technologies. The impact of Microwave Engineering in general and microwave integrated circuits where planar transmission lines are the backbone, in particular, cannot be expressed or described in words. For example, the invaluable noble service of Magnetic Resonance Imaging (MRI) in "Medical Diagnosis" is known to everyone. The contribution of "Computer Tomography Scan" is being felt by all. Biomedical Instruments' role in human healthcare Engineering cannot be expressed so easily. While MRI and CT take care of the diagnosis part which technology will safeguard CT, MRI, etc.? Which Science has started augmenting the rare diagnosis of dreadful disease? ……! The only answer for all these million-dollar questions is "Metamaterial." Metamaterial acts as a perfect shielding (Electro-mechanical Interference EMI) for MRI, CT, and also the medical and paramedical professionals who are continuously exposed to radiation.

The advancement in the medical field has saved the lives of millions of people. The research in electromagnetics is redefined with the advent of metamaterials. The negative refractive index property of metamaterial is finding numerous applications

in the medical field. The unit cell of metamaterial structures has an average size less than the guided wavelength. The split ring resonator exhibits negative permeability while the complementary split-ring resonator has negative permittivity by interacting with the electric field. The sensitive performance behavior of strain, dielectric properties, chemical and biological sensing, the metamaterials are found to be the best suitable candidate for biomedical applications.

THz spectroscopy and metamaterial are used in biomarker detection for the early diagnosis of cancer. The detection of cancer in the cell has witnessed splendid advancement with geometrical variations in metamaterials. An array of gold ring resonators is used to detect cell apoptosis about changes in resonant frequency. The cross-polarization effect is reduced by designing a periodic SRR array on the dielectric layer. The water absorption problem can be overcome by integrating the metamaterial with a microfluidic chip. The sensitivity response can be further increased by the Asymmetric double split-ring resonator.

Magnetic resonance imaging also reaped the benefits of using metamaterial. The metamaterial ring antenna system in MRI helps improve the SNR value of the imaging. An eight-channel RF head coil for human head imaging operating at 1.5 Tesla uses ZOR metamaterial sensors.

Metamaterial (beyond nature and whose properties are not usually found in nature, the theory of Maxwell's equations with negative permittivity and permeability) was coined by Rodger M Walser (in 2001) and pioneered and predicted by "father of radio and millimeter waves" Sir J.C. Bose (in 1898, Dr. Victor Veselago (in 1967)). Sir John B. Pendry discovered the negative value of permeability (Cheng et al. 2011). The term left-handed material was coined by Veselago. Metamaterials have negative dielectric constant, permeability, and earn their properties from structure rather than composition. It is a known fact "Microwaves" which are otherwise called "cm" and "mm" waves cannot be bent. Smith et al. had recognized the gradient refractive index (RI) of the medium resulting in bending the electromagnetic waves. Metamaterials make microwaves bend. Laparoscopic surgery and gastrointestinal operations need a catheter to reach the organs of the human with the help of optical fibers.

Metamaterials' indispensable impact usage can be utilized in every specialty of the medical field apart from a uniquely valuable role in Radar Cross-section reduction, size reduction of the antennas, spatial filters, resonant frequency alteration, high-frequency communication systems and components, stealth technology, etc. The importance of metamaterials in unconventional fields is getting known in recent times only. Gynecology, obstetrics, state-of-the-art diagnostic centers, oncology, orthopedics, neurology, ENT, dentistry, and telemedicine are a few of the divisions of metamaterial applications among an extensive list. In short most of the gadgets used for clinical medical metamaterials roles are slowly being felt. The metamaterial cladding MRI images play a confirmation test for the most dreadful community virus (Fig. 1).

Telemedicine is a branch of medicine that combines communication and information technology, as well as medical engineering and science. The telemedicine system consists of customized hardware and software at both the patient and doctor's

Fig. 1 A Gynecologist (author's wife) delivering a baby. Metamaterial has an impact on fetus diagnosis also. Metamaterials part of Laparoscopy procedure "Hysterectomy" is in development

end, with some diagnostic modalities provided at the patient's end, such as ECG, X-ray, and pathology. They are linked by a communication backbone, which could consist of VSAT or ISDN lines, leased lines, or even telephone lines.

Metamaterial impregnated with conventional microwave engineering will enhance the implantable antenna parameters and flexibility. This mechanism improves the Signal-to-Noise Ratio (SNR) which facilitates many desirable properties. The research in Bio Electromagnetics has gone through a sky change with the advent of Metamaterials. The negative refractive index property of metamaterial that is the result of the interaction of the electromagnetic wave with material at a particular frequency is finding numerous applications in healthcare engineering. This property is not readily available in nature and hence is defined as artificially homogenous materials (Pendry et al. 1999) (by Victor Veselago in 1967). Single negative (SNG) materials, such as mu negative (MNG) and epsilon negative (ENG), or double negative (DNG) materials are available. Artificial and effectively homogeneous with cell size p, negative refractive index providing for inversion of Snell's law, inversion of the direction of propagation, and inversion of Doppler's shift are all characteristics of Metamaterial (Kumar et al. 2011).

In the biomedical field, microwave antennas and sensors can be used for diagnostic purposes such as Magnetic Resonance Imaging (MRI), Microwave Computed Tomography (CT), or therapeutic purposes such as Hyperthermia, Coagulation, and so on. Only correct diagnosis and treatment can shorten the stay in the hospital length, and they are the need of the hour. The focusing and directing capabilities are achieved by utilizing a mix of Split Ring Resonators (SRR) and metallic thin wires

(Roseline et al. 2013). The metamaterials' negative refractive feature makes it behave like a perfect lens (Cheng et al. 2011). Pendry demonstrated that these materials can restore evanescent waves while also providing better image resolution than traditional lenses. It is demonstrated in Rennings et al. (2011) that metamaterial structures may focus even evanescent waves to the appropriate focal point. This characteristic is used in telemedicine to improve the spatial resolution of tissues or organs of interest. Microwave signals are commonly employed in biomedicine because they relieve symptoms more effectively than medications, with "diathermy" being the greatest example.

Implantable microwave antennas and sensors are designed for detection and therapy. Miniaturization mantra is made a reality due to metamaterials intervention for accurate diagnosis. Many antennas are designed in ISM (Industry, Scientific, and Medical) frequency is generally used during design specification.

MTM plays an indispensable role in lifesaving/extending healthcare engineering. The applications include Magnetic Resonance Imaging (MRI), Microwave Computed Tomography (CT) Hyperthermia, Coagulation, breast tumors diagnosis, strain sensing, organic tissue analysis, Hyperthermia, Fetal monitoring, and Specific Absorption Rate (SAR) analysis. Metamaterials significantly have made an impact in the prevention, detection, and therapy of a few disorders with accuracy and in lesser time. Miniaturization (the mantra which facilitated BioMEMS even) in medical devices has come a long way in the medical arena and metamaterial facilitates this magic. Medical devices operate in the ISM (Industrial Scientific and Medical) band of the frequency range.

Metamateral Applications further include cancer detection, imaging, microwave hyperthermia, strain sensing, and reduction of SAR in glucose monitoring for diabetes management to overcome the issues associated with the retrieval of implantable (resorbable MTM), endoscopy to track gastrointestinal passage} point-of-care (POC) technologies, diabetes management, traumatic brain injury (TBI), concussion, stroke, healing fractured bones using split-ring resonators breast cancer, heart failure and brain stroke detection, mammography, telemedicine and Orthopedic injuries. Complimentary Split Ring Resonators (CSRR) loaded Metamaterial Endoscope facilitates tissues abnormalities vision with better enhancement. Cancer detection, imaging, microwave hyperthermia, strain sensing, SAR reduction, and glucose monitoring for diabetes management. To overcome the issues associated with the retrieval of implantable (resorbable MTM), endoscopy in tracking gastrointestinal passage MTM play a greater role. Point-of-care (POC) technologies, diabetes management, traumatic brain injury (TBI), concussion, stroke, healing fractured bones using split-ring resonators breast cancer, heart failure, and brain stroke detection, mammography are other health issues taken care of by the metamaterial inspired technologies.

While X-ray technology has greatly improved the accuracy and efficacy of mammograms, they still rely on ionizing radiation that can be dangerous to the body over time, MRI uses non-ionizing radiation. MRI uses a strong magnetic field and radio waves to generate images of parts of the body.

MTM will facilitate a shorter span time to relieve the patient remaining completely immobile for a longer time as is needed in the conventional MRI.

It can make scans quieter and faster and can increase measurement sensitivity by a factor of five by increasing SNR.

Radiofrequency Micro Electromechanical Systems (RF MEMS) based metamaterial sensors find prominent applications in the measurement of strain in telemedicine due to the inherent properties of the high-Q factor, sensitivity, and sharp resonance multifunctionality.

"Surgical wound monitoring by MRI with a metamaterial-based implanted local coil" was reported by Hanan Kamel, Richard R.A. Syms, Evdokia M. Kardoulaki, and Marc Rea.

To reduce the postoperative convalescence. Noninvasive or minimally invasive surgical procedures using laparoscopy technology will be certainly a preferred one. Of course, childbirth is one example that cannot be done by this laparoscopy method which involves optical fibers and metamaterial amalgamation with the machine learning process also. "Imitating human soft tissue on basis of a dual-material 3D print using a support-filled metamaterial to provide bimanual haptic for a hand surgery training system" was reported by Johannes Maier, Maximilian Huber, Christophe Palm N Weiherer, Michaela (Fig. 2).

Implantable antennas are required for physiological monitoring and administering therapy. Miniaturization is needed and that is answered by metamaterial.

One of the highly advanced metamaterials with many unique and varied properties is GRAPHENE which plays a greater role in the diagnosis and treatment of neurological diseases, according to a report. "It is expected that neurosurgery will significantly benefit from several graphene-based technological developments in the next decades." Mattei and Rehman write. Richard Feynmann's sentence "There is plenty of room at the bottom" which inspired many to come out with lots of engineering wonders in MEMS first, RF MEMS next, and finally "BioMEMS." Incidentally MEMS and RF MEMS impact in the biological field of medicine is invaluable. It's again an ocean. Orthopedic injuries are very common and present a financial burden to the medical community. The development of RF MEMS-based metamaterial sensors with multiple split-ring resonators in nested architecture is used to measure strain telemetrically. The high-Q factor, sensitivity, and sharp resonance make metamaterial supersede conventional sensors.

Graphene finds its usage for neuron surgeons in neuro-oncology, neuro-intensive care, neuron regeneration research, peripheral neurosurgery, functional neurosurgery and spine neurosurgery, cancer treatment, a prominent role in tumor-targeted imaging, as well as possible new therapeutic approaches involving photothermal or alternating electrical field stimulation therapies, and intensive care unit monitoring. Neurological monitoring in patients with stroke and traumatic brain injury, to promote regeneration of nervous system tissues, functional neurosurgery, improved electrophysiological monitoring systems, epilepsy, and movement disorders, spinal surgery, help in the next-generation instrumentation for spinal surgery, implants biocompatible Graphene-based metamaterial will create an unprecedented impact (Tobias et al. 2014).

The malignant cells have significantly higher water content than normal tissues. The permittivity and the conductivity of the malignancy are different from those of

Fig. 2 Laparoscopy instruments

normal tissue at high frequencies. Change in the basic parameters namely permittivity and conductivity of the tissues due to water content facilitates cancer detection. This is more predominantly seen at microwave frequencies where "windowing" is a notable phenomenon.

MammoWise, a radio-wave imaging system for breast cancer screening developed by Metamaterial Inc., expand screening procedures to a large pool of younger patients in a cost-effective and reliable manner and enhanced early detection, without side effects and stress.

Magnetic resonance imaging, MRI, the first full-body scans of a human to diagnose cancer for medical applications were developed in 1970.

Frequency selective surface (FSS) (spatial filter is one of the patents of Marconi and was developed and contributed by Sir. J.C. Bose and B. Munk) which belongs to the family of metamaterial plays a significant role in medicine even though it is very much applicable in communication and RADAR technology. It is used as an

absorber to suppress unwanted radiation in hospitals. As a band-stop filter, it suppresses unwanted signals and eliminates external signal interference with clinical data. To monitor patients' intake of doses a dosimeter in the hospitals by increasing the antenna gain. The medical body area network (MBAN) monitors physiological signals like ECG, EEG, heart rate, and blood pressure. High isolation between the human body, antenna, and external environment is obtained when FSS combines with Electronic Photonic Band Gap structures. MRI performance enhancement and SAR reduction are also obtained with the help of FSS.

FSS in Biomedical

Frequency selective surfaces (FSSs) popularly known as periodic arrays have the filtering properties of electromagnetic waves in free space. These periodic arrays are extended in two dimensions to cover or protect the sophisticated equipment to protect from unwanted EM radiation. The EM spectrum spreads from 300 MHz to 30 GHz, with which several types of equipment are connected. It is always important to protect the specific device/equipment to avoid malfunctioning by the effect of unwanted radiation. FSS is found effective in controlling the transmission of EM (electromagnetic) waves that pass through the human body to protect from hazardous radiation. The best such example is to protect the pacemakers from interference from unwanted frequency bands.

A novel FSS structure is proposed based on biodegradable materials for biological tracking and sensing requirements. The proposed FSS is gradually broken down and dissolved into biofluids after its functional usage period. This shows its advantages of harmlessness and convenience. The study was made to observe the resonance frequency for various incident wave angles and polarization cases. Also, the study was conducted for defection cases in the dissolving process of the degradable material. The model shows promising performance with high stability and superiority for biomedical applications (Zhao et al. 2014) (Fig. 3).

The easy and low-cost fabrication availability made the FSS a profound candidate to use for bio-medical application. Also, the different fabrication technologies like micromachining, surface machining, ink-jet printing, 3-D fabrication, and conformal fabrications made the FSS available for most sophisticated and crucial shielding applications.

Matameterial in Health Care Technology [Raghavavan and Rajeshkumar 2013*]

Communications systems and biomedical sensors can be used to supervise a patient's condition outside of the clinical setting. Over the last decade, there has been a rapid increase in the demand for surgically implanted telecommunication devices for clinical applications. Microwave antennas and sensors are critical

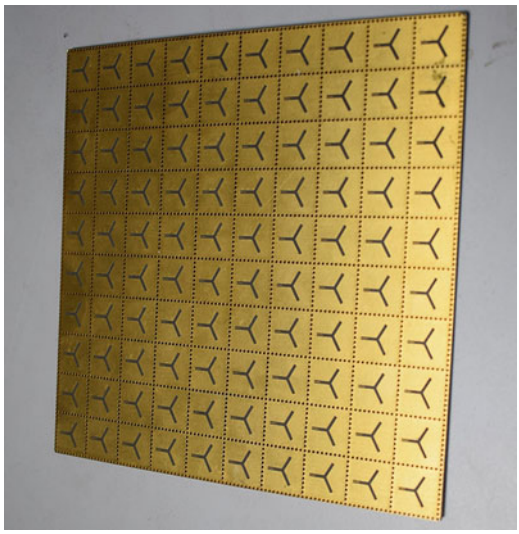

Fig. 3 Periodic FSS structure (prototype) (Krushna Kanth and Raghavan 2018)

components of medical location tracking. The unusual characteristics of metamaterial have led to the development of more efficient metamaterial antenna elements, sensors, and metamaterial lenses for miniature wireless systems. Metamaterials respond extremely sensitively to strain, dielectric media, chemical, and biomedical sensing application forms.

Introduction

Due to the requirement for early disease detection and continuous monitoring of physiological data, the implementation of wireless telemetry systems in medicine has expanded dramatically in recent years. Because they facilitate communication between the patient and the base station, microwave antennae and sensors are essential parts of modern telemetry systems. Electromagnetic (EM) properties of metamaterials can be changed to something that does not present in nature. As a result of their unique electrical characteristics at mm wave frequencies, metamaterials have piqued the curiosity of microwave engineers and physicists alike. A split-ring resonator (SRR) is one of the metamaterial particles that offer negative permeability (Pendry et al. 1999), while a complementary split-ring resonator (CSRR), the duality of SRR, interacts with the electric field and introduces negative permittivity (Pendry et al. 1999), which are most used in biomedical sensors. It is well known that SRRs are resonant structures, and consequently, it is possible to employ these resonances as well as radiating modes. SRRs are very compact, and it is easy to design them with dual-band characteristics. The review of metamaterials towards different healthcare applications is described in the subsequent sections.

Metamaterial-Based Antennas for ISM Band

The ISM radio bands are sections of the wireless spectrum that have been reserved globally for the use of radiofrequency (RF) energy for industrial, scientific, and medical uses other than communication. In slotted microstrip antennas, metamaterial structures such as EBG (Electromagnetic Band Gap) structures can be used to improve efficiency and return loss features (Kumar et al. 2011). An EBG structure has been employed as a metamaterial above the ground plane in a patch antenna with a slotted ground plane. The developed antenna exhibits adequate properties in two radio frequencies: the UMTS band (2067–2245 MHz) and the ISM band (2390–2795 MHz) (Kumar et al. 2011).

A wireless endoscope is a capsule-shaped device that is used to monitor and/or treat the gastrointestinal tract. Xiaoyu Cheng et al. developed, constructed, and analyzed a 2.4-GHz ISM-band folding patch antenna for a wireless capsule endoscope application. A self-packaged folded patch antenna with a quasi-omnidirectional radiation pattern has been built on a flexible substrate and folded in a cylindrical shape (Cheng et al. 2011). For a wireless endoscope, the CSRR loaded patch has a small design, EMI protection, and a quasi-omnidirectional radiation pattern (Cheng et al. 2011). When compared to its typical patch equivalent without CSRR loading, the CSRR contributes a 74% size reduction (Cheng et al. 2011). The enormous enclosed compartment might be used to house digital controllers, EMI sensitive sensors, cameras, batteries, and other electronic devices.

A high gain pencil beam array antenna for the applications within the 2.4 GHz ISM band was designed by A. Rennings et al. (2011). The high-gain pencil-beam array antenna is based on composite right/left-handed (CRLH) metamaterial transmission-lines, referred to as CRLH meta-lines.

A metamaterial-inspired handset antenna is designed to reduce the specific absorption rate (SAR) for mobile communication. The SAR reduction was achieved by miniaturizing the physical size of the antenna for the resonance frequency of 2.4 GHz, by devising a metamaterial-inspired radiator much smaller than the conventional planar inverted F antenna (PIFA) (Kahng et al. 2012). Further, the dual-band microstrip patch antenna was designed for implantable medical devices, which covers the ISM (2.4–2.48 GHz) band. The antenna has a multilayer configuration where the feeding line and the radiating elements are located at different levels. The radiating elements of the antenna are a split ring resonator (SRR) coupled to a spiral and both elements are short-circuited to the ground plane to achieve size reduction (Sanchez-Fernandez et al. 2010).

Metamaterials in Cancer Detection

Cancer is the uncontrolled growth of abnormal cells in the body, called malignant cells. Developing microwave devices and combining them with structures inspired by metamaterials, can lead to a very cost-effective device that can localize with high precision and abnormality within the human body. The basic principle behind cancer

detection is a small change in the water content of biological tissues produces changes in the permittivity (ε) and conductivity (σ) values of the tissues (La Spada et al. 2011). The malignant cells have significantly higher water content than normal tissues. Hence the permittivity and the conductivity of the tumor are higher than those of normal tissue at microwave frequencies. The proposed biosensor consists of an array of complementary metallic metamaterial resonators. The reason for choosing SRRs are their strong response to an electromagnetic field (La Spada et al. 2011).

An electromagnetic source generates an electromagnetic wave impinging on the metamaterial array and a detector is placed to reveal the signal after the array. The biosensor without any material under test has a specific resonant frequency (La Spada et al. 2011). The variation of the permittivity caused by the presence of the material under test acts on the capacitance of the resonators, leading to a high-sensitive variation of the sensor resonant frequency. Thus, the shift in resonant frequency and the shape of the response is extremely useful to detect the tumor. Further, the frequency multiplexed two-dimensional sensor array based on SRR for organic tissue analysis has been proposed by Margarita Puente et al. A frequency multiplexed two-dimensional sensor array was developed using microstrip-line excited SRRs which are designed to have different resonant frequencies as shown in Fig. 1a (Puentes et al. 2012). Therefore, the resonant frequency of individual SRR will indicate the dielectric properties of the MUT around the ring and its location within the array (Puentes et al. 2012).

Terahertz spectroscopy, Metamaterial with a microfluidic chip along with double split-ring resonators help a lot in early cancer detection especially to avoid water absorption problems.

Metamaterials inspired antenna arrays are used to detect tumors and cancer. By measuring the reflected or transmitted signals malignancy can be detected.

Metamaterials in Medical Imaging

A super lens or perfect lens is a lens that uses metamaterials to go beyond the diffraction limit. The diffraction limit is an inherent limitation in conventional optical devices or lenses. In 2000, a type of lens was proposed, consisting of a metamaterial that compensates for wave decay and reconstructs images in the near field (Pendry 2000). And most importantly both propagating and evanescent waves contribute to the resolution of the image. In 2004, the first super lens with a negative refractive index provided resolution three times better than the diffraction limits and was demonstrated at microwave frequencies. In 2005, the first near-field super lens which exceeded the diffraction limit was demonstrated by N. Fang et al..

The higher focusing resolution will be provided by a flat left-handed metamaterial (LHM) lens if compared to the convex dielectric lens and elliptical reflector focusing system. High-resolution near-field microwave target detection and imaging with a flat LHM lens can be implemented (Wang et al. 2007). Due to higher focusing resolution, the LHM lens has the potential to acquire higher imaging resolution and easy in-depth scanning, which will simplify the detection system design (Wang et al. 2007).

MRI image quality can be improved by placing the metamaterial adjacent to the organ to be imaged due to the improvement of the Signal-to-Noise Ratio (SNR).

X-rays were discovered in 1895. The use of x-ray imaging to detect breast cancer was suggested as soon as the early 1900s, but it was not until the 1960s that image quality improved enough for mammography to become an accepted technology. X-ray technology involves ionizing radiation which is harmful and this is not so in non-ionizing radiation technology.

MRI Metamaterials Influence in MRI

It is found that metamaterials improve the quality of Magnetic resonance imaging (MRI). MTM facilitates lower power MRI scanners which then felicitates quality images with lesser cost, this becomes a boon to many hospitals that make it available at affordable costs as compared with powerful devices. Boston University confined on improving the fidelity of the radio waves that are emitted by the body into the MRI and dispense with powerful magnets which are otherwise used to improve MRI image quality. The device is made to off its functionality when the scanner is transmitting its radio signal and turns on immediately after, only boosting the power of the return signal.

Magnetic Resonance Imaging (MRI) is superior to X-rays in picture quality and non-ionization technology. MRI scanner that uses a large magnet in which the patient is passed through has become an indispensable diagnostic tool and gained the benefits and unique advantages of using metamaterial. Signal-to-noise ratio is improved by a metamaterial ring antenna system.

Metamaterial Inc. is working to develop metasurface a passive resonator designed to increase the SNR of an MRI machine up to 40 times, resulting in higher image quality and shorter scanning time. The image resolution is directly proportional to the field strength of the magnet. Cost becomes more as the field strength increases.

Generation of conventional MRI images is based on the spatial variance of the phase and frequency of the emitted and absorbed RF energy (Sanchez-Fernandez et al. 2010). These common principles are just altered with the use of metamaterials that employ collimation of RF energy. Higher resolutions can also be achieved by increasing scanning time at an increasing cost, however; SNR increase will even shorten the scan time.

Metamaterial Inc. (CSE: MMAT) has developed metamaterial films that cancel skin reflections, increasing the signal-to-noise ratio. The company is developing several applications which integrate this technology, including pain-free, noninvasive blood glucose monitoring for diabetics; earlier breast cancer screening; higher quality, lower cost MRI imaging; and more rapid treatment of stroke.

Metamaterial makes an MRI quieter and faster.

Researchers at Michigan Technological University have found a way to possibly solve one of the biggest challenges, getting light waves to pass through the lens without getting consumed.

Fig. 4 MRI wherein Metamaterial play dominant roles for enhanced performances

Negative index metamaterial, in addition to passing through propagating light waves, also amplifies decaying light waves. Metamaterial using Specific Absorption Rate analyses the SAR value of the fetus of a pregnant woman and helps a lot in the antenatal process.

Magnetic resonance imaging has become an indispensable tool for medical diagnostics. Metamaterials can increase measurement sensitivity, make scans faster and quieter, and examine more pleasure, since the beating of the heart. Teams of researchers from the Fraunhofer Institutes MEVIS and FHR working on a Fraunhofer project (Dr Thomas Bertuch, team leader at Fraunhofer FHR) succeeded in boosting the sensitivity of MRI machines under certain circumstances many times over (Figs. 4, 5, and 6).

MTM in EMI EMC

Metamaterials are also helpful in protecting the human body against electromagnetic radiation. Metamaterial structures when incorporated with implantable antenna helps to reduce the SAR value. A decrease in 70% of SAR is witnessed when metamaterial is placed between the antenna and forearm. Wearable devices emit harmful radiation and this poses a serious health concerns as the device is in direct contact with the body. Artificial magnetic conductors, a class of metamaterials, are used to control the radiation pattern of the antenna and prevent harmful electromagnetic waves. Frequency selective surfaces are effective shielding for CT, etc. The antenna is an

Fig. 5 MRI diagnosis procedure

Fig. 6 CT SCAN which also has better output images due to the impact of MTM

Fig. 7 C-arm equipment used as an imaging scanner intensifier. C-shaped arms are used to connect the x-ray source and x-ray detector. Used during surgical, orthopedic, and emergency care procedures

important component in the wireless endoscope. When the endoscope is developed with CSRR, loading tissue abnormalities can be seen with better enhancement (Figs. 7, 8 and 9).

MEMS Fabrication Instrument. MEMS Metamaterials-based sensors play a prominent role in biomedical engineering.

Metamaterials in Microwave Hyperthermia

Hyperthermia is a technique that uses microwaves to destroy or ablate damaged soft tissue by heating it to a temperature that triggers cell death. The generator generates microwave energy, which is transferred to the patient via the antennas. When the temperature of tumor cells is raised, the cell membranes are damaged, and the cancer cells are destroyed. The goal of hyperthermia cancer treatment is to deliver a precisely regulated dose of heat to the malignant tumor and surrounding bodily tissue. The capacity of the metamaterial lens' negative-refractive index (NRI) to focus a source's electromagnetic field is its most notable feature (Wang et al. 2007).

As a result, it can provide an adequate concentrating spot in biological tissue, as required in microwave hyperthermia treatment (Gong and Wang 2009). The

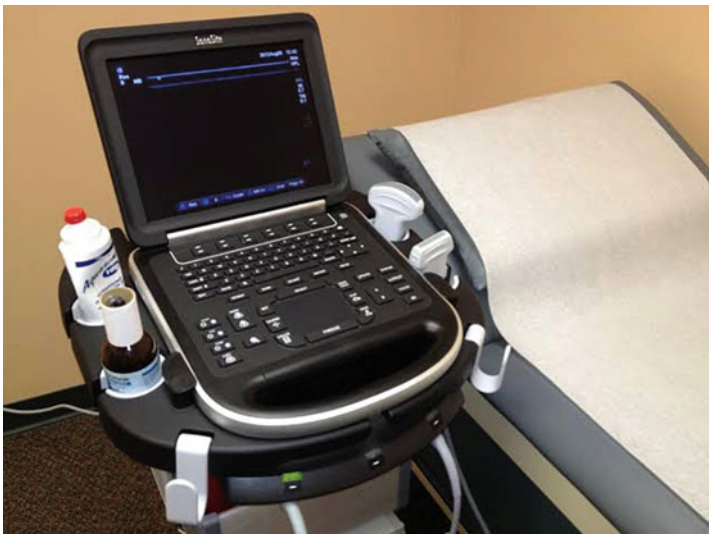

Fig. 8 Physicists in the US have built an acoustic lens from artificial "metamaterials" and used it to focus ultrasound waves to a tight spot. They say that modifying their apparatus could lead to very-high-resolution ultrasound imaging, for use in medicine and non-destructive material testing

Fig. 9 MEMS fabrication: one stage

microwave radiation released from the microwave source was focused using a flat metamaterial slab as a lens. A planar array of split-ring resonators is sandwiched between two parallel metallic plates, and the split-rings are excited by a tiny loop antenna (Velazquez-Ahumada et al. 2011). Furthermore, the ability to easily modify the position of the focusing or heating area in the tissue region increases the likelihood of successful microwave hyperthermia (Tao and Wang 2012). The heating

point in tissue can be changed by shifting the source, without the need for a sophisticated deployment and control system as in a traditional array applicator (Gong and Wang 2009). Conformal microwave array applicators with correct source spacing and a low loss left-handed metamaterial lens have recently been proposed for hyperthermia treatment of vast area tumors. Joint heating can be used to perform microwave hyperthermia in this case.

The negative refractive index of metamaterial helps to localize the target spot in hyperthermia. Hyperthermia makes considerable use of a flat left-handed lens. Better performance is achieved with multiple lens systems because more microwave energy is focused on tissue and the focusing depth of numerous microwave sources is enhanced (Tao and Wang 2012).

Metamaterials in Wireless Strain Sensing

Strain sensors based on RF-microelectromechanical systems (MEMS) that make use of recent metamaterial advancements. R. Melik et al. created a wireless strain sensor to monitor the healing processes of shattered long bones... Because of the inability to monitor fracture healing, approximately 10% of fractures do not heal adequately. By monitoring the amplitude of operating frequency shift under applied load, the SRR-based sensor exhibits a characteristic resonance frequency under no load, the progression of fracture healing can be observed (Melik et al. 2009). The nested SRR-based strain sensor was developed after the single SRR-based wireless metamaterial strain sensor was produced to improve sensitivity. Multiple SRRs are combined in a compact nested architecture on a single chip in nested metamaterial-based strain sensors to considerably improve telemetric sensing sensitivity (Melik et al. 2010). The nested SRR-based strain sensor outperforms the standard SRR sensor when metrics like unloaded operating frequency, sensitivity, and nonlinearity errors are compared (Melik et al. 2010).

SAR Reduction Using Metamaterial Absorbers

The specific absorption rate (W/kg) is a unit of measurement for how much radiation is absorbed by the human body. The metamaterial absorbers work on the idea that when one of the effective medium parameters is negative and the other is positive, the medium will show a stopband. Metamaterials' stopbands can be tailored to fit within cellular phone operating bands (Hwang and Chen 2006). The SRRs were designed to operate in the 900 and 1800 MHz bands. Around the 900 and 1800 MHz bands, the structure parameters of SRRs were designed to be the effective medium parameter is negative. When a metamaterial absorber is compared to other SAR reduction approaches such ferrite sheet attachment and using a perfect electric conductor, it is found that metamaterials can reduce the peak by 22.2 percent (Hwang and Chen 2006). Although the ferrite sheet attachment effectively reduces peak SAR, it also significantly reduces radiated power from the antenna.

Metamaterials can be created at the desired operating frequency, as opposed to using a ferrite sheet (Hwang and Chen 2006). The metamaterials can be constructed on a circuit board and readily integrated into a cellular phone.

Engineers and physicists are increasingly needed to concentrate on and create technology specifically for healthcare. Metamaterials have been a hot topic in study over the last decade, with possible applications in healthcare. A review of existing metamaterial antennas designed to work in the ISM band is presented, as well as the usage of metamaterials in cancer diagnosis, imaging, microwave hyperthermia, strain sensing, and SAR reduction. Metamaterials are also used in glucose monitoring to help with diabetic treatment. Furthermore, the utilization of resorbable metamaterials will eliminate the problems associated with implantable device retrieval after their operational lifetime. Metamaterials in biomedical applications will soon open up a large area of research (Raghavan and Rajeshkumar 2013).

Metamaterial –"Quotable Quotes" of few Scientists

"A lot of people are surprised by its simplicity," said Zhang. "It's not some magic material. The 'magical' part is the design and the idea."

"If you can deliver something that can increase SNR (Signal to Noise ratio) by a significant margin, we can start to think about possibilities that didn't exist before," said Anderson, such as the possibility of having MRI near battlefields or in other remote locations. "Being able to simplify this advanced technology is very appealing," he said.

"Could a small ring-like structure made of plastic and copper amplify the already powerful imaging capabilities of a magnetic resonance imaging (MRI) machine? Xing Zhang, Stephan Anderson, and their team at the Boston University Photonics Center, can picture such a feat." With their combined experience in engineering, materials science, and medical imaging, Zhang and Anderson, along with Guangwu Duan and Xiaoguang Zhao, developed a new magnetic metamaterial that can increase MRI quality and cut scan time in half, as described in Communications Physics.

According to Zhang and Anderson, their magnetic metamaterial might be used as an additive technology to boost the imaging capability of lower-strength MRI equipment, allowing clinics to see more patients while lowering expenses, without the hazards associated with employing higher-strength magnetic fields. They also foresee using the metamaterial with ultra-low field MRI, which employs magnetic fields that are thousands of times lower than those utilized in present machines. This would allow MRI technology to become more freely available around the world. According to Anderson, a School of Medicine professor of radiology and vice chairman of research in Boston Medical Center's radiology department, "this [magnetic metamaterial] creates a crisper image that may be produced at more than double the speed" of a present MRI scan.

Wolters Kluwer Health: Lippincott Williams & Wilkins. "Revolutionary 'metamaterial' has potential to reshape neurosurgery." ScienceDaily, 25 April 2014.

"The intelligent metamaterial consists of an array of metallic helical resonators closely packed with [a passive sensor]," said Xiaoguang Zhao, one of the

researchers. "When the high-energy radio waves are coming in, the metamaterial detects the high energy level and 'turns off the resonance automatically. With low-energy radio excitation, the metamaterial [turns on] the resonance and enhances the magnetic component of the radio wave."

"Aluminum and silver are the best choices so far in the visible light spectrum, not just for a perfect lens but all metamaterials," Güney says, explaining that metamaterials have been successfully created with these metals, although they still tend to absorb light waves. "Loss—or the undesired absorption of light—is good in solar cells, but bad in a lens because it deteriorates the waves," he explains

"Our aim is not to replace current breast screening or stroke detection imaging methods, but rather to make them accessible to all, regardless of their age, location, or wealth. This requires low-cost, portable POC devices that are widely accessible but also accurate enough to be trusted by clinicians and doctors. Our imaging algorithms and metamaterial solutions to enhance data quality gives us confidence that such devices can become reality shortly with the use of radio-wave technology that is affordable and operation-independent," said Dr Panagiotis Kosmas, Co-Founder of MediWise, and META's VP in Biomedical Engineering

And if additional magnetic field switching for imaging can be eliminated, the process also becomes much faster. "According to theoretical calculations, our technology should enable us to complete scans up to one thousand times faster. Only when the experiments have been done will we know just how fast that will be in practice," says Günther. Patients would then be able to enjoy much faster and quieter examinations.

"Our metamaterial can be embedded directly into the patient table of any commercially available MRI scanner. However, in the future we see even more potential in the concept of special smart clothing for MRI scanning," explained Alexey Slobozhanyuk, a researcher at the International Laboratory of Applied Radiophysics.

"If the MRI works with high-frequency coils that are placed on the patient's body, depending on the question, we can improve the dynamics by up to 20 per cent," says Dr. Thomas Bertuch, the team leader at Fraunhofer FHR. "If the coils installed in the MRI machine are used, the measured signal can even be increased five-fold."

"With our device, we can pick up and transmit the evanescent waves, which contain a substantial fraction of the ultra-subwavelength information from the object, so that we can realize super-resolution acoustic imaging," said first author Jie Zhu, a post-doctoral fellow in the Center for Scalable and Integrated Nanomanufacturing (SINAM), a National Science Foundation-funded Nanoscale Science and Engineering Center at UC Berkeley.

"Without the metamaterial, it would be impossible to detect a deep sub-wavelength object at all," Yin said.

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Metamaterial-FSS for THz Applications

22

Rahul Kumar Jaiswal, Nidhi Pandit, and Nagendra Prasad Pathak

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Abstract

This chapter provides an overview of the metamaterial-Frequency Selective Surfaces (FSSs) for terahertz (THz) applications. In order to support various next generation wireless applications, the demand of the high data rate capacity

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systems keeps on increasing day by day. Lower frequency ranges such as microwave frequency bands offer smaller bandwidth, and these bands are already congested because of supporting the several wireless applications. To achieve a higher data rate, larger bandwidth is required, which can be obtained at high frequency, i.e., utilizing THz band. In this chapter, first, the electromagnetic spectrum is discussed with the focus of THz bands and its applications. Frequency selective surfaces (FSSs) are currently an active area of research due to their capability to manipulate the EM waves and unprecedented features. FSSs act as spatial filters, i.e., some particular band of frequency is transmitted, and others are reflected and found a wide applications in communications, radar systems, and sensing. However, traditional FSSs have the unit elements sizes which are in the order of half-wavelength at frequency of operation, which limit its applications, and hence they need to be miniaturized for use in various applications with better performances. Electromagnetic metamaterials are the man-made structures exhibiting the unique electromagnetic characteristics, which do not occur naturally. Recently, metamaterial has shown its capability toward the miniaturization of the microwave structures as well the antennas. In this regard, metamaterial concept plays a vital role in the miniaturization of the FSS by using subwavelength unit elements. Here, a brief overview of the metamaterial inspired FSS is discussed in detail. In the last, simulation setup and design flow for designing FSS using EM simulator such as HFSS and CST are described.

Introduction: Overview of EM Spectrum and THz "Gap" and Its Related Applications

Terahertz (THz) frequency region of the electromagnetic spectrum, which typically spans between millimeter wave and infrared, lies in the range from 0.1 THz to 10 THz and represents a vast spectrum resource with enormous potential applications (Siegel 2002). In terms of the wavelength, the range varies from 3 mm to 30 mm as shown in Fig. 1.

This range of frequency is also called as THz gap because technology development is in its initial phase in this frequency region as compared to the radio frequency (RF)/microwave frequency ranges (Nagel et al. 2002; Bahl 2003; Steve 2006; Siegel 2007; Tonouchi 2007; Nagatsuma et al. 2013; Akyildiz et al. 2014; Kurner and Priebe 2014; Seeds et al. 2015; Sengupta et al. 2018; Song and Nagatsuma 2011). Researchers are continuously trying to develop theories and other equipments to design, fabrication, and characterization of the THz components and systems.

With the rapid development of the fabrication and characterization facilities at high frequencies (up to THz ranges), a huge development is taking place in THz domain. For the realization of the high-frequency integrated circuits, monolithic and hybrid technology using planar transmission line is widely used. For example, microwave monolithic integrated circuits (MMIC) technology, all the lumped components (resistor, inductor, and capacitor), as well as diode/transistors are fabricated

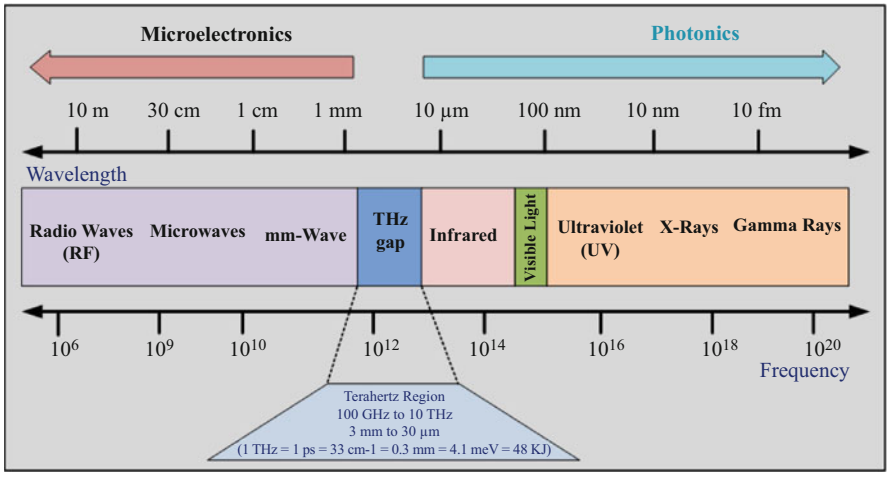

Fig. 1 Electromagnetic spectrum

monolithically on semiconductor substrates like Si, GaAs, etc. (Pickwell and Wallace 2006). However, in hybrid technology, lumped components (resistor, inductor, and capacitor) chips, and transistor chips are integrated using planar transmission line, e.g., microstrip and CPW and are fabricated using PCB technology as a conductor pattern on the substrate (Bolivar et al. 2003).

Due to the smaller wavelengths, greater bandwidth, and more interaction with atmospheric constituents, THz integrated circuits (ICs) have found their broad applications in diverse engineering and scientific fields, e.g., satellite applications, biomedical applications (Huang and Wang 2011; Piro et al. 2016), nondestructive technology (Phillips and Keene 1992; Zeitler et al. 2007), cellular communications like 5G and beyond (Ostmann and Nagatsuma 2011; Federici and Moeller 2010; Federici et al. 2005), submillimeter wave astronomy (Rotaru and Sykulski 2011), military and defense for security and imaging, and sensing applications (Collin 1991; Zhang and Li 1999; Balanis 2012; Ma et al. 2013; Cong et al. 2015).

In the recent years, THz frequency bands have been utilized for many applications. They have the potential to meet the demands of the emerging next-generation communication systems and Gigabit wireless applications, medical and bioimaging, sensing applications, and cognitive radars.

Theory of Periodic Structures

Electromagnetic periodic structures have been extensively used from decades with the antennas for a variety of RF/millimeter wave applications due to their unusual electromagnetic characteristics such as improved gain, and radiation response by utilizing the electromagnetic bandgap structures and also controlling the EM absorption, reflection, and transmission. These EM periodic structures have been classified into EM

metamaterials, frequency selective structures (FSSs), and electromagnetic bandgap structures (EBG). In these EM periodic structures, metamaterials are the subwavelength man-made materials and exhibit unusual electromagnetic characteristics, which are not found in nature, whereas electromagnetic bandgap structures and frequency selective structures are the spatial filters providing band-pass or band-stop properties for the incident EM waves. The good thing with EBG structures is that it can work in all angles of polarization of the incident wave whereas FSS structures are limited to some specific angles of the incident EM wave. Moreover, the electromagnetic band-gap characteristics of the EBG structures direct the EM waves through the periodic structures so that they can be used to reduce the surface waves from the antenna. This property is also extended to use the EBG as high-impedance surfaces as well as artificial magnetic conductor (AMC) along with ground plane to design the low-profile antenna.

Floquet Theorem for Periodic Structure

Floquet theorem is used to analyze the periodic systems in a simpler manner. This theorem states that, "In a periodic system, for a given mode of propagation at a given steady-state frequency, the fields at one cross section differ from that one period (or an integer multiple of periods) away by only a complex constant" (Munk 2000; Caloz and Itoh 2005). Figure 2 shows the one-dimensional periodic structure with a period "d."

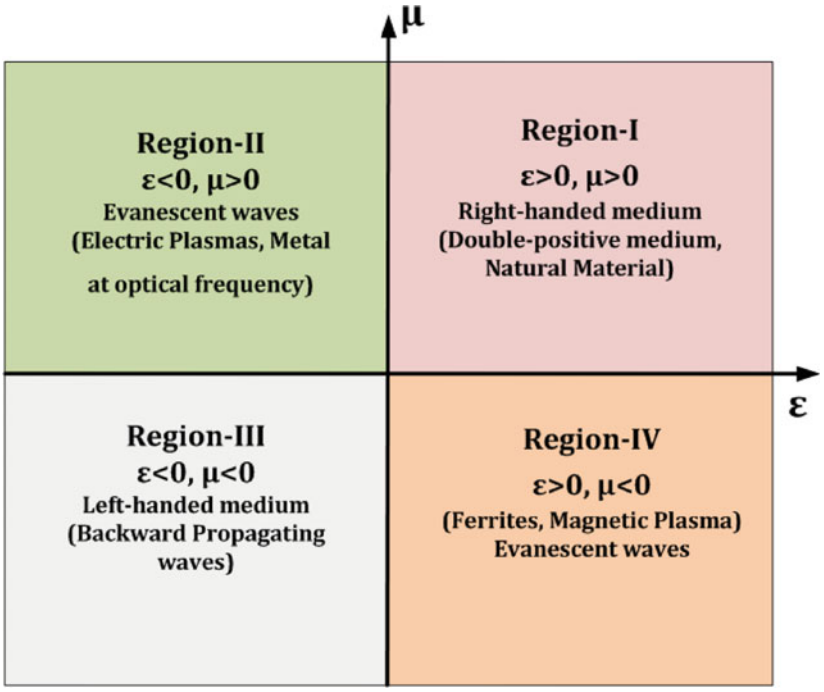

Fig. 2 Classification of material (based on the permittivity s, and permeability p)

The periodicity in the periodic structure forces the field to be periodic. If one defines the field as Y(x), then the Floquet theorem states that the field in a period distance differs only by a complex constant $\psi(x + d) = B\psi(x)$. If one moves an integer number of periods n, then the field becomes $\psi(x + nd) = B^n\psi(x)$. In general, the complex constant B can be expressed as an exponential $e^{jk_{x0}d}$, where k is a complex constant. The field relation between neighboring unit cells would then become

$$\psi(x + nd) = e^{jk_{n0}d}\psi(x) \tag{1}$$

Now, a periodic function can be defined as

$$P(x) = e^{-jk_{n0}d}\psi(x) \tag{2}$$

To show that P(x) is a periodic function,

$$P(x + d) = e^{-jk_{n0}d(x+d)}\psi(x + d) = e^{jk_{x0}x}e^{jk_{x0}d}e^{jk_{x0}d}\psi(x) = e^{jk_{x0}x}\psi(x) = P(x) \tag{3}$$

Since P(x) is a periodic function, it can be expressed as Fourier series

$$P(x) = \Sigma_{n=-\infty}^{\infty} a_n e^{j\frac{2\pi n}{d}x} \tag{4}$$

Substituting (4) in (2)

$$\psi(x) = e^{jk_{x0}d}\Sigma_{n=-\infty}^{\infty} a_n e^{j\frac{2\pi n}{d}x} = \Sigma_{n=-\infty}^{\infty} a_n e^{jk_{x0}x}e^{j\frac{2\pi n}{d}x} = \Sigma_{n=-\infty}^{\infty} a_n e^{jk_{xn}x} \tag{5}$$

where,

$$k_{xn} = k_{x0} + \frac{2\pi n}{d} \tag{6}$$

This description of the field for periodic systems is known as Floquet spatial harmonic expansion, and each term n represents a Floquet Space Harmonic (FSH). Floquet's theorem is essentially an extension of the Fourier theorem for periodic functions.

Any function or field (e.g., $\psi(x)$) which repeats itself periodically except for a multiplicative exponential factor (e.g., e^{jkx0d}) can be expressed as an infinite set of modes (a_n). This formulation is adequate to describe the field in the vicinity of a planar 2D infinite periodic phased array with excitations of the unit cells which are uniform in amplitude and vary linearly in phase.

Background of Metamaterial

Metamaterials are composite structures that can be artificially engineered to exhibit unique electromagnetic properties. Metamaterial structures are made up from sub-wavelength unit elements, which control and manipulate the electromagnetic fields

according to the design geometry. This manipulation gives rise to different characteristics such as reversal of Doppler effects, reversal of Snell's law (negative refraction), and properties, which are not found in nature (Wu 1995; Vardaxoglou 1997). The subwavelength metallic structural features provide the necessary strong resonant interaction with the EM field, which also gives rise to dissipative losses, which needs to be taken care at high frequencies.

At high frequencies such as microwave-THz frequencies, metamaterials are generally constructed as periodic arrangement of the current-conducting elements that have appropriate capacitive and inductive characteristics. The feature size of the metamaterial structures is very much smaller than the wavelength at the frequency of operation so that the structure behaves as a homogenous material for an incident EM wave. Metamaterials have the EM properties that are not occurring naturally, and their EM properties are different from those of the constituent materials (permittivity and permeability). In the metamaterial structure, the structural dimensions such as width and periodicity are significantly smaller than the wavelength (λg) at operating frequency. Such structures are called as effective homogeneous EM structures. Hence, to build a medium where phenomena of refraction dominate over the scattering or diffraction, medium should follow the periodicity $p<\lambda g/4$. Hence, the medium seems to be homogeneous for the incident EM wave, and this effective medium is electromagnetically uniform in the direction of propagation (Vardaxoglou 1997). Thus, such material is known as metamaterial, and it is characterized under effective medium theory. The EM response of such material is due to the response through the effective medium.

The wave-matter interaction depends on the macroscopic properties of the material such as its constitutive parameters, electric permittivity (ε), and permeability (μ). The electrical permittivity of a material is defined as the ability of the material to store electric energy after the electric field is applied, and similarly, when a magnetic field is applied to a material, then the permeability of the material tells its ability to get magnetized in response to that magnetic field.

The free space permittivity and permeability is given as $\varepsilon_0 = 8.85 \times 10^{-12}$ F/m and $\mu_0 = 4\pi \times 10^{-7}$ H/m. The permittivity and permeability of all materials is defined with respect to the free space permittivity (ε_0) and permeability (μ_0), respectively. Figure 2 gives a classification of the material based on the material's constitutive parameters. The first quadrant shows the material for which both the permittivity and permeability are positive and hence is known as double positive materials. These are the natural materials existing in nature. The second quadrant shows the materials which have permittivity with negative values and positive permeability. Metals at optical frequencies behave as plasma with negative permittivity value. The third quadrant represents the materials having both the permittivity and permeability negative and hence is known as double negative materials. For such materials, the group and phase velocities are antiparallel to each other and hence are known as backward propagating waves. These materials show the characteristics which are not found in nature such as reversal of Doppler effect and reversal of Snell's law and are called as left-handed material or negative index materials or metamaterials. The fourth quadrant represents the materials having

negative permeability and positive value of permittivity. Ferrites shows such properties, and other gyrotropic materials show such properties. These materials also show the effect of Faraday Rotation.

For the realization of the metamaterials, first, thin wires with split ring resonators have been utilized. These two individual structures are combined to make a composite structure, which shows the unique feature of the simultaneous negative values of permittivity and permeability. However, the disadvantage of such assembly is it exhibits narrow bandwidth due to its resonant nature and high losses. Due to this reason, such composite structure does not have much utilization in most of the engineering applications. After that transmission line, metamaterials were developed, and they have been widely used for the development of the antennas and various microwave circuits and components (Vardaxoglou 1997). In such metamaterials, unit cell geometry plays an important role to determine the value of ε and μ and hence the behavior of the metamaterial structure. These structures may be resonant or nonresonant in nature depending on the specific applications. These metamaterials have subwavelength size of the unit cells, which control and manipulate the EM waves and hence found applications in the miniaturization of the many microwave structures.

Frequency Selective Surfaces and Their Evolution

Frequency selective surfaces are spatial filters used to obtain band-stop, band-pass, and absorption phenomena in response to an electromagnetic radiation. Frequency selective surfaces are long periodic structures or surfaces having identical elements, such as patches, slots, wire grids, and resonators, arranged in one-dimensional or two-dimensional array. When FSS is exposed to electromagnetic radiation, these surfaces produce scattered wave with prescribed frequency response. These FSSs have found several applications, which include antenna radomes, reflectors, absorbers, high-selective filters, and radar-absorbing materials as well as metamaterial, biomedical, and biochemical sensors (Rittenhouse 1786; Munk 1974; Sanz 2011).

Initially, single FSS is a single layer of thin surface of patterned conductor or resistive material on a planar or curved surface with a dielectric layer as a mechanical support. Multiple FSSs can have multiple layers, which are constructed using multiple dielectric layers and FSS sheets embedded in between. Another type of FSS, which is referred as circuit analogue structures, contains lumped elements such as resistor, capacitors, inductors, diodes, and transistors on the surfaces. These FSSs may offer deliberate losses for radar-absorbing material (Mittra et al. 1988).

History of FSS

History of the FSS dates back in 1786, when David Rittenhouse has observed that some color of light is being suppressed and is transmitted when a silk scarf

(handkerchief) is put in front of a lamp (Huang et al. 1994). This is the first filtering properties of the FSS that were discovered. The first experimental verification had come in 1919 when Marconi and Franklin had designed a parabolic reflector from the wire conductors arranged in a periodic manner. The first patent on periodic surfaces has been granted in 1974 to B. A. Munk (Langley and Parker 1990).

The features that are required for designing FSSs are high selectivity, stability in polarization and incidence angle, harmonic free, low-loss, and thickness. The problem with the design of FSS structures is unwanted couplings, roll-off, band edges, and larger separation between the resonating bands. These issues can be eliminated by using lumped components. However, this solution has the limitation due to the limited range of the lumped components values.

In order to obtain frequency filtering functions in the antenna, one can integrate the filters into the antenna-feeding system. However, this technique requires more space and is not favorable in many applications such as satellite communication. In such cases, spatial filters or FSSs may be used with the antenna, which block the incident wave to reach to the aperture of the antenna.

FSS plays a vital role in the military applications and satellite communications. The features required for FSS for such applications are multiband, reconfigurable, and with reduced size without replacing the feed (Schennum 1973; Schennum 1996; Romeu and Rahmat-Samii 2000; Sarabandi and Behdad 2007; Bayatpur and Sarabandi 2008; Kesawan 2019).

The technique to design multiband FSS includes fractal structures, perturbed structures, multiresonant cells-based structures, multilayered structures, and SIW (substrate-integrated waveguide) technology-based multiband FSS structures. Recently, liquid crystal materials have been used to develop reconfigurable FSS structures, as dielectric permittivity and loss of the liquid crystals are dependent on the applied voltages (Pelton and Munk 1974; Pelton and Munk 1979; Parker et al. 1981; Parker and Handy 1981; Hamdy and Parker 1982; Lagley and Drinkwater 1982; Langley and Parker 1982; Langley and Parker 1983; Tsao and Mittra 1984; Lam et al. 1988; Wu 1992; Huang et al. 1994; Lima et al. 1994; Schneider and Munk 1994; Sievenpiper et al. 1999; Broas et al. 2001; Yeo and Mittra 2001; Anderson et al. 2003; Bao et al. 2006; Xiong et al. 2013).

FSS can also be used in curved surfaces because of the small sizes of the unit-resonant elements and used for designing leaky wave antennas. Multilayered FSS, which is made up of layers with multiple-resonant cells with multiperiodicity, is designed to get sharp roll-offs, high selectivity with reduced grating lobes in the operating reason. However, in multilayered structures, one needs to take care of the mutual coupling among different layers during the design. Multilayered structures are also constructed by printing metallic patches on one side of the dielectric substrate and metallic wires on the other side. Another approach to get the sharp roll-offs and high selectivity is the cascading technique. Cascading technique is the technique to combine multiple FSS layers. However, cascading gives thicker and bulky structures, and its frequency response may also vary with the angle of incidence.

In the single-layered FSS designs, the number of identical resonant elements is increased to get the higher order FSS. The different resonant elements may be utilized to obtain multiband response.

However, more FSS structures with single-layer design along with multiple band frequency responses are designed and analyzed nowadays. The passive resonant elements are being used for further miniaturization of the FSSs. The design of metamaterial FSS with multiband capability with controlled propagation direction and polarization of the EM waves at high-frequency regimes such as millimeter wave and THz frequency ranges is still going on.

Traditional Frequency Selective Surfaces Design Elements

A frequency selective surface (FSS) is a periodic array of surface elements printed on the dielectric substrate, which selectively controls the frequency characteristics of the incident electromagnetic (EM) wave. The response depends on the shape of the structure as well as the polarization and incident angle of the incident wave. For the numerical analysis, these arrays are assumed as infinitely long, which reduces the whole problem of analysis to the frequency response of the single element of the periodic array. Various element geometries used for the FSS design have been discussed in subsections below.

Geometries of the Element

There are several elements used to analyze FSS, which are broadly categorized in patch- and aperture-type elements. First, square patch and its complementary slot-type planar periodic arrays were considered for the FSS. Figure 3a shows the structure consisting of periodic array of metallic patches, which behaves as a capacitive surface and thus shows low-pass characteristics as shown in Fig. 3c. When a plane wave is incident on the surface, this surface transmits the low frequency of the wave while reflecting the higher frequencies. Figure 3b shows the array of slot type elements, which is the complementary structure of Fig. 3a. This surface behaves as an inductive surface and hence provides high-pass response as shown in Fig. 3d. When these two surfaces are put together, a band-pass response is produced (Schennum 1973).

Depending on the applications of FSS, different FSS elements have been introduced for band-pass and band-stop applications and can be chosen to fulfill the demands. These demands include the dependence on the incidence angle of the incoming wave, level of cross-polarization, bandwidth, and level of band separation.

A variety of FSS elements have been introduced over the years for the band-pass and band-stop applications as shown in Fig. 4. These FSS elements are categorized in different groups.

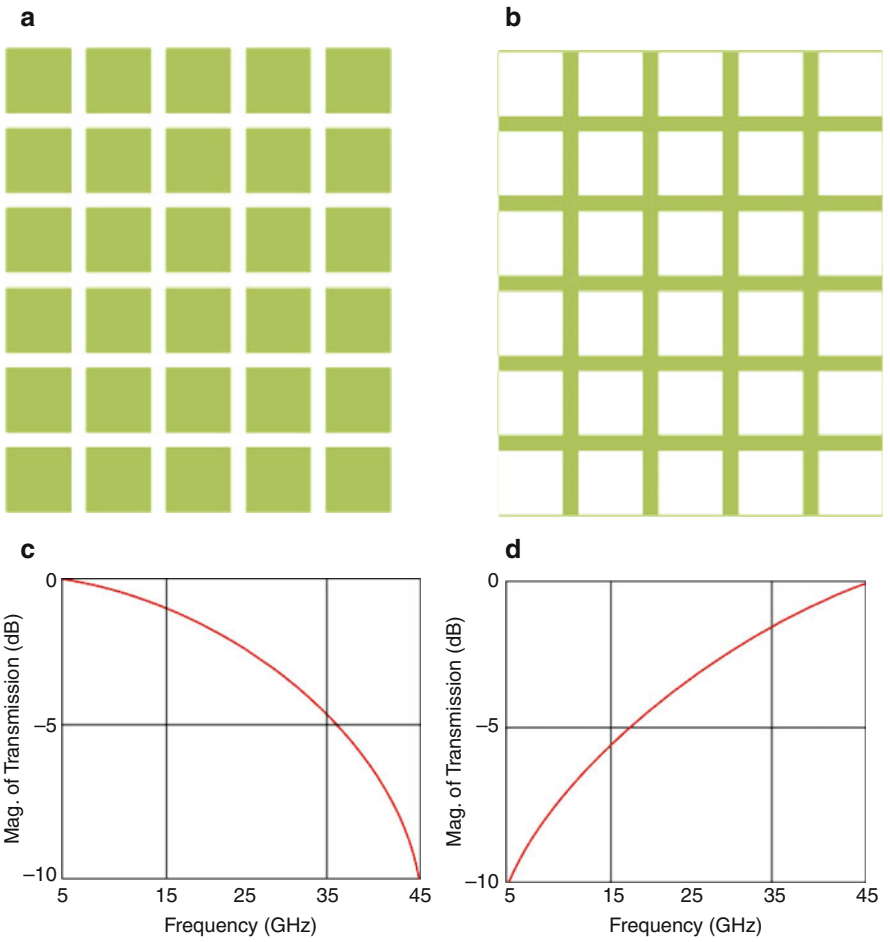

Fig. 3 (**a**) Patch type element array; (**b**) slot type element array; and (**c, d**) transmission response of the Fig. (**a**) and (**b**), under transverse electromagnetic (TEM) plane wave excitation, respectively

Selection of a particular group or specific element types depends on the applications of the FSS. However, some of the features should always be desirable in most of the applications, which are the following:

1. Less sensitivity or low dependence of the FSS on the angle of incidence of the exciting plane wave
2. Bandwidth
3. Miniaturization
4. Harmonic free

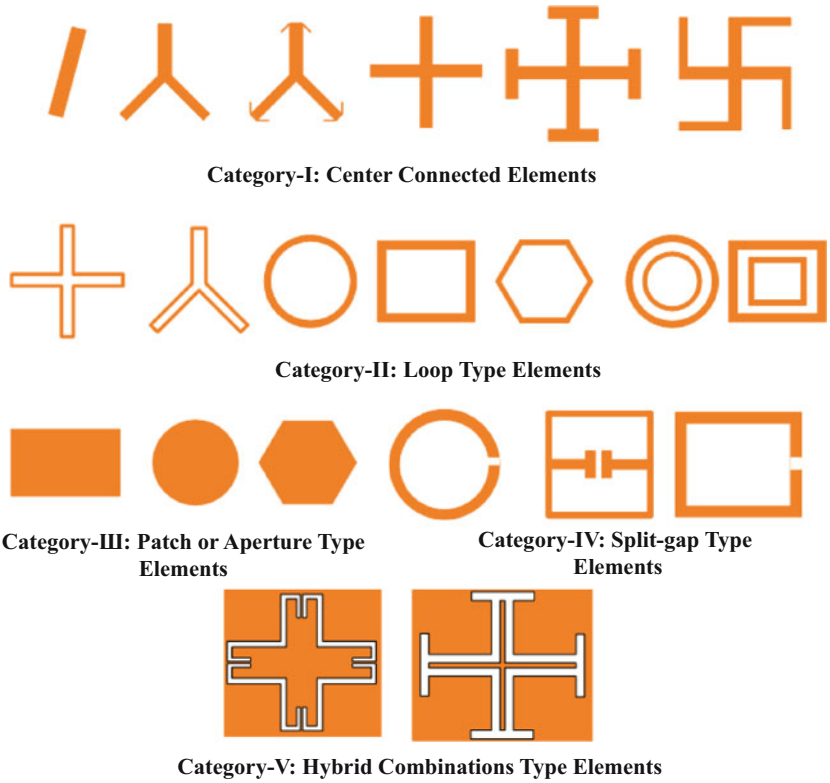

Fig. 4 Various types of FSS elements

Category-I: Center-connected elements, which include dipole, three and four-legged elements, Jerusalem cross, square spirals, etc. One always takes care of choosing the specific shape to reduce the inter-element spacing (miniaturized FSS) along with the enhanced bandwidth. By reducing the interelement spacing, one can tightly pack the elements and miniaturize the overall size of the FSS. In comparison with the tripole array, Jerusalem cross-based FSSs provide better stability as the angle of the incidence, bandwidth, and harmonics free.

Category-II: Consists of the loop type elements, which include square, circular, and hexagonal loops, multiple ring loops, etc.; in these elements, hexagon-shaped element provides better performance and response characteristics.

Category-III consists of solid patch type elements such as (a) array of metallic patches having shapes such as circular disk, square, rectangular, etc. which acts as reflecting arrays, (b) array of slots on metallic surface with shapes like rectangle, square, circle, etc. These slot type structures behave transparently. These are the

traditional design of FSS in which the patches have the dimensions of the order of X/2. Also, these patches suffer from angle of incidence dependence and early onset of grating lobes, and hence these are generally not suitable for filter design. Using the patches with metamaterial can provide a better response.

Category-IV consists of split gap type elements, which are widely utilized for metamaterial designs.

Category-V includes hybrid combinations of different elements, which can be from any group elements. One can create infinite number of elements using the combination.

Element Dimensions

As FSS is a periodic structure, which is designed using resonant elements arranged in periodic fashion over a dielectric substrate. However, traditional FSS unit cells have the dimensions of multiple wavelengths at frequency of operation. FSSs that are made up of dipoles (Parker and Handy 1981) and square loops (Langley and Parker 1982) have the length of half of the wavelength.

Moreover, the performance characteristics of the FSS depend on the shape of the unit elements and the way these unit elements are arranged in the periodic fashion. The frequency response also depends on the characteristics of the dielectric substrate used such as loss tangent and height.

Application of FSS

Initially FSSs were introduced as spatial filters; however, they have been employed in various other applications such as radomes, filters, absorbers, polarizer, artificial magnetic surfaces, high-impedance surfaces, high-gain antennas, multiband and reconfigurable FSS-based absorbers and filtering characteristics, electromagnetic shielding, beam-switching, and beam shaping (Jackson and Alexopoulos 1985; Yang and Alexopoulos 1987; Ghodgaonka et al. 1989; Li et al. 1995; Liu et al. 1995; Liu et al. 1997; Bayatpur 2009; Brito 2010; Fallahi 2010; Edalati and Sarabandi 2014; Costa et al. 2016; Ebrahimi 2016; Hussein 2018; Lv et al. 2019; Mahmoodi 2020). Radomes in antenna reduce the radar cross section (RCS) outside of the operating frequency, where the radome acts as a band-pass filter. FSSs have been used in reflector antennas for communication data links. A high-gain FSS reflector antenna was proposed in Mahmoodi (2020). As FSSs have reflective characteristics with a certain amplitude and phase, these can be used in the design of impedance surfaces (Li et al. 1995). Phase-shifting property of FSS has been also shown experimentally in Liu et al. (1995), which has been utilized in the development of reactive impedance surfaces. FSS structures are also used to design high-impedance surfaces, artificial magnetic conductor (AMC), and electromagnetic band-gap structures. In Li et al. (1995), artificial magnetic conductor is used to suppress the surface wave inside the substrate. FSS has one of the prominent

applications in the development of the microwave/THz absorbers. These absorbers exhibit the absorption at RF/THz frequency. These single and multiband absorbers are reported in literature (Costa et al. 2016; Lv et al. 2019).

FSS-Based Absorber Offers the Following Advantages

1. Structures are ultrathin small in the order of λ/20.
2. Z-direction thickness is also small.
3. Scalable: Change the unit cell dimension and material: frequency of operation can be changed.
4. Fabricated using PCB technology.
5. Low cost.

The absorptivity, A, is given as Eq. (7).

$$A = 1 - |S_{11}|^2 - |S_{21}|^2 \tag{7}$$

When both the reflected power ($|S_{11}|^2$) and transmitted power ($|S_{21}|^2$) are simultaneously minimized, then the absorptivity (A) is found to be maximum.

In most of the FSS absorbers' design, metal sheet is placed on the other side of the dielectric substrate. This ensures zero transmission (i.e., $|S_{21}| = 0$). This metal sheet will block the transmission, and only reflection will be there from the metal sheet. Now, Eq. (7) reduces to Eq. (8).

$$A = 1 - |S_{11}|^2 \tag{8}$$

Now, by reducing the reflections ($|S_{11}| \rightarrow 0$), one can obtain maximum absorption. So now the question is how to reduce the reflection coefficients to zero? One of the approaches is to use the impedance matching with the free space impedance. This needs to design the unit element in such a way that the input impedance is matched with free space impedance.

FSS-Based Filter

FSS is essentially a filtering structure, which is used to transmit one band of frequency and reflect all other frequency outside the band. FSS-based filter geometry is used in several applications where high selectivity (or large quality factor) in frequency characteristics is required. (i.e., very small bandwidth). There are many FSS-filtering structures reported in the literature such as narrow band, broadband, multiband, and tunable structures, based on the requirement and applications. Multilayered multiorder FSS structures are used to obtain high selectivity and flat band response. Aperture-coupled concept is used, where metallic patches are placed on the top and bottom side of the dielectric and slots are made in the dielectric.

FSS-based filter structure minimizes either S_{11} or S_{21} whereas FSS-based absorber structure minimizes both the S_{11} and S_{21}.

FSS for Sensing Applications

Recently, FSS structures are being used widely for many sensing applications. As FSS frequency response is affected by the changes in the dielectric substrate parameters such as permittivity, los tangent, and thickness as well as geometry shape; FSS has found a potential application in the domain of wireless sensing applications. Various research have been reported for monitoring the structural health-like crack detection, strain and temperature sensing, inspection of multilayer structure, normal and shear strain sensing, etc. (Costa et al. 2016). Various performance metrices-resonant frequency, depth of resonance, and quality factor are used to quantify the effect of substrate parameters on sensor performance.

The other applications of FSS are in electromagnetic shielding, reconfigurable FSS in beam-switching antennas, and high-impedance surfaces for suppression of surface waves and gain enhancement.

However, the traditional FSS elements suffer from the high sensitivity of the angle of incidence, larger inter-element spacing, and hence offer lower bandwidth and harmonics. These affects can be minimized by using subwavelength elements so that they can be tightly packed, and their EM response can be manipulated according to the applications. This can be done using metamaterial-based subwavelength unit cell elements, which is discussed in the next section.

Metamaterial Frequency Selective Surface

Many applications require lower sensitivity in terms of angle of incidence of EM wave and to maintain uniform phase front. To fulfill these demands, smaller size of the FSS is required. In traditional designs, the overall frequency response of the FSS is based on the mutual interactions among the unit cells, which requires FSS size to be large to get the specified response. However, in order to get low sensitivity of the FSS frequency response for the incident wave having oblique angles of incidence, the unit cell size needs to be miniaturized. The subwavelength features are particularly important when the FSS structure is kept near the antenna or any radiation source. Therefore, metamaterials-based frequency selective surfaces have been formed based on miniaturized subwavelength element sizes. These types of FSS unit cells have the subwavelength dimensions such as $\lambda/10$ or $\lambda/20$, and hence the FSS surface behaves as a homogenous medium for the incident EM waves. These type of FSS allows excellent performance over the conventional FSS structures. In metamaterial FSS structures, nonresonant unit cells with subwavelength dimensions have been utilized instead of using resonant elements. These miniaturized unit cell elements behave as lumped capacitors or inductors, and these lumped elements are arranged in a manner that they couple to the incident electromagnetic wave. In doing

so, the electrical size of the FSS unit cells decreases, and thus the size of overall structure decreases (Mahmoodi 2020).

Metamaterial FSSs usually employ planar periodic metallic arrays or dielectric arrays in two-dimensional space. The frequency response of the FSS structure is determined by the shape and geometry of the unit cell element considering the infinite size of the FSS surface. The metamaterial FSSs have found their application in multiband reflectors, radomes for antennas in airborne applications, multiband and broadband polarizers, electromagnetic interference (EMI) shielding, diplexers, circuit analog absorbers, beam steering and beam switching antenna arrays, and high-impedance surfaces (HISs), artificial magnetic conductors, or electromagnetic bandgap (EBG)-resonating elements array to increase the directivity and gain of the antennas.

A compact FSS-based filter geometry will serve two purposes:

1. The design will occupy lesser space (if we make the unit cell compact), and thereby a greater number of unit cells can be fitted.
2. The grating lobes will appear delayed, and angular stability will be enhanced (grating lobes are the type of sidelobes coming in the picture which reduces the overall response).

In order to remove or minimize the grating lobes, one option is to use fractal geometry. By doing so, the physical dimension is the same, but one may increase the order by fractal, and hence frequency is decreasing, and thus electrical size is reduced. As frequency is reduced, wavelength is increased. And as electrical size is reduced, so one can obtain miniaturization of the structure. FSS designs based on fractal geometry provide multiband response (Bao et al. 2006). Another way to minimize the grating lobes is to use the convolution geometry where meander lines concept is utilized. The values of the inductance and capacitance are increased in meander line, and hence frequency is reduced, and one obtains miniaturized structure.

Apart from that, reconfigurable metamaterial FSSs have also been reported. Varactor diodes are used as a tuning element. These varactors have been incorporated in series with the unit element. DC power supplies are used to bias the varactors and to change the voltage across the diode. By using the varactors, one can tune the center frequency, bandwidth, as well as the transmission zeros on either side of the band in the FSS structure (Fallahi 2010).

The metamaterial-based FSSs were designed to obtain the high-frequency selectivity, greater out of band rejection, spurious free responses, and low sensitivity to angle of incidence and polarization without increasing any fabrication complexities. FSS structures show reflective characteristics with a particular phase and amplitude that allow frequency changes in a medium and are not restricted to a fixed frequency response. This type of metamaterial FSS suppresses the surface waves when used as a ground plane in the antennas and provided better radiation patterns. FSSs can also be used as a superstrate to an antenna to enhance performance such as directivity and gain. FSS superstrates are easily fabricated with the use of planar technology such as

microstrip whose thickness is much less as compared to the dielectric superstrates. These FSSs have homogeneous surfaces which also suppresse the grating lobes, and hence a high-gain antenna can be developed (Edalati and Sarabandi 2014).

Design and Analysis of Metamaterial FSS

FSS is a large periodic structure, which has frequency control characteristics. So, the question is how to analyze such a large periodic structure. In order to analyze the FSS characteristics, one unit cell is designed and analyzed.

In this section, design, and analysis, of the metamaterial FSS is demonstrated using the HFSS and CST microwave studio. A step-to-step procedure for designing FSS is provided for the beginner.

Design and Simulation Setup in HFSS

Figure 5 shows the unit element of the FSS along with the boundary conditions and excitation. First of all, a unit cell is designed according to the specification (band-pass or band-stop) in the modeler of the HFSS. After that, an air box is created as shown in Fig. 5. In order to define the periodic boundary condition, master-slave boundary conditions have been defined in both x and y directions. Floquet port has

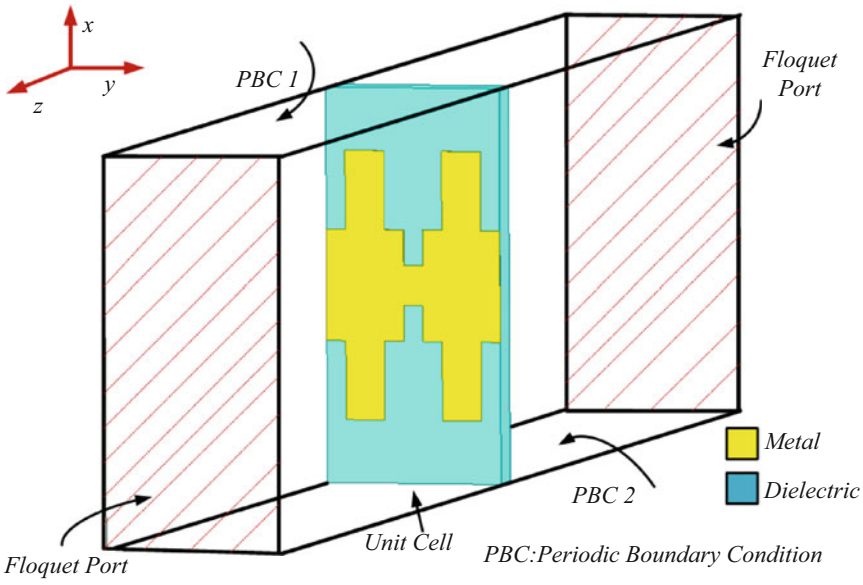

Fig. 5 Unit element of the FSS along with the boundary and excitation

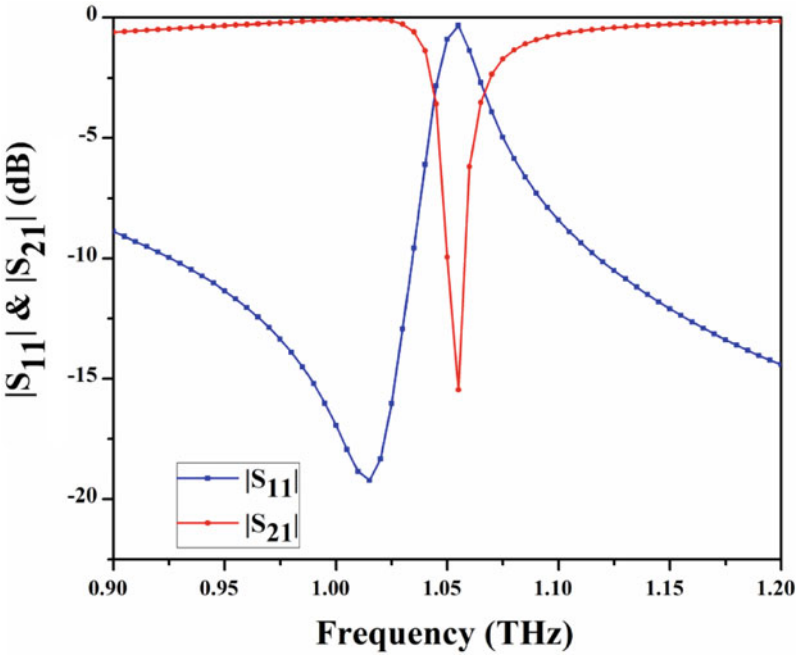

Fig. 6 S-parameter response of the designed unit cell of Fig. 5

been used to excite the unit cell resonator. The metal is taken as perfect electric conductor, and the dielectric substrate is Rogers 5880 having the dielectric constant of 2.2 and loss tangent of 0.0009, with a height of 20 micrometer.

After setting the solution setup, this FSS unit cell is simulated and its results are analyzed. The same unit cell has been simulated according to the abovementioned procedure, and its reflection and transmission coefficients have been recorded and provided in Fig. 6.

As FSS structures are the large periodic structure, unit cell elements are periodically arranged on a dielectric substrate. In order to perform numerical analysis, the FSS array sizes are considered to be infinite. The assumption reduces the whole array analysis into calculating the frequency response of a single unit cell element in the array given the periodic nature of the FSSs.

Figure 6 depicts the S-parameter response of the designed grooved stepped impedance resonator type unit cell as shown in Fig. 5. This FSS unit cell resonates at frequency of 1.05 THz and provides band-stop behavior. This FSS unit cell gives narrow band response. In order to broadband response, unit cell shown in Fig. 7 is utilized. This unit cell is a combination of loop and slot apertures. Figure 8 presents the S-parameter response of the FSS unit cell as shown in Fig. 7. Therefore, as discussed above, the loop and slots provide better bandwidth and flat response in the broadband as compared to patch type elements (Fig. 9).

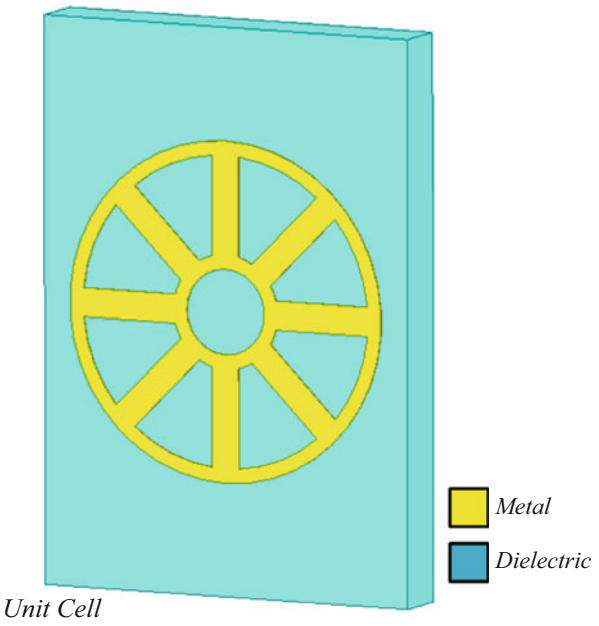

Fig. 7 Unit element of the FSS

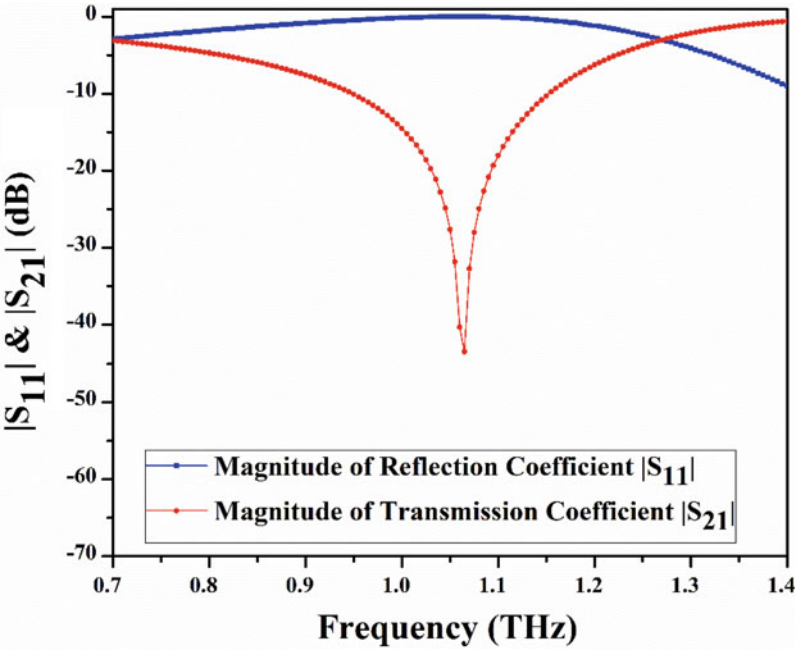

Fig. 8 S-parameter response of the designed unit cell of Fig. 7

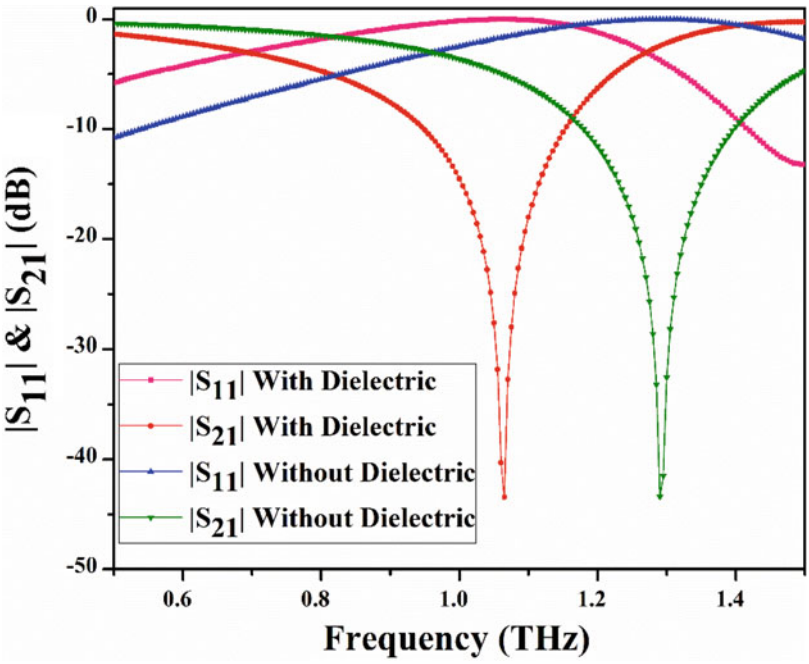

Fig. 9 S-parameter response of the designed unit cell of Fig. 7 with and without dielectric slab

Design and Simulation Setup in CST

The same FSS unit cell can also be analyzed in CST microwave studio. The only difference here is the boundary condition. Here, unit cell boundary conditions are employed in x and y directions. For the excitation, floquet port is used.

After designing and analyzing the response of the single unit cell, the same unit cell is used to make an array, which has the size in the multiple of wavelength, i.e., 4 λ, 5 λ, 10 λ, etc. These structures' size should be in the order of 5 $\lambda \times 5\ \lambda$, where λ is the wavelength corresponding to the frequency of operation. Then the structure is almost behaving as the infinite structure. In case of lower size (1 $\lambda \times 1\ \lambda$), the simulated and measured results will be different.

Fabrication and Measurement Method

After analyzing an array, the structure needs to be fabricated for the experimental verification. The FSS array can be fabricated with printed electronics technology such as lithography, ink-jet printing, direct deposition, aerosol jet printing, and nowadays 3D printing methods. After fabricating the structure, the FSS is experimentally characterized in Anechoic chamber.

Measurement Steps

1. In anechoic chamber, the two antennas are placed in near proximity (very close to each other); we put some pyramidal absorber in between them to avoid any coupling between one antenna to another antenna.
2. One antenna will be used as transmitter antenna, and another will be used as receiving antenna.
3. Now background noise is checked and recorded. While measuring the background noise, no FSS structure will present there. Only the signal is transmitted in free space, and it is transmitted and reflected from the anechoic chamber, and then the signal is received by the receiving antenna. This background will be very small based on the specification of the anechoic chamber.
4. After that, a metallic plate is placed (mostly copper) in front of the antennas and transmitted power ($|S21|^2$) has been measured. (Transmitted power will be reflected from copper plate, and it is received by the receiving antenna and recorded in the VNA).
5. Now the metallic plate is removed and replaced with the fabricated FSS structure. Again, signal is transmitted through antenna, and transmitted power ($|S21|^2$) is recorded coming from the fabricated FSS.
6. After getting the transmitted power from both the metallic plate and FSS structure, the transmitted power from the copper plate is subtracted from the power obtained from the FSS sample, which gives reflection. By doing so, one may remove the path losses such as scattering loss and diffraction loss and get the actual results. This subtraction removes the diffraction effect due to the finite size of the FSS sample. There may be small deviation in the results that may be due to the fabrication tolerance and etching. One does not need to do such subtraction in case of simulation as we model the lossless free space in the simulation, so there are no such effects in the simulation model.

Conclusion

This chapter provides an overview of the Metamaterial FSS for THz applications. Here we focus on the various aspects of the FSS such as design, applications and how to setup the simulation and do the measurement for the designed FSS etc.

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Metamaterial for RFID and Sensor Applications

23

Jamal Zaid and Tayeb A. Denidni

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Abstract

Radio frequency identification (RFID) is a rapidly evolving technology that enables wireless identification and tracking with simpler devices, used to mark objects or people at one end, called tags, and more complex devices on the other end, called the reader. RFID is a new technology and is one of the fastest growing segments of the modern automatic identification and data capture (AIDC)

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industry. RFID is currently used in hundreds, if not thousands, of applications. RFID is revolutionizing supply chain management by replacing barcodes, the most important object tracking system, and is rapidly becoming a cost-effective technology. However, developing a tag that can cover all controlled UHF-RFID tapes while providing consistent read performance is an important challenge. Cost savings are a special issue when implementing microwave RFID systems, as they usually use active tags that need to reduce energy consumption. Another challenge is the size of the RFID antenna, where metamaterials concepts can be used for the miniaturization of RFID devices. In this chapter, we describe about RFID technology and how magneto-dielectric substrate (MDS), which operates as a split ring resonator (SRR)-like loop, used for the miniaturization of various RFID antennas.

Keywords

RFID · Multi-port patch antenna · RFID tag based-sensor · Antenna miniaturization · Magneto-dielectric substrate · Reading range · Minimum required power

Introduction

RFID Technology

Radio frequency identification (RFID) is a rapidly evolving technology that enables radio identification and tracking. Today there are many applications such as theft prevention of car, smooth toll collection, building access, restricted access communities, vehicle access control to business campuses and airports, providing lift access, library product tracking, asset identification, retail and supply chain management, animal tracking, and other functions using RFID systems (Finkenzeller 2003; Hunt et al. 2007). This section provides an overview of RFID technology, provides a brief description of key frequency bands and a brief history of RFID.

RFID usually has a simple device called a tag or responder at one end and a more complex device called a reader at the other end. The tag contains an application-specific integrated circuit chip (ASIC) with memory to store the data and an antenna. Sometimes a matching network can be included between IC and the antenna to ensure appropriate impedance matching. The reader contains an antenna, an electronic RF module that is responsible for tag communication, and a control electronic module that communicates with a host computer (or controller), usually connected to the reader, to process information from readers. Figure 1 shows a diagram of a typical RFID system. RFID systems can be read-only (data flow in one direction from tag to reader) or read-write (two-way communication). Tags can be tagged either by battery (active tag) or by a radio signal (passive tag) transmitted by the reader.

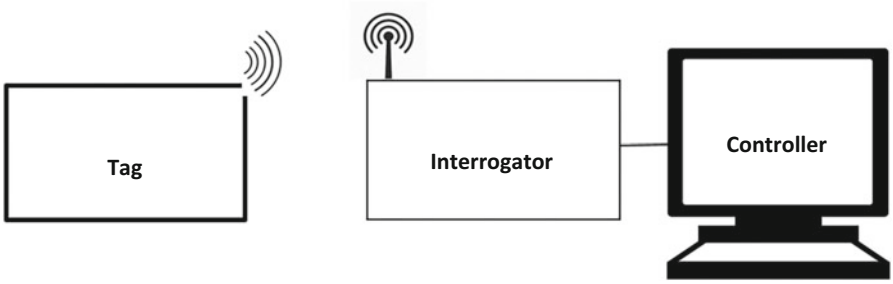

Fig. 1 Schematic of RFID system

History of RFID

The first milestone on the basic foundation of RFID passive technology was published by Harry Stockman in 1948 and is called communication by means of reflected power (Stockman 1948). Stockman noted in Ref. (Stockman 1948) that a lot of research and development is required before solving the remaining major problems with reflected power communication and before exploring useful applications. It actually took 30 years of research before the first RFID transponders were developed. The important events related to RFID history are listed below (Landt 2005).

In the 1950s, researchers studied many RFID-related technologies using advances in radio frequency communication and radar theory made in the 1930–1940. In the 1950s, the US military commenced operating an early type of long-range identification system called IFF (identification, friend or foe) designed to differentiate between allied and enemy aircraft. The merchandising activity of RFID technology began mainly in the 1960s due to research into RF: Harrington's "Fields Measurements Using Active Scatterers" (Harrington 1963) and "Theory of Loaded Scatterers" (Harrington 1964) published in 1963 and 1964. During these decades, electronic article surveillance (EAS) tools for theft prevention and protection were developed. Such systems were one-bit systems that could only detect the presence of an object and could not identify it.

The RFID explosion occurred in the 1970s as more and more companies, academics, and government laboratories started working on RFID. During these decades, RFID technology has been used for animal tracking, vehicle tracking, and factory automation. Significant advances were made in a 1975 article by Alfred Kelle, Stephen Depp, and Robert Freiman entitled "Short-Range Radio-Telemetry for Electronic Identification using Modulated Backscatter" (Koelle et al. 1975). In 1978, the passive microwave transponder was introduced. In the 1980s, retail RFID systems began to grow in many parts of the world. Examples such as cattle management, employee punching systems, keyless access, and transport applications that emerged later this decade. The world's first customs clearance application appeared in Europe in 1987, followed by in the USA in 1989. With the development of PCs in the same period, RFID applications have spread rapidly, enabling more

convenient and cost-effective data processing. Tags started using custom CMOS ICs and discrete components. This allowed to further reduce tag size and improve functionality. As a result, the size of the antenna has become important in determining the tag size.

Also, the 1990s became an important decade as RFID began to penetrate common businesses and technologies. RFID systems have begun to be used in customs, access control, and other retail applications. In the mid-1990s, toll-based RFID systems operated at highway speeds, allowing vehicles to pass through customs collection points without being blocked by barriers. The success of electronically toll collection has led to significant advances, such as the first use of tags in various business segments. This allows a single tag to be used for various applications such as electronic tolls, parking access, and closed public access. New RFID applications have also been created for fuel distribution, ski passes, and vehicle systems. For the initial time, microwave RFID tags containing only one integrated circuit were created (previously their capacity was limited to inductively coupled RFID transponders). This, along with the increasing number of computers and Internet, has given the RFID industry a unique problem with expensive tags that must be overcome to develop commercially viable systems. The Auto-ID Center was established at MIT in 1999, with manufacturers, researchers, and RFID users working with other standards organizations such as the International Standards Organization (ISO) and the European Conference of Postal and Telecommunications Administrations (CEPT).

In the beginning of 2000s, microwave trays were made with separate labels with two elements: a simple IC and an antenna. Later then, the size of the spike is restrained by the size of the antenna. In 2003, Auto-ID merged with the Electronic Product Code (EPC) to take over the standard issue of supply chain applications. In the early 2000s, ISO was also very active in developing standards for a variety of applications. All of these organizations are currently working on RFID technology standards.

Frequency Ranges and Applications

RFID systems operate in a variety of frequency bands, including the 125 kHz to 5.8 GHz frequencies in the microwave range.

Low Frequency (LF)

These systems operate from 125 kHz to 134.2 kHz and can be used in the USA, Europe, and Japan. Due to the electromagnetic properties of these frequencies, the LF tags can be read even when attached to objects containing water, animal tissue, metal, liquids, and wood. However, it is suitable only for contactless applications as it can poll at very close distances of a few centimeters (LF tags are usually passive). Antennas for LF tags usually consist of a copper coil wound around a ferrite core hundreds of times. Due to these features of LF tags, they are used in some

applications such as asset tracking, animal identification, vehicle immobilizer, access control, medical, and various POS applications. Since the early 1980s, LF tags have been widely used for animal tracking.

High Frequency (HF)

This band operates with a center frequency of 13.56 MHz and is available in the USA, Europe, and Japan with very similar power levels. HF tags are passive and work the same way as LF tags. However, HF tags have a wider reading range than LF tags and can be read up to 0.5 m away. The tag contains a resonant LC antenna made by rotating a conductive material such as copper, aluminum, or silver several times (~5–20 turns) in a flat spiral, usually connected to a capacitor. So, RF tags are usually very thin (as a paper). Like LF tags, HF tags are easy to read when attached to objects that contain water, cloth, metal, wood, and liquids. However, nearby metal objects will affect performance. It is difficult for tagging metal objects, because it significantly suppresses the magnetic field perpendicular to the object required to excite RF tags attached to the metal surface. Applications include smart cards, credit cards, NFC, library bookmarks, airline tagging, and asset tracking.

Ultrahigh Frequency (UHF)

The UHF systems works in the range of 433 MHz (active tag) and 840 MHz to 960 MHz (passive and semi-passive tags). There exist important differences among the rules of many countries around the world, such as the USA, China, Europe, and Japan. UHF tags have a maximum reading range of approximately 15 m. In contrary to LF and HF tags, all UHF protocols have some kind of anti-collision feature, allowing to read multiple tags at once. UHF tag antennas are mainly based on dipole antennas and are made of copper, aluminum, or silver deposited on a substrate. At 900 MHz, the length of the resonant half-wave dipole antenna is about 17 cm. However, with proper design techniques, the overall size of the antenna can be reduced. UHF antennas are easy to produce and can be thin with a flat design like HF tags. UHF tags offer more memory and better read range compared to LF and HF tags. As we know, the performance of UHF tags changes when applied to other objects. This can usually be solved by taking into account where the tag will be placed in at the design stage. However, objects containing metal and water can interfere with the visible reading area due to energy absorption and decomposition. This can be solved by creating tags specifically for labelling metal or liquid content. However, tags that work well on metal objects are usually too expensive, oversized, and heavy in some applications. The reason metal objects reduce the efficiency of conventional tags (not specifically designed for metal) is that most UHF tags must be excited by a directional electric field, tangentially oriented to the metal surface, which cancels the electric field. In the case of water (or liquid), short UHF waves are usually strongly absorbed by the liquid. UHF RFID applications include supply chain management, retail product management, and parking access management.

Microwave

Systems operating at industrial, scientific, and medical (ISM) mid-band frequencies of 2.45 GHz and 5.8 GHz come into this category. Passive, semi-passive, and active tags occur in this frequency range and in most regions. Passive tag reading range is similar to UHF tag interval (up to about 15 m), but active tags can be read at intervals of up to 100 m or more. If the tag is near water or metal, like a UHF tag, performance may be degraded.

Since anti-collision protocols are available, allowing multiple tags to be read simultaneously. However, microwave devices that use the same ISM band, such as cordless phones and microwave ovens, generate RF noise. Applications include fleet management for customs, vehicle identification, and access control for gated centers, corporate cities, real-time location systems (RTLS), and hospitals and airports.

The most used physical coupling techniques in low frequency and high frequency bands are based on inductive coupling, but there are also several capacitive coupling systems. The significant reason is that only objects and metals with high permeability affect the magnetic field, and these objects are rare in everyday life. Additionally, capacitive coupling systems are influenced by objects with high dielectric permittivity and losses. In contrary, UHF and microwave systems are connected using electromagnetic fields and sometimes surface acoustic waves.

UHF-RFID Systems

In particular, ultrahigh frequency (UHF) RFID systems are very attractive because they can cater higher read range, faster signal processing, secure data transmission, and more memory as compared to RFID regulatory systems that use low frequency (LF) and high frequency (HF) bands (Finkenzeller 2003). Despite microwave RFID systems offer similar realization to UHF systems, rules and standards for this frequency band are available. Since it operates in the ISM band, it is particularly susceptible to interference from other microwave devices that share this frequency band.

Frequency Bands and Operating Principle

The regulated UHF RFID bands depend on different places of the world and cover frequencies from 840 MHz to 960 MHz. The UHF RFID frequency band operates worldwide at 840–845 MHz in China, 866–869 MHz in Europe, 902–928 MHz in the USA, and 950–956 MHz in Japan. As a result, developing UHF tags that can cover all regulated bands (e.g., global band tags) has become an important challenge.

Most passive UHF RFID systems benefit the modulated backscatter principle. In this category of RFID system, the reader transmits the modulated signal received by the tag antenna at CW intervals. The chip converts the unmodulated signal to DC and sends the information back by changing the complex RF input impedance Z_{chip}. This impedance is usually switched between two different states, such as a coupled coupling, and another impedance that effectively modulates the backscattered signal (Seshagiri Rao et al. 2005).

Read Range

The most significant tag parameter that indicates the realization of an RFID system is the read range. It is defined as the maximum distance the reader can read information from the tag. Since the sensitivity of the reader is usually higher than that of the tag, the read range is determined by the response threshold of the tag. The read range depends on the orientation of the tag, the material the tag is attached to, and the environment (Seshagiri Rao et al. 2005). The reading area r can be calculated by the Friis formula for free space.

$$r = \frac{\lambda}{4\pi}\sqrt{\frac{EIRP\, G_t P_t}{P_c}} \tag{1}$$

where λ is the wavelength and EIRP is the equivalent isotropic radiated power, determined by region regulations (EIRP = 3.3 W in Europe and EIRP = 4 W in the USA). Pc is the minimum threshold power needed to activate the RFID chip, P_t is the power transmission coefficient between the chip and the antenna, and G_t is the gain of the receiving tag antenna. It is important to emphasize that Eq. (1) gives the reader the maximum distance in free space to detect backscattered signals from a given tag. The reader antenna and tag antenna are oriented together in the direction of maximum gain and their polarization is the same.

There are several ways to measure the reading range of an RFID tag. This can be done with the most cost-effective method, using a commercial RFID reader (Fig. 2). The tag is moved away from the reader as far as the tag ceases responding. The read range curve as a function of frequency is achieved by redoing the same procedure, changing the frequency within a given RFID frequency range. The fundamental issue with this approach is that commercial readers are designed to work only in certain frequency bands (e.g., Europe). This will reduce the reading for the read range to a limited frequency range. However, as long as there are not multiple readers, they can all operate in different UHF RFID frequency ranges. However, it cannot get a continuous curve over the range as a function of frequency due to the frequency band gaps between the

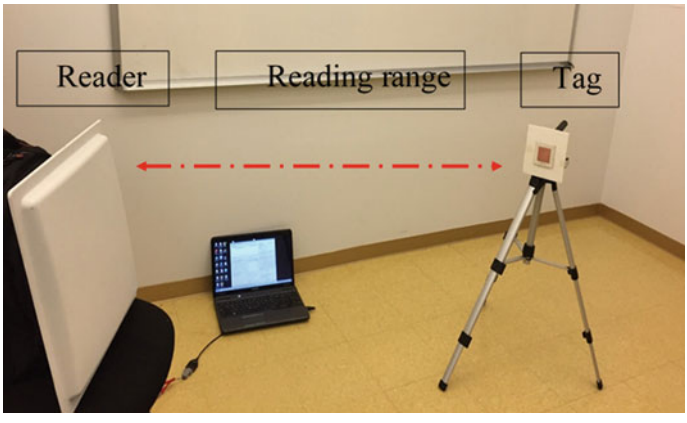

Fig. 2 Measured reading range

adjustable bands. To solve this issue, instead of an interrogator, we can use an arbitrary function generator that can generate an RFID frame and attach it to a specific antenna.

Magneto-Dielectric Substrate-Based Microstrip Antenna for RFID Applications

Microstrip patch antenna miniaturization has been an attractive research topic for the past 20 years. This miniaturization presents complex challenges due to trade-offs between antenna size and performance related to matching bandwidth, radiation efficiency, and gain. Numerous methods have been considered to decrease the physical size of microstrip antennas (Chakraborty et al. 2014; Reddy and Vakula 2015; Khan et al. 2015; Colburn and Rahmat-Samii 1999; Mookiah and Dandekar 2009; Mosallaei and Sarabandi 2007; Pinsakul et al. 2012; Ikonen et al. 2006). The abovementioned miniaturization techniques fall into two groups. The first category includes changes in the topology of the radiating element (Reddy and Vakula 2015; Khan et al. 2015). For example, incrementing the electrical length of the current path. These methods are advised easy to manufacture and can be used for industrial utilization. However, as reported by Reddy et al. (Reddy and Vakula 2015), it has several drawbacks, such as degradation of the radiation pattern of antenna due to changes in the model of the radiating elements. In addition, the use of these technologies in MIMO systems can enhance the mutual coupling between adjoining radiating elements due to surface waves generation (Khan et al. 2015). These approaches typically diminish the operating impedance bandwidth of the antenna as a result of the new resonances generated along the new elongated current path.

In the second category, we can change the overall specifications of the host substrate (Colburn and Rahmat-Samii 1999; Mookiah and Dandekar 2009; Mosallaei and Sarabandi 2007; Pinsakul et al. 2012; Ikonen et al. 2006). One method uses substrates with high electrical permeability, as shown by Colburn et al. (Colburn and Rahmat-Samii 1999). In this method, the increase in permittivity increases the electrical energy density of the substrate, which increases the patch edge impedance, making it difficult to match the impedance. On the other hand, surface waves increase the permittivity of the substrate, reducing the overall radiation efficiency. In addition, the electrical energy density of the substrate raises the quality factor (Q), which reduces the operating bandwidth. Another approach uses a magneto-dielectric substrate that has a higher permeability than free space. The substrate consists of the artificial metamaterial unit cells proposed in Refs. (Mookiah and Dandekar 2009; Mosallaei and Sarabandi 2007; Pinsakul et al. 2012; Ikonen et al. 2006). Nonetheless, the magneto-dielectric presented in Refs. (Mookiah and Dandekar 2009; Mosallaei and Sarabandi 2007; Pinsakul et al. 2012; Ikonen et al. 2006) are bulky and cannot be included in planar circuits. Hence, the frequency bandwidth decreases as a result of the enhanced permeability of the host substrate.

The intrinsic impedance of a magneto-dielectric substrate defined as. $\eta = \eta_o \sqrt{\varepsilon_r/\mu_r}$. To design a magnetic dielectric substrate with $\varepsilon_r/\mu_r \cong 1$, the intrinsic impedance of the magnetic dielectric substrate is approximately equivalent

to the free space intrinsic impedance (η_o). This observation leads to this design in which the microstrip patch antenna is mainly radiated from both edges of the antenna to obtain a good matching, where the intrinsic impedance of the magneto-electric substrate is same as free space impedance. Thus, the reduction in bandwidth can be compensated by increasing the permittivity by manipulating the magneto-dielectric substrate.

Microstrip patch antennas are widely used and have become common in wireless communication systems. However, the microstrip patch antenna has a narrow band as a result of the resonant characteristics of the antenna (Balanis 2016). The dominant resonant frequency of the microstrip patch antenna can be formulated as:

$$f = \frac{c}{2(l+h)\sqrt{\mu_e \varepsilon_e}} \tag{2}$$

$$\varepsilon_e = \frac{\varepsilon_{r+1}}{2} + \frac{\varepsilon_{r-1}}{2}\left(1 + \frac{12h}{w}\right)^{-0.5} \tag{3}$$

where ε_e is the effective electrical permittivity, μ_e is the effective magnetic permeability of the main substrate, ε_r is the relative electric permittivity, w and l are the width and length of the rectangular patch antenna, and h is the thickness of the substrate. In accordance to Eq. (2) and Eq. (3), the resonant frequency is inversely proportional to the effective permittivity and permittivity. As a result, you can shift the resonant frequency down by increasing the dielectric constant product and the dielectric constant. In metamaterial structures such as SRR mentioned in Ref. (Mosallaei and Sarabandi 2007), the perpendicular magnetic field vector induces current in the ring. The induced current generates a magneto-dipoles that acts on the primary magnetic field, increasing the magnetic permeability of the substrate.

Magneto-Dielectric Substrate

As shown in the Fig. 4, the unit cell consists of three substrates: top, middle, and bottom. The base substrate has two vias, a ground plane, and a metal strip. The center substrate includes of a via and a metal strip. The top substrate comprises of a PEC layer on the top. Two strips and two vias with ground form an SRR-like structure. The magneto-dielectric substrate is kept under the patch antenna etched on the upper substrate. The magneto-dielectric substrate consists of 40 unit cells placed in two rows (2 x 20) encapsulated in the lower and center substrates as appeared in Fig. 4. The main dielectric substrate is made out of Rogers RO4003, $\varepsilon_r = 3.55$, and tan $\delta = 0, 0027$.

The height of the lower substrate (hl), the center substrate (hm), and the upper substrate (hu) are 0.813 mm, 0.203 mm, and 0.203 mm, respectively. Every unit cell has two vias, 0.813 mm and 1.016 mm long. The main parameters of the magneto-dielectric substrate can be deduced from the unit cell frequency response under certain boundary conditions (Fig. 3a) by the approach proposed by Smith et al. (Smith 2005).

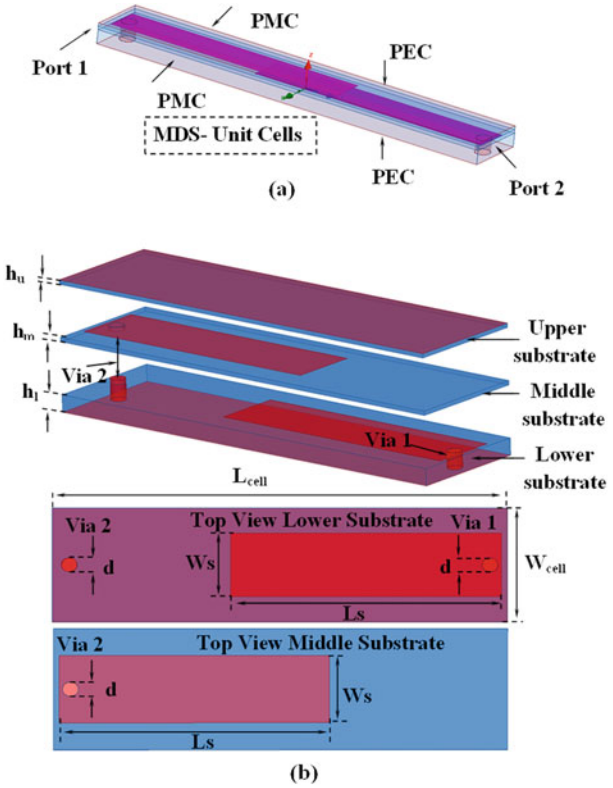

Fig. 3 Schematic of magneto-dielectric unit cell. The unit cell parameters are Wcell = 2.75, Lcell = 20, Ws = 1.6, Ls = 11.9, d = 0.7, hl = 0.813, hm = 0.203, and hu = 0.203 (all parameters in mm). (**a**) The unit cell under described boundary condition and (**b**) top and side views of the unit cell (Zaid et al. 2017)

As shown in Fig. 3a, the top and bottom sides are perfect electric transmitters (PEC) and the two sides (yz plane) are perfect magnetic conductors (PMC). The resonant frequency can be altered by changing the cell size of the unit cell. Unit cell length (Lcell) is the significant parameter, because it affects the resonant frequency of the unit cell. The increased unit cell length (Lcell) reduces the resonant frequency. The main specifications of the magnetic dielectric substrate are shown in Fig. 4. It is clearly shown that the electric permittivity steadily decreases as the frequency increases, and the magnetic permeability increases at higher frequencies.

The product of dielectric permittivity and magnetic permeability is about 20 in terms of frequency bandwidth (500–900 MHz) and decreases when the frequency exceeds 900 MHz. The amount of magnetic and electrical losses is about zero in the frequency bandwidth (500–900 MHz) and the losses raise when the frequency exceeds 900 MHz. At 850 MHz, the optimal value is a compromise between the product of dielectric permittivity and magnetic permeability and entire loss, so maximum efficiency can be achieved with the smallest possible size.

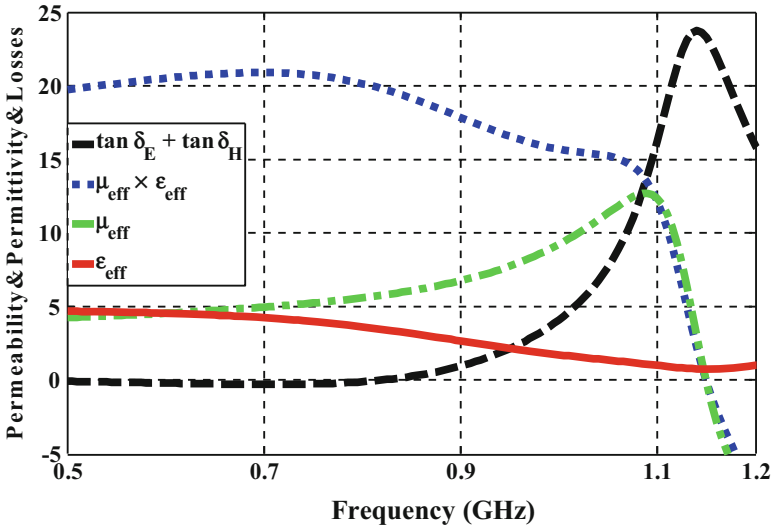

Fig. 4 Effective characteristics of the magneto-dielectric substrate. The dashed line (black) shows the sum of the magnetic and electric losses, the dotted line (blue) shows the dielectric permittivity times magnetic permeability, the solid line (red) shows the electric permittivity, and the dashed-dotted line (green) shows the magnetic permeability (Zaid et al. 2017)

Antenna Design Using Magneto-Dielectric Substrate

Using the magneto-dielectric substrate, the antenna can be made smaller compared to the conventional antenna as the effective permittivity and effective permeability increase (L = 89.6 mm, W = 118.4 mm using a Rogers RO4003 substrate with effective electric permittivity equal to 3.55). Therefore, we developed a miniaturized patch antenna with dimensions L = 40 mm and B = 55 mm as shown in Fig. 4. These dimensions were calculated for a magneto-dielectric substrate with an effective permeability multiplied by an effective permeability equal to 19, as this is the best trade-off value between miniaturization and loss. The cell specifications of MDS were calculated according to the size of the patch antenna.

The unit cell depends on two significant dimensions: the length of the strip and the cross section area between the strips. The resonant frequency of the unit cell is inversely proportional to the length of the strip and the cross-sectional area between the strips. Indeed, the two strips and the bottom PEC wall with two vias act as a split ring resonator (SRR). The resonant frequency of the MDS unit cell is determined by the loop capacity and inductance. Increasing the cross section area between the strips increases the loop capacitance and decreases the resonant frequency. Likewise, the length of the strip raises the loop inductance, reducing the resonant frequency. We place a microstrip patch antenna on the magneto-dielectric substrate and compare it with three antennas with a specific magnetic permeability and electric permittivity.

In these four cases, the same substrate thickness (hl + hm + hu) is maintained and all antennas operate at the same frequency. The four categories are:

1. In Case 1, the patch is etched on a high dielectric constant substrate with $\varepsilon_e = 19$ and $\mu_e = 1$.
2. Case 2 is a conventional patch antenna without miniaturization; the patch antenna is etched on an Rogers RO4003 substrate with $\varepsilon_e = 3.55$, $\tan\delta_e = 0.0027$, and $\mu_e = 1$.
3. Case 3 is the magneto-dielectric antenna, with the magneto-dielectric unit cell of Fig. 3. The product of effective magnetic permeability and effective electric permittivity is about 19 as illustrated in Fig. 4.
4. Case 4 is an equivalent magneto-dielectric substrate with constant specifications ($\varepsilon_e = 4.35$ and $\mu_e = 4.35$). The product of effective magnetic permeability and effective electric permittivity is about 19.

The simulated reflection coefficient for the four cases are shown in Fig. 6. Table 1 summarizes the simulation results for operating frequency, bandwidth, gain, and radiation efficiency for the four cases. For Case 1, Case 3, and Case 4, the dimensions are Wf = 2.73, Lf = 60.25, We = 0.85, Le = 46.09, L = 37.32, and W = 55, and for Case 2, the conventional patch antenna dimensions (without miniaturization) are Wf = 2.72, Lf = 60, We = 1.35, Le = 50, L = 89.6, and W = 118.4 (all units of dimensions in millimeters). Case 4 is the good antenna in terms of bandwidth, radiation efficiency, and gain. In that case, the intrinsic impedance of the substrate is equal to the intrinsic impedance of the free space (η_o). This behavior is because, in this case, we have better matching between the intrinsic impedance and the impedance of the free space (η_o).

As can be seen from Table 1, it is clear that Case 2 (conventional antenna without miniaturization) has the highest value of gain compared to other cases. However, its size is quite large. This behavior is due to miniaturization of the patch antenna, reducing the distance between the two radiating edges of the patch, which does not affect the efficiency of the antenna. The reduction in gain is due to the reduction in the distance between the two radiating edges, just like the antenna array. As the distance between the radiating elements in the antenna array decreases, the antenna gain decreases slightly. In the first case, the antenna substrate with a high dielectric constant is miniaturized, but the impedance band is narrow due to the high Q, and the radiation efficiency is low due to the large amount of energy captured by the dielectric.

The magneto-dielectric substrate antenna has similar impedance bandwidth and radiation efficiency as the conventional patch antenna, except the gain is reduced

Table 1 Results of the four substrates (Zaid et al. 2017)

Case	F_0 (GHz)	BW (MHz)	Gain (dB)	η%
1	0.851	3	−5	30
2	0.850	7	6	77
3	0.850	6	2	72
4	0.845	10	4	85

because of miniaturization the antenna, the distance between the two radiating edges of the microstrip patch is reduced, decreasing the gain. Nonetheless, the efficiency is almost the same as in the conventional microstrip patch. In both Case 1 and Case 4, the product of ε_e and μ_e is equal to 19, but the magnetic permeability (μ_e) is equal to the given values (for Case 1, it is 1, and for Case 4, it is 4.35), and the effective electric permittivity (ε_e) can be calculated by Eq. (2).

This calculation is different for Cases 1 and 4: for Case 1, $\varepsilon_r = 19$ and for Case 4, $\varepsilon_r = 4.35$; these other values cause the difference in resonant frequencies between Cases 1 and 4 (see Eq. (2)). The size of the magneto-dielectric antenna is decreased by about 80% compared to the conventional antenna. Figure 5 shows the structure of the magnetic dielectric antenna, and Fig. 7 shows the fabricated antenna prototype.

Figure 8 shows the comparison of measured and simulated input return loss for the designed magneto-dielectric patch antenna, where the measured input return loss is varied slightly towards higher frequency. The corresponding frequency characteristics of the simulation and measurement results differ by 7 MHz (frequency shift 0.8%). This frequency change between simulation and measurement is due to the adhesive layer (dielectric adhesive-EPO-TEK TJ1104-LH) connecting the three substrates together. The size of the designed magneto-dielectric antenna is decreased by about 80% compared to the conventional microstrip patch antenna and has similar frequency bandwidth. The radiation pattern of the designed magneto-dielectric antenna is shown in Fig. 9, where the measured results shown considerable similarity with the simulated ones.

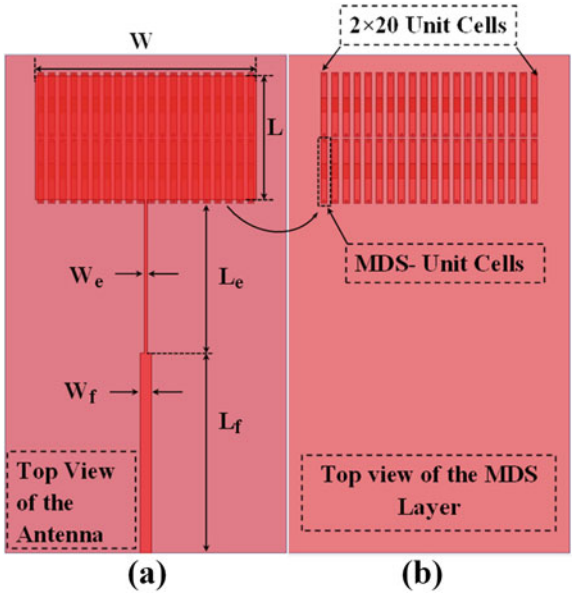

Fig. 5 Microstrip patch antenna with MDS. (**a**) Top view microstrip antenna and (**b**) top view of MDS (Zaid et al. 2017)

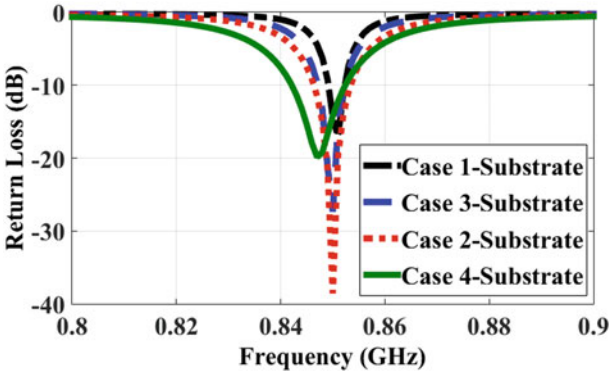

Fig. 6 Reflection coefficient of the four cases. (Zaid et al. 2017)

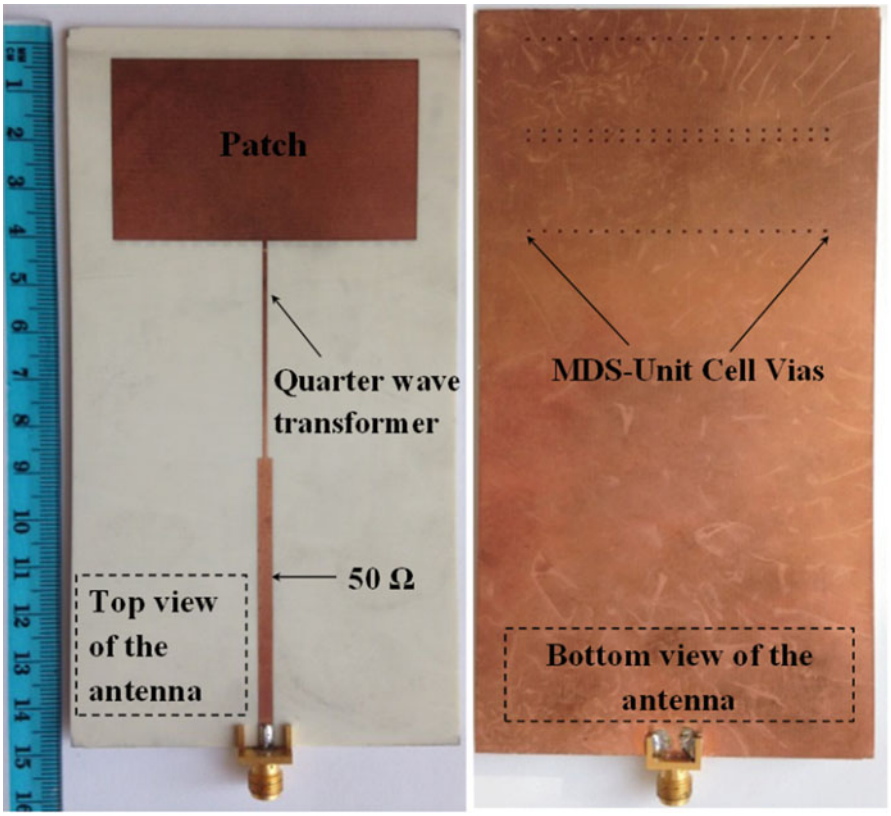

Fig. 7 Top and bottom views of the fabricated magneto-dielectric antenna. (Zaid et al. 2017)

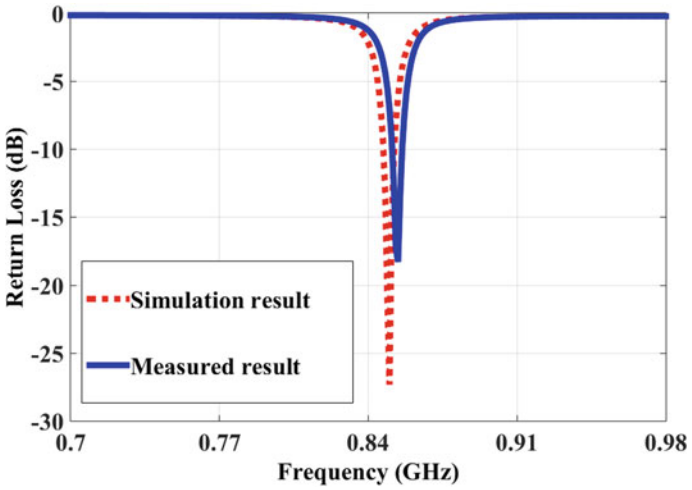

Fig. 8 The simulated and measured return loss of the magneto-dielectric patch antenna. (Zaid et al. 2017)

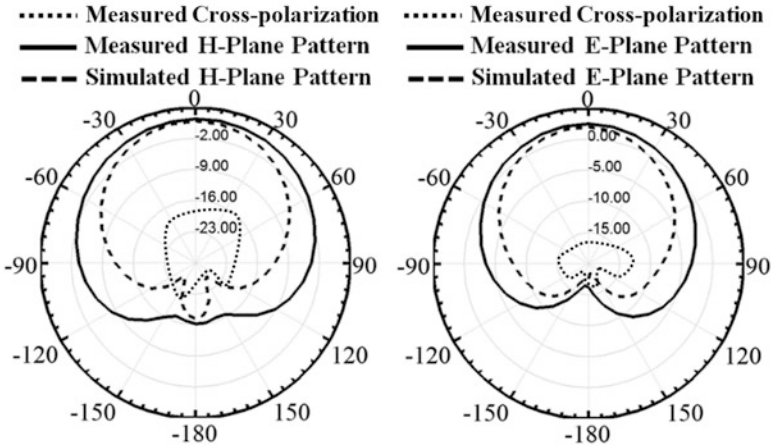

Fig. 9 Simulated and measured radiation pattern. (Zaid et al. 2017)

Self-Powered Multiport Patch Antenna Using MM for Passive UHF RFID-Tag Sensor

RFID-based sensors have become very appealing and interesting research topic among researchers because of their simultaneous use for detection and identification, low energy consumption, long life span and low cost (Marrocco et al. 2008; Babar et al. 2012; Gao et al. 2011; Merenda et al. 2018). Various self-powered RFID-tag

sensors have been presented in the literature presented in Refs. (Marrocco et al. 2008; Babar et al. 2012; Gao et al. 2011, 2013; Merenda et al. 2018; Sidén et al. 2007; Kim et al. 2015; Jiang and Yang 2012; De Donno et al. 2014; Abdulhadi et al. 2017; Abdulhadi and Abhari 2012; 2016).

Two nodes are used in the self-powered RFID tag-based sensor. One of the nodes is used as a reference node and the other node considered as a sensor node. These nodes make the required operating power from the signals received by the RFID reader. Passive sensors are used to measure physical quantities such as temperature, humidity, pressure, etc., in the sensor node circuit. The other node connected to the RFID chip is considered as reference only. Considering both nodes are activated by a signal received from the reader, changes in temperature or humidity will result in a false match between the antenna and sensor device. So, the sensor node needs further power to activate the RFID chip in comparison to the reference node which generates activation power from the same RF signal received by the RFID reader. Through measuring the minimum power, the reader should activate the sensor and reference nodes, and the measured physical value is determined based on the measured power ratio for both nodes. For example, some previous sensor systems used omnidirectional antennas, a dipole, folded loop antenna meander monopole as introduced in Refs. (Marrocco et al. 2008; Babar et al. 2012; Gao et al. 2011, 2013; Merenda et al. 2018; Sidén et al. 2007; Kim et al. 2015; Jiang and Yang 2012; Abdulhadi et al. 2017; Abdulhadi and Abhari 2012), respectively. However, since these systems are self-powered, directional antennas cannot be mounted on metal surfaces. Antenna characteristics such as radiation pattern, resonant frequency, and efficiency are reduced. Therefore, the sensor cannot provide accurate results. Half-omnidirectional antennas such as patch antennas are used to solve the previous problem in other systems. Thus, the patch antenna can be fixed on a metal surface without compromising performance, as suggested in previous work (Abdulhadi and Abhari 2016; Abdulhadi and Denidni 2017; Jamal Zaid et al. 2017). Some systems used multiport antenna as described in Refs. (Abdulhadi and Abhari 2016; Abdulhadi and Denidni 2017; Jamal Zaid et al. 2017), containing a reference node and a sensor node on one antenna. However, this method has several drawbacks. For example, the antennas are large and use batteries, microcontrollers, and solar cells (Abdulhadi and Abhari 2016; Abdulhadi and Denidni 2017).

To solve the previous problem, a completely self-powered antenna and work in the North American 902–928 MHz frequency band is described. This antenna is connected to two RFID chips, one of which is marked as a reference node and the other is soldered parallel to the resistive sensor to serve as the sensor node. The antenna can be mounted on a metal surface, and a magneto-electric substrate can be used to miniaturizing microstrip patch antenna by about 75% compared to the conventional patch antenna, and the antenna can replace the use of two antennas. In addition, this antenna has a small size and completely self-powered compared to the previous works (Abdulhadi and Abhari 2016; Abdulhadi and Denidni 2017; Jamal Zaid et al. 2017). In this section, we have looked at two examples. The two antennas are used in the first case, for the reference node and for the sensor node. A

multiport antenna consists of the reference node and a sensor node in one antenna as the second case. Instead of a resistance sensor, a variable resistance (10, 20, 50, 100, 200 Ohm) is used and is measured in all cases.

MM Unit Cell Design

Microstrip patch antennas are becoming more popular in wireless networks because they have many preferences, such as less expensive, easy to produce, and integrated with flat circuits. However, the microstrip patch antenna has narrow and relatively less bandwidth. The size of a microstrip antenna is inversely proportional to the product of the magneto permeability and dielectric permittivity. The split-ring resonator (SRR) was used to reduce the size of the antenna as suggested in Refs. (Mookiah and Dandekar 2009; Pinsakul et al. 2012). According to this, a magnetic field perpendicular to the ring (loop) induces a current in the ring, and a magnetic dipole is created as a result of the inductive current. As a result, the product of permeability and permittivity in the substrate constant increases.

This approach can reduce the antenna size by compensating for bandwidth degradation. However, in previous works (Mookiah and Dandekar 2009; Pinsakul et al. 2012), the layer of the metamaterial substrate was thick and could not be used for a flat design. The thickness of magnetic dielectric substrate is less enough for a planar design. As depicted in Fig. 10, under some boundary conditions, the zy-plane wall is the perfect magnetic conductors (PMC), the xy-plane wall is the perfect electric conductors (PEC), and the xz-plane wall includes ports. The effective

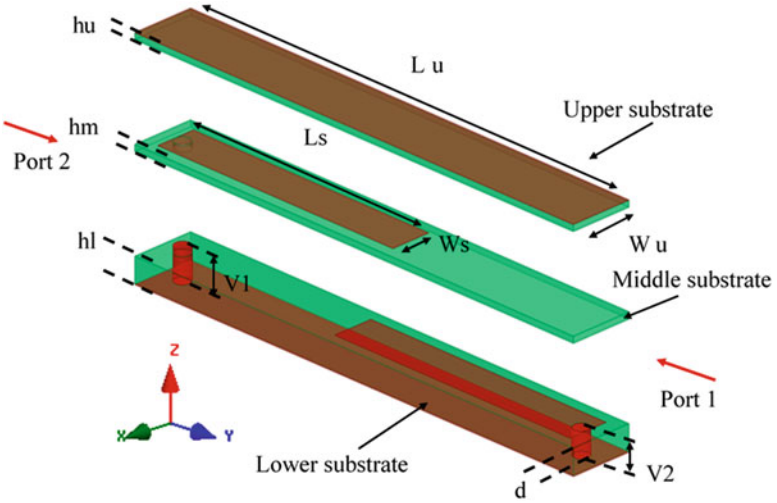

Fig. 10 The MDS unit cell under specified boundary condition. Dimensions are Wu = 2.76, Lu = 23.5, Ws = 1.8, Ls = 13.65, d = 0.7, hl = 0.813, hm = 0.203, and hu = 0.203 (all dimensions in mm). (Zaid and Denidni 2019)

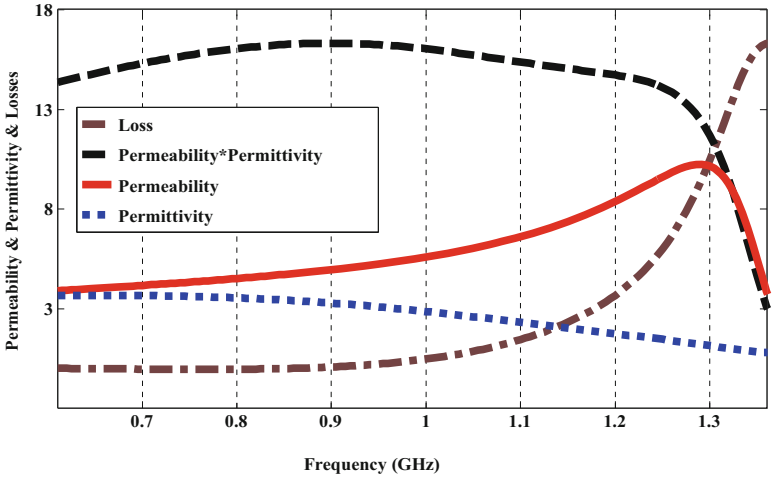

Fig. 11 Effective constitutive parameters of the MDS unit cell. Dashed-dotted line is shown the sum of the magnetic and electric losses, dashed line is shown the product of dielectric permittivity and magnetic permeability, solid line is shown magnetic permeability, and dotted line is shown electric permittivity. (Zaid and Denidni 2019)

parameters of the unit cell are extracted as shown in Fig. 11. The product of electric permittivity and magnetic permeability is about 16 at 915 MHz and losses are almost zero at the same frequency. The magneto-dielectric substrate layer consists of 34 unit cells (2×17) and is placed under the patch. The unit cell is etched on Rogers substrate (RO4003 with $r = 3.55$ and $\tan \delta = 0.0027$) and consists of three substrate layers of varying thickness: top substrate ($hu = 0.203$ mm), central substrate ($hm = 0.203$), and lower substrate ($hl = 0.813$ mm). The bottom substrate consists of two vias, a ground, and a strip (PEC) on top. The central substrate has via and a strip at the top. The top substrate contains a PEC layer, which is considered a patch antenna.

Multiport Microstrip Patch Antenna Design

Using the designed magneto-dielectric substrate, the size of the patch antenna is miniaturized at about 75% in comparison to the conventional patch printed on the substrate (RO4003 with $\varepsilon_r = 3.55$ and $\tan\delta = 0.0027$). Passive RFID is activated by a signal received by the reader. Therefore, the received signal depends on the characteristics of the tag antenna and the matching between the tag antenna and the RFID chip. The patch antenna is connected to the RFID chip by matching an inductive loop for conjugate matching ($Z^{*}_{chip} = Z_{antenna}$). So, most of the power received by the reader is sent to the RFID chip and vice versa. At 915 MHz, the input impedance of the RFID chip (RI-UHF-IC116–00) is 25-j190, and the conjugate impedance is $25 + j190$. The RFID chip is fed by a lumped port (24 ohm) and a capacitor (0.95 pF) as the equivalent element. Figure 13 depicts the return loss of the

designed antenna at the operating frequency (915 MHz) of about −16 dB, which is a better antenna/loop equivalent. Figure 14 shows a manufactured antenna A matched to one RFID chip.

The patch antenna is implanted on a magneto-dielectric substrate consisting of 34 unit cells (2 × 17) as shown in Fig. 12. Thus, the single port antenna (antenna A) in Fig. 5 is designed and manufactured using a magneto-dielectric substrate. The inductive loop method is used to match between the patch antenna and the RFID chip. The imaginary part can be controlled by changing the dimensions (length and width) of the induction loop, and adjusting the real part by varying the distance between the patch antenna and the induction loop to ensure a match. The antenna and

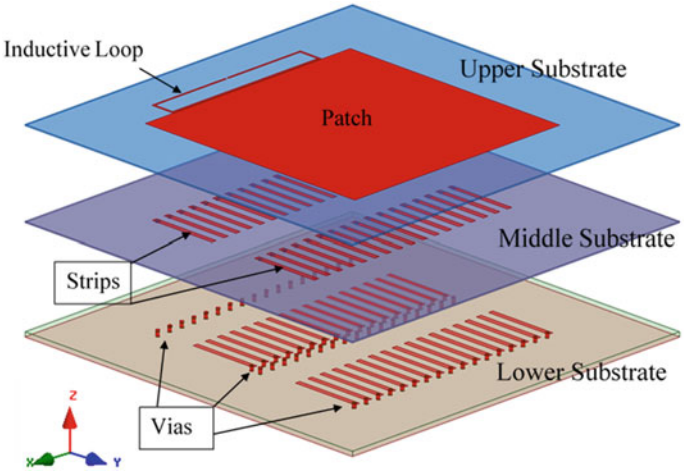

Fig. 12 Structure of the proposed antenna. (Zaid and Denidni 2019)

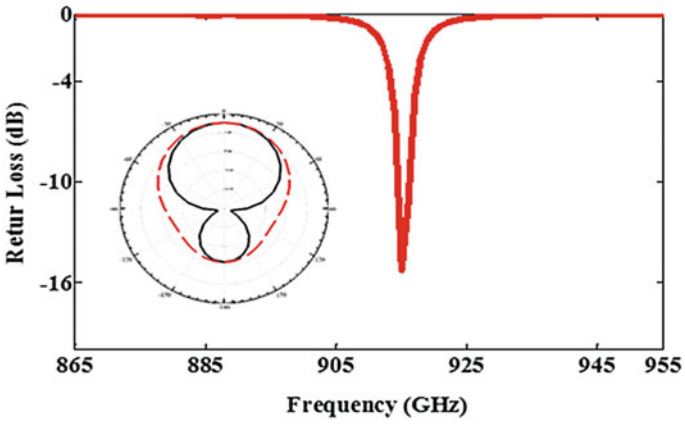

Fig. 13 Return loss of the designed antenna and radiation pattern. (Zaid and Denidni 2019)

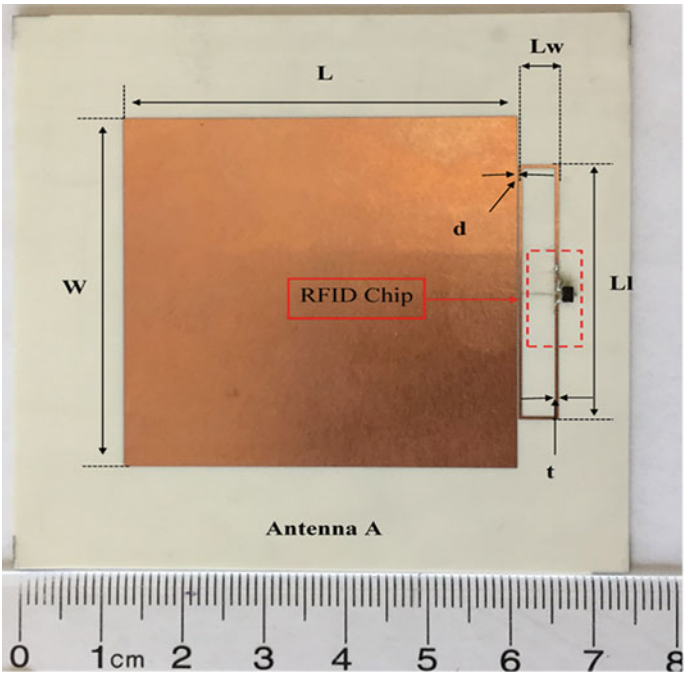

Fig. 14 Fabricated prototype. Dimensions are L = 48, W = 47, Lw = 4.8, Ll = 34, d = 0.2, and t = 0.4 (all dimensions are mm). (Zaid and Denidni 2019)

reader (GAO 216010) are adjusted according to the line of sight and then gradually increase the distance until the maximum reading interval is reached. The maximum value of the reading range is about 11 m at 937 MHz as a result of dielectric tolerance ±0.05 and manufacturing accuracy. The resonant frequency shifts from 915 MHz to 937 to about 22 MHz as shown in Fig. 15. The power sensitivity of RFID is about -20 dBm. However, the strength of the tag builds upon the characteristics of the antenna and the compatibility between the antenna and the RFID chip. The power required to activate the RFID chip is measured and shown in Fig. 16. According to these figures, the minimum power required is about 13 dBm at 937 MHz.

Single-Port Antenna-Based Sensor

Two dipole antennas were used in some systems introduced in the literature (Chakraborty et al. 2014; Colburn and Rahmat-Samii 1999; Mookiah and Dandekar 2009; Mosallaei and Sarabandi 2007; Pinsakul et al. 2012; Ikonen et al. 2006) as sensor nodes and the other as reference nodes. The dipole has omnidirectional radiation, so it cannot be mounted on a metal surface. It is also affected by the human body, as demonstrated in Ref. (Colburn and Rahmat-Samii 1999). Patch antennas have been used for RFID-based sensors described in Refs. (Abdulhadi and Abhari 2016;

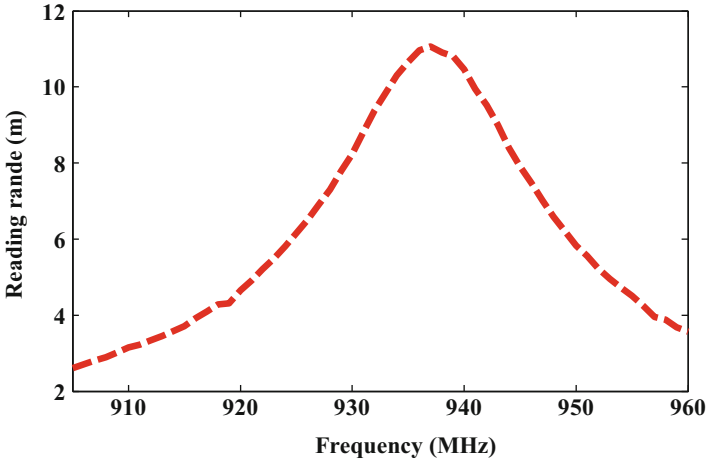

Fig. 15 Measured reading range of antenna A

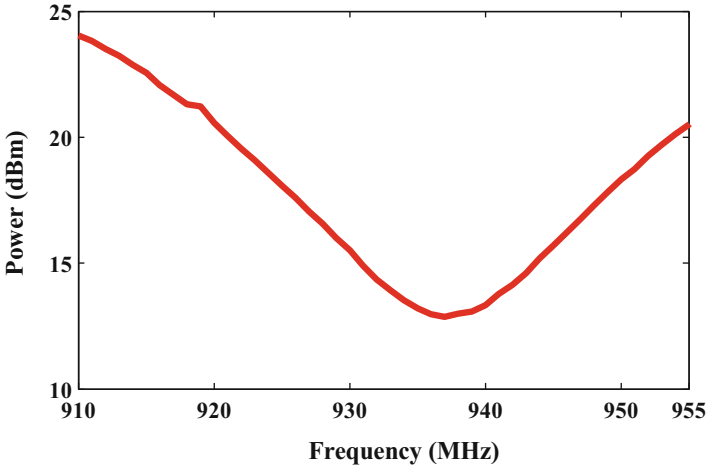

Fig. 16 Measured minimum required power for activating antenna A RFID chip. (Zaid and Denidni 2019)

Abdulhadi and Denidni 2017; Jamal Zaid et al. 2017). Nevertheless, the size of the patch antenna is large. Due to this, in this chapter, for miniaturization of the proposed antenna, a magnetic dielectric substrate layer is placed underneath, which is the smallest size compared to previous studies (Abdulhadi and Abhari 2016; Abdulhadi and Denidni 2017; Jamal Zaid et al. 2017).

In order to use the RFID tag as a sensor, a resistance sensor (varying resistance) is connected in parallel to the RFID chip as depicted in Fig. 17 (antenna B). As far the air conditioning room cannot be used during measurements, a variable resistor is

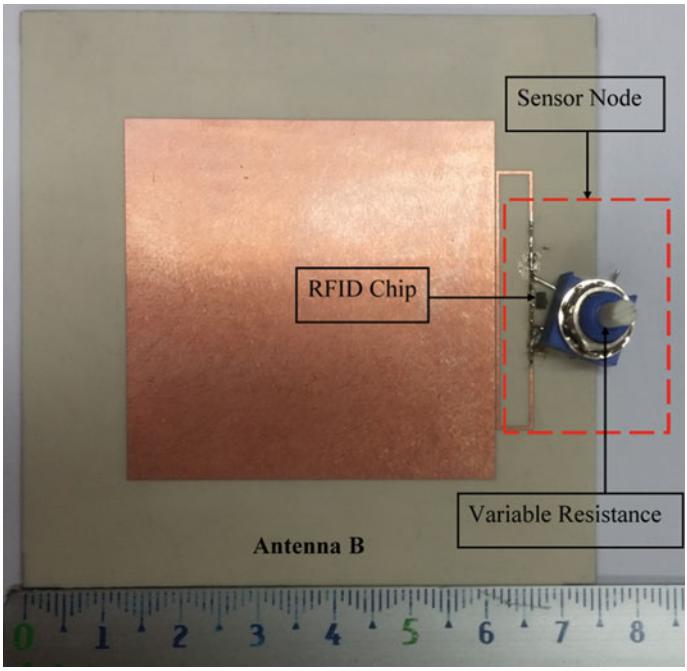

Fig. 17 Fabricated antenna with resistive sensor (antenna B). (Zaid and Denidni 2019)

used as a substitute for the resistive sensor. Antenna A is used as a reference node, and antenna B is used as a sensor node. By changing the variable resistor to 10, 20, 50, 100, and 200 ohms so that the match between the reference nodes does not change the resistor value at all. The minimum power required for frequency is measured at each resistance value (10, 20, 50, 100, and 200 ohms) and is shown in Fig. 18. Due to this, the minimum power required is inversely proportional to the resistance value, as shown in Fig. 18. The maximum value of the reading range is measured at each resistor depicted in Fig. 18. Lowering the resistance value increases the required power. The minimum power to activate the sensor node are 25.36 dBm, 23.5 dBm, 21.72 dBm, 19.53 dBm, and 17.56 dBm at 10 ohms, 20 ohms, 50 ohms, 100 ohms, and 200 ohms, respectively. The minimum value of power of the reference node is constant (13 dBm) for each sensor resistance value. Comparing the minimum power of a reference node at 937 MHz with the minimum power of a sensor node, for different resistance values at the same frequency as shown in Fig. 19, we can determine the temperature or humidity.

Multiport Antenna-Based Sensor

In the previous section, the RFID tag sensor used two antennas, a reference node and the other as a sensor node. Since some previous systems used multiple antennas, the

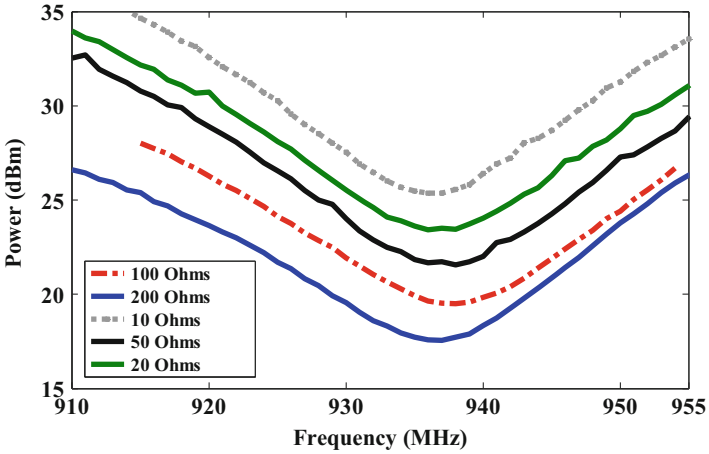

Fig. 18 Measured minimum power to activate sensor node with different resistance (antenna B)

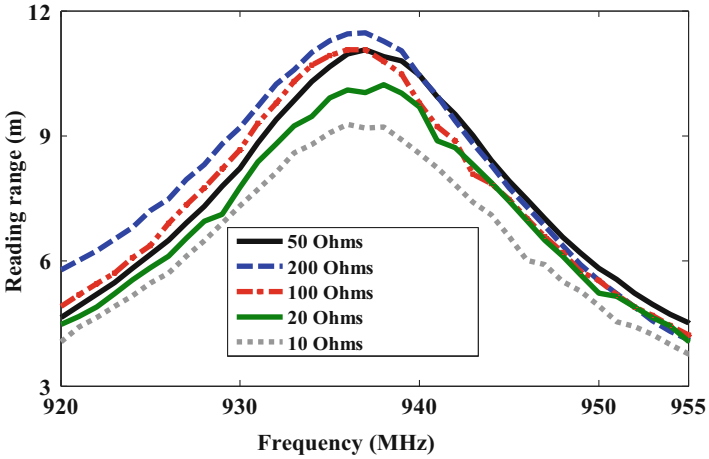

Fig. 19 Measured reading range for variable resistance (antenna B). (Zaid and Denidni 2019)

reference node and sensor device are part of the same antenna (Abdulhadi and Abhari 2016; Abdulhadi and Denidni 2017; Jamal Zaid et al. 2017). However, the multiport antenna sensor has preferences such as low cost, the sensor unit and the reference unit receive the same signal from the reader, and both nodes are still large in the same ambient weather conditions. Accordingly, the designed antenna is a miniaturized multi-antenna. The patch antenna is implanted on a layer of a magneto-dielectric substrate and interfaces with two RFID chips using an inductively coupled loop. One of the RFID chips is used as a reference node and the other is a sensor node, which is connected next to a resistive sensor (variable resistor) (see Fig. 20) (antenna C). The variable resistor was changed to 10, 20, 50, 100, and 200 Ohm with

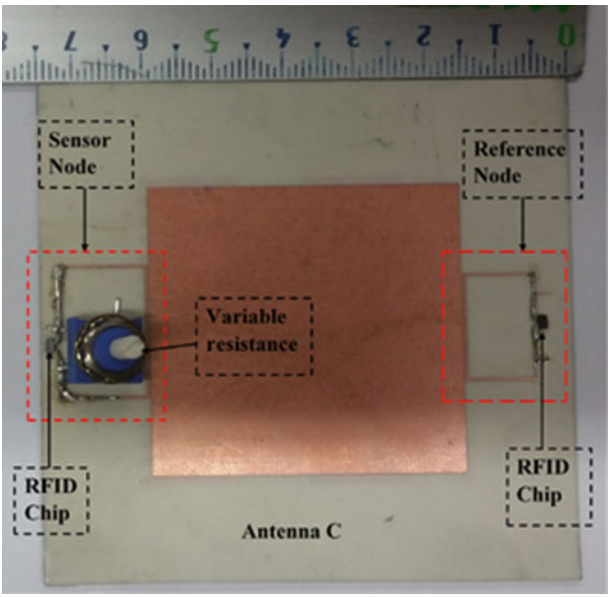

Fig. 20 Fabricated multi-port antenna (antenna C). (Zaid and Denidni 2019)

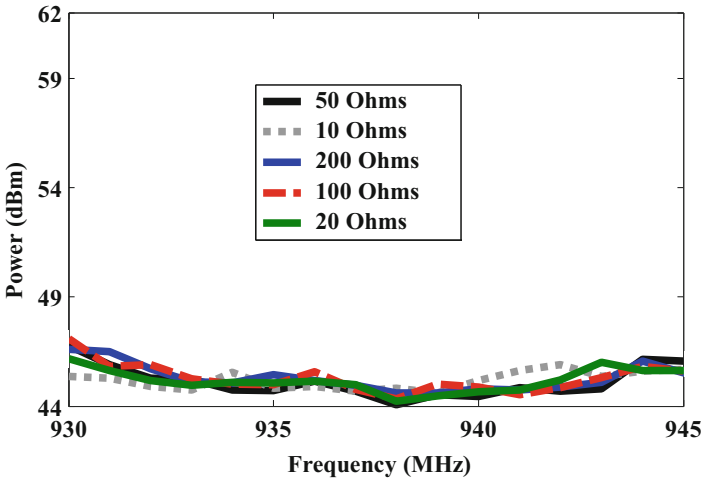

Fig. 21 Measured minimum power needed to activate the reference node with variable resistance (antenna C). (Zaid and Denidni 2019)

the minimum power required at the reference and sensor nodes depending on the frequency measured at each resistor. The minimum power needed to activate the sensor node at 937 MHz is 35.79, 33.8, 32.03, 29.16, and 27.15 dBm with RTDs of 10, 20, 50, 100, and 200 ohms, as shown in Fig. 21. However, the minimum power

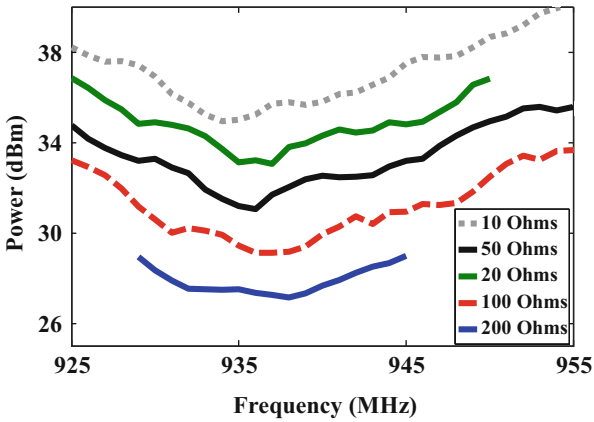

Fig. 22 Measured minimum power needed to activate the sensor node with variable resistance (antenna C). (Zaid and Denidni 2019)

required for the reference node is approximately the same at about 44 dBm, as shown in Fig. 22. Temperature or humidity can be extracted from the power differences. The minimum power of the multi-part antenna (C) is higher than the minimum power of the single port antenna (B) due to the coupling between the two ports.

Conclusion

RFID is a novel technology and is one of the fastest growing segments of the modern automatic identification and data capture (AIDC) industry. In case we comprehend it or not, RFID is an essential part of our lives and being used in hundreds of day-to-day applications. In this chapter, we have described application of metamaterial for RFID antennas. The magneto-dielectric substrate metamaterials were used with the UHF RFID antennas for the miniaturization of size.

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Additive Manufacturing of MTM-FSS

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S. S. Karthikeyan and R. Adeline Mellita

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Abstract

Additive Manufacturing (AM) techniques have emerged as a rapid prototyping technique for microwave components/circuits and antennas; this chapter reviews different AM technologies and its applications for the development of frequency selective surfaces (FSSs). The 3D/metamaterial (MTM) FSS models which perform better than their planar counterparts were tedious to develop with

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primitive PCB manufacturing techniques. Adaption of AM for microwave component manufacturing has rendered the prototyping of such 3D FSS models an effortless possibility. AM technologies such as: Stereolithography (SLA), Selective Laser Sintering (SLS), and Fused Deposition Modelling (FDM) which have been used for microwave component manufacturing are discussed. Further, the techniques used to characterize the electrical properties of AM materials are also discussed. Incorporating conductive elements in an additively manufactured microwave component is challenging. Methods like inkjet printing, aerosol jet printing, silver nanoparticle paint, and conductive tapes were used for this purpose. This resulted in two-step manufacturing where the dielectric part was additively manufactured and the conductive part was further added using alternative methods. Advancements in FDM process made it possible to print the conductive elements seamlessly to the microwave component, eliminating the need for a two-step manufacturing process. Although the adaptation of AM provides benefits such as low cost and easy manufacturing, the major obstacle is the lack of highly resistive or conductive AM materials. With the growing interest on AM in the field of RF and microwave engineering, the material research in the field of AM shows promising signs of discovering electrically efficient AM materials which could prove beneficial for manufacture of future microwave components.

Keywords

Additive manufacturing · Frequency selective surfaces · Metamaterials · PCB manufacturing

Introduction

The manufacturing domain has transformed drastically with the advent of Additive Manufacturing (AM) technologies (Ian Gibson 2015). This technology commonly called as "3D-printing" is an alternative to the conventional PCB manufacturing (Clark 2012) and Computer Numerical Control (CNC) machining which are examples of subtractive manufacturing. AM has applications (Gebhardt 2011) in automotive, health-care, art, architectural modelling, landscaping, and product development. It also helps in rapid and economic prototyping with its ability to build faster, utilizing a single tool. AM techniques directly use 3D-CAD data to add layer-upon-layer of material without using any part-dependent tools. These manufacturing techniques can be used with materials like polymers, ceramics, and metals (Zhang and Jung 2018). The current sophistication available with AM technology has made economical fabrication of complex and arbitrary 3D geometries without machining a possibility (Celik 2020). Recently, many research groups have successfully demonstrated the use of 3D printed microwave components proving AM to be a promising manufacturing technology for future RF and microwave devices. The dielectric part of RF components such as horn antennas (Zhang

et al. 2018; Oktafiani et al. 2019), patch antenna (Fawaz et al. 2018), cavity antenna (Zhu et al. 2020), array antenna system (Ketterl et al. 2015; Yu et al. 2018), transmission lines (Cai et al. 2016; Lopez-Villegas et al. 2019), waveguide components (Naeini and van der Weide 2019; Shu et al. 2020), and frequency selective surfaces (FSS) (Kronberger and Wienstroer 2017; Lleshi et al. 2019; Yang et al. 2020) have been prototyped using AM technologies. However, the conductive parts of the component have been manufactured by inkjet printing (Cui et al. 2019), aerosol jet printing, electroplating (Lopez-Villegas et al. 2019), wire-mesh embedding (Yu et al. 2018), or even manually coating with conductive paint (Palazzi et al. 2019). This results in a two-step prototyping process, where the dielectric part was built using AM and conductive parts were later added using alternate techniques mentioned above. Materials with high conductivity required by the conductive part of the RF component and its compatibility with the AM infrastructure prevents the integration of this two-step manufacturing process into a single-step process. But, the adaptation of AM for RF and microwave component manufacturing has fueled the study and development of new materials. For example, materials with higher dielectric constant (Czarny et al. 2018), magnetic materials (Panwar et al. 2015), polymer filaments with metal powders (Kronberger and Soboll 2016) for improved conductivity were developed. The current chapter discusses the application of AM for FSS prototyping. FSS are periodic structures (Munk 2005) which either pass or block a certain range of frequencies of electromagnetic waves. In a conventional FSS, a patch-type metallic structure provides band-stop frequency response while a slot-type structure gives a band-pass response. FSS can function as a reflector, absorber, spatial filter, or polarizer based on its frequency response, which in turn depends on the geometrical and material properties of the structure. 3D FSS models have superior performance compared to their printed circuit board (PCB) manufactured planar counterparts (Zhang et al. 2020). By using the theory of metamaterial (MTM) design, these 3D models can be made more efficient (Capolino 2017). Another class of FSS called all-dielectric FSS has come into prominence since they have no conductive parts, its frequency response completely depends on the geometrical and material aspects of the dielectric (Barton et al. 2015). The geometrical aspect of the structures would not only include the outer structure but also the layer-wise micro-structure. However, manufacturing such 3D structures using substractive manufacturing processes is a challenge as this process requires removing or cutting away excess material from a larger block of material. In contrast, these complex 3D structures can be effortlessly manufactured using AM techniques.

This chapter is divided into four sections. The first section introduces the adaptation of AM technologies in RF and microwave engineering. The next section explains the working principles of a few selective AM techniques, their advantages over PCB process, and their contribution to microwave component prototyping. The following section explains the working of an FSS and presents a brief summary of recent works, where adapting AM proved to be advantageous for MTM-FSS fabrication. The final section outlines the current challenges and future prospects of this technology for MTM-FSS prototyping.

AM in RF Engineering

Conventionally, RF component are manufactured using PCB technology, this limits the type of structures that can be used in a RF component. AM is an alternative to PCB manufacturing process, it offers greater flexibility in the type of structures that can be used in an RF component. These techniques include Stereolithography (SLA), Selective Laser Sintering (SLS), and Fused Deposition Modelling (FDM). This section reviews these important AM techniques which are commonly adopted for manufacturing RF components. These AM techniques are then compared with PCB manufacturing process and the benefits of manufacturing FSS using AM are discussed.

Stereolithography

Stereolithography (Bártolo 2011) is a liquid based process where a photosensitive polymer is cured or solidified by an incident ultraviolet laser. The curing process is an exothermic process which enables the bonding of large numbers of small molecules forming a highly cross-linked polymer. The laser beam supplies the energy required for the curing process. An SLA printer contains arrangements to facilitate this curing process. It has the following components, resin vat, re-coating blade, build-platform, and a UV laser source. This printer takes the STEP file of a CAD model as input. The instructions to move the build-platform and focus the UV laser are provided via the STEP file. For complex models, supporting structures are added to the CAD model, these are thin cylindrical structures used to securely attach the part to the print platform, they also prevent the model from warping during the curing process. Figure 1 illustrates the working arrangement of an SLA printer. Based on the STEP file information, the laser beam is focused on the liquid polymer resin in the vat. The UV radiation solidifies the polymer resin in the required shape, forming the first layer of the model. The platform then moves down by one layer dipping the model into the resin filled vat. The re-coating blade sweeps over the model to evenly cover the model with resin. The laser is then directed to form the next layer of the model which bonds with the previous layer. This process of polymer solidifying, moving the platform downward and re-coating with the blade is repeated until all the layers are formed. The completed model is then removed out of the resin vat and any supporting structures if present are eliminated to obtain the final finished product.

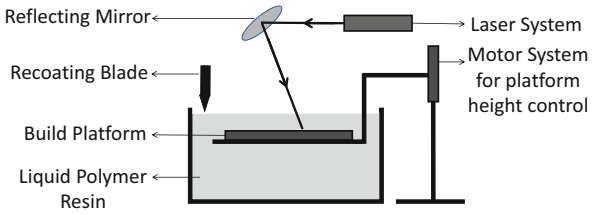

Fig. 1 Schematic of stereolithography process

Compared to a conventional PCB manufacturing process, SLA offers greater flexibility in FSS manufacturing. For a planar FSS manufactured using a PCB manufacturing process, the thickness of the dielectric and conductive layers is fixed. While each PCB machine has its own minimum spacing value, most of them have conductive trace dimensions limited to 0.0762 mm. However, if an SLA-based AM technique is used to print the substrate layer, its thickness can be varied from 0.01 mm to 200 mm. The conductive traces can be further added on the dielectric substrate using inkjet or aerosol jet printing. For a moderate resolution and a minimum trace width (up to 0.05 mm) inkjet printing can be used, while aerosol jetting may be used for higher resolution and trace width of up to 0.01 mm, such high resolution is required for FSS operating at THz frequencies. For such an FSS, a minor discontinuity may result in large deviation from the desired response. While at GHz frequencies the resolution provided by inkjet printing is adequate. Inkjet printing technology is cost-effective and compatible with a wide range of substrates; it is also possible to achieve an improved conductivity by jetting multiple layers.

In case of 3D structure manufacturing, PCB process cannot be used unless the 3D structure is a cascaded layer structure. Whereas with AM, it is feasible to manufacture both 3D and cascaded models. SLA gives very high resolution for manufacturing 3D objects with a smooth finish for flat surfaces at a moderate cost. Even complex 3D models can be made with the help of removable support structures.

In an SLA process, the print pattern or the print fill percentage cannot be varied resulting in lesser flexibility in controlling the dielectric constant and the loss tangent. Moreover, a two-step process is generally required as SLA supports only photosensitive polymer material but not conductive materials. However, a recent work by Quan et al. (2021) demonstrated the use of conductive fillers like carbon nanotubes (CNTs) within the polymer resin to improve its conductivity. An FSS developed using SLA process cannot be used in applications with prolonged UV exposure, since the AM material used is sensitive to UV light. Thus, with further advancement in AM, FSS with conductive elements may be printed with a single step SLA process.

Selective Laser Sintering

Selective laser sintering (SLS) is a powder-based process (Gebhardt 2011) where small particles of either plastic, ceramic, metal, or glass are sintered by a high power laser, preferably a carbon dioxide laser. This technique is advantageous for mechanically tough and complex 3D prototypes, where cost can be compromised for resolution. However, for planar FSS models, PCB process could prove more feasible than SLS technique.

First, the supply chamber is filled with the required powder, while filling the build chamber for a single layer. The STEP file of the required CAD model is loaded in the printer to provide model information in the form of layers. Unlike an SLA process, support structures are not required for SLS printing since the un-sintered powder provides support. Based on the model information, the laser is focused on the heated

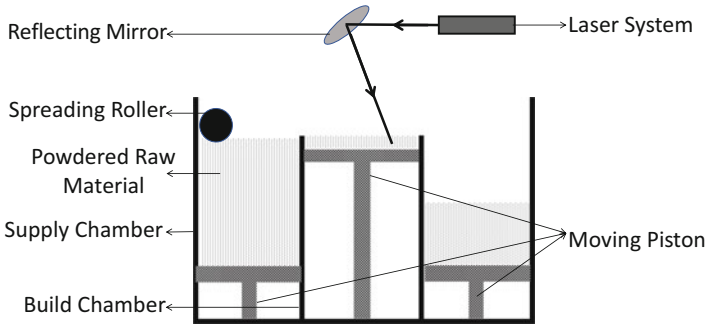

Fig. 2 Schematic of selective laser sintering process

build platform, which sinters the powdered material to the required shape forming the first layer of the model. Then the build platform moves down by one layer while the platform in the supply chamber moves up. A roller slides across the chambers to supply the next layer of powder to the build chamber. Any excess powder in the build chamber is dumped in the collection chamber. The laser is again directed to form the next layer of the model which then bonds with the previous layer. These steps of sintering and re-powdering are continued until all the layers of the model are built. Figure 2 illustrates the working of an SLS printer. The completed model encased in the powdered shell is removed out of the build chamber. Finally, unsintered powder is removed by manual brushing and low pressure sand blasting.

Similar to SLA, SLS technique can also build a substrate layer with varied thickness from 0.01 to 300 mm. But the conductive layers need to be manually coated using silver paint. However, for FSS-absorber applications powdered composites with improved conductivity can be used to build 3D structures avoiding the need for manually coated conductive layers. Although a copper foil needs to be attached at the bottom of the absorber to support zero transmission (Zhou et al. 2016; Abdullahi and Ali 2021). SLS technique is more advantageous for complex 3D structures because of its high resolution and lack of support structures. Like SLA, the print pattern or the print fill percentage cannot be altered even in SLS technique.

Fused Deposition Modelling

Fused deposition modelling (FDM) also known as fused filament fabrication (FFF) is a process (Gebhardt 2011) which melts and deposits the raw material layer by layer with heated material extrusion to form the required model. The raw material filament is fed from a large spool of thermoplastic. The FDM printers can have more than one extrusion nozzle to support printing of multi-material parts. Figure 3 illustrates the working of a conventional FDM printer.

The STEP file of the required CAD model is loaded into the printer. The printer parameters, such as temperature for the bed and nozzle, are set according to the chosen material. The layer height, print fill percent for the model, and necessary

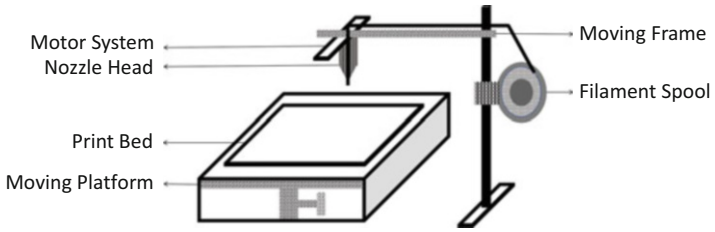

Fig. 3 Schematic of fused deposition modelling process

support structures are added using the slicing software. Based on the model information, the heated nozzle is moved to deposit the first layer on the printer bed. The bed is maintained at a lower temperature to solidify the melted thermoplastic material. The heated nozzle and the bed moves in the horizontal and vertical plane respectively. After the deposition of the first layer, the print bed is lowered by one layer. This process of moving and depositing continues until the whole model is fabricated. FDM can be considered as the most versatile AM technology for microwave component manufacturing. For manufacturing planar FSS, the dielectric layer thickness can be varied from 0.1 mm to 600 mm. The electrical properties of the dielectric substrate layer can also be altered by modifying the print fill percent and print patterns (Mirzaee and Noghanian 2016). This gives us an additional design freedom to minimize the weight of the FSS for the required electrical properties. With advancements in AM material science even conductive type FDM filaments have been made available. Using FDM, planar FSS have been manufactured with single step process where both the conductive and dielectric layers are printed seamlessly.

FDM has been used to fabricate FSS with resistive type elements, where screen printing technology has been used to generate resistive patches which are added on the FDM made dielectric layer. For 3D structured FSS, composite FDM filaments which contain conductive material like graphite or carbonyl iron have been used (Kjelgård et al. 2018; Liu et al. 2019). FDM technique is preferred for low-cost prototyping of moderately complex FSS. Like SLA, FDM also requires support structures to manufacture complex 3D structures, these support structures can be removed in post-processing along with surface smoothening. The material library of FDM is very large enabling high design flexibility. Moreover, the cost for prototyping is quite low, making it a go to choice for RF engineers.

A Comparison of AM Technologies

This section presents a brief comparison of the previously discussed AM technologies and their application in manufacturing RF components, Table 1 summarizes the comparison. This comparison may be useful to decide on the type of AM technology required for a microwave component. The SLA technique is more preferable for moderate cost prototyping with 3D structures with no exposure to direct sunlight. The material currently available is mostly polymer resins although recent studies

Table 1 Comparison between popular AM techniques

Attributes	SLA	SLS	FDM
Resolution	Very high	High	Moderate
Surface finish	Smooth	Slightly rough	Rough
Design complexity	High	Very high	Moderate
Ease of use	High	Moderate	Very high
Min. layer thickness	0.01 to 0.015 mm	0.01 to 0.05 mm	0.127 to 0.5 mm
Min. wall thickness	5 mm	0.8 mm	1 mm
Support structures	Required	Not Required	Required
Cost	Moderate	High	Low
Pro	High accuracy	Strong and complex parts	Fast printing
Con	UV sensitive	Limited material options	Lower resolution

show a trend to make conductive polymers by adding fillers to the resin. The SLS technique is more preferable to build mechanically strong and complex 3D structures. It also has a wide material choice including polymers, metals, ceramics, glass, and various composite material powders. However, the cost of these prototypes are very high mostly used for aerospace and medical applications. FDM is the preferred low-cost rapid prototyping technique; it also provides additional design freedom to modify the components electrical properties by suitably choosing the print fill pattern, which in turn may also reduce the weight of the product. Although the fast printing speed of this technique supports rapid prototyping, it suffers from low resolution making it difficult to print intricate structures. However, the availability of a wide range of thermoplastic filaments and recently introduced composite filaments with high conductivity makes it an attractive choice for microwave component prototyping and manufacturing. It is also preferred for its ease of use as the FDM printers are relatively easy to operate and the printed products require lesser or simpler processing techniques.

Overall, AM techniques offer immense design flexibility in terms of structural dimensions, material properties, and structure type (planar, 3D or cascaded) which can be modified individually or collectively. From the above discussion, it can be concluded that migrating to AM from PCB techniques not only improves performance but also ensures design freedom and low-cost rapid prototyping. AM also makes it feasible to manufacture structures obtained from optimization algorithms (Zhu et al. 2017), which are otherwise impossible to build with PCB techniques. Printing a complete multi-chip module using multiple AM techniques is also possible. For instance, a multi-layered RF front end multi-chip module may have its nonconductive parts printed with SLA and conductive parts printed with inkjet printing (He et al. 2019). The printed interconnects eliminate free-standing bond wires and provide a mechanical stress buffer for the module which in turn increases the reliability of the device. However, a major challenge to build a whole module using AM is in choosing an appropriate material and printing process. Compared to microwave components manufactured using conventional techniques, AM components are reported to have reduced module weight, a reduction of up to 84% has been reported in the literature (Salim et al. 2018).

AM Material Characterization

The type of material needed to manufacture an RF/microwave component using AM process is dictated by its functionality, operating frequency, deployed environment, and expected mechanical stiffness and strength. The geometric design parameters of a microwave component are determined by the electromagnetic properties of the material, namely, dielectric constant and loss tangent. The thermo-mechanical properties like melting temperature, coefficient of thermal expansion, Young's modulus, and impact resistance aids in material selection by considering the mechanical and environmental factors. For example, components which need high conformability can be manufactured using thermoplastic polyurethane (TPU) which has a small young's modulus indicating high mechanical flexibility. Similarly, for applications where the environment undergoes drastic temperature changes, material with low coefficient of thermal expansion is preferred to preserve structural integrity of the component. RF structures used in pressure variant environments should use materials with high impact resistance to avoid breakage. For AM materials, the mechanical properties are temperature dependent while the electromagnetic properties depend both on temperature and frequency of operation. Thus, the electromagnetic properties of AM materials used to prototype an RF/microwave component need to be characterized at the frequency of operation to ensure the predicted response. Although the AM materials are temperature dependent, their transition temperatures are higher than the standard operating temperature of RF/microwave components. Thus, AM-based RF/microwave components have good temperature stability. In case of power handling capability of AM-based RF/microwave components, the materials with lower loss tangent can be chosen for higher power handling capability.

Estimation of Electrical Properties of AM Materials

Measuring the electromagnetic properties of an isotropic homogeneous material involves computing the complex relative permittivity (ε_r) and complex relative permeability (μ_r). The real part of the complex permittivity determines the dielectric constant, while the loss tangent ($\tan\delta$) is determined as the ratio between the imaginary and real part of the complex permittivity. In case of complex relative permeability, it can only be computed for magnetic materials. While the permeability is close to free space permeability for non-magnetic materials. EM characterization of a material can be done using a non-resonant or resonant method. Depending on the frequency band and accuracy, the characterization method is chosen. Non-resonant method is used for broadband and fairly accurate characterization while a resonant method is preferred when an accurate characterization over a discrete set of frequencies is required.

The transmission/reflection (T/R) line method (Costa et al. 2017) is a nonresonant method which involves the measurement of S-parameters. The permeability and permittivity of the material are then computed from the S-parameters using the

Nicholson-Ross-Weir method. For calculating the S-parameters, a small cuboidal/cylindrical sample of the material is placed inside a two port microwave waveguide/co-axial line. A vector network analyzer (VNA) is then used to measure the S-parameters. Instead of a two-port measurement, the material can also be characterized using single-port measurements (Sahin et al. 2020). Although such T/R line waveguide methods offer EM characterization for broadband frequency response, they have lesser accuracy due to air gap errors. As an alternative, free space measurement approach can also be used to characterize material not only for a broad range of frequencies but also temperatures (Rocha et al. 2013; Singh and Bhattacharyya 2021).

The resonant method for EM material characterization is the resonant cavity method (Chen et al. 2004) where the resonant frequency and Q-factor of the sample are measured. Using the relationship between sample volume, resonant frequency and Q-factor the EM properties of the material are determined. Even though the resonant methods can measure material properties only for discrete frequencies, it is preferred due to its high accuracy and sensitivity. Waveguide cavities are the go to setup for material property estimation using resonant method. With advancement in planar technology the substrate integrated waveguide (SIW) serves as a good substitute for conventional waveguides. An SIW cavity is chosen over a waveguide cavity, as it not only has a high Q-factor but also does not suffer the drawbacks of a conventional waveguide such as, bulkier model, non-integrability with planar technology, and costly manufacturing. Recently, a rectangular SIW cavity with a novel transition offset proposed by Varshney and Akhtar (2021) provided highly accurate EM characterization for an arbitrary material sample. The proposed cavity had a high Q-factor of about 515 and high average sensitivity of nearly 20 MHz. The sample to be measured is inserted in a slot of dimension $4 \times 4 \times 1.57$ mm^3 drilled at the position of the electric field maxima. To ensure a close cavity like enclosure and avoid leakage, the sample thickness is precisely modelled to be equal to the sensor's substrate thickness. Its electrical properties are calculated using the following equations,

$$\varepsilon_s' = \varepsilon_r' \left[A \left(\frac{f_1 - f_2}{f_2} \right) \frac{V_C}{V_S} + 1 \right] \tag{1}$$

$$\varepsilon_s'' = B \left(\frac{1}{Q_2} - \frac{1}{Q_1} \right) \frac{V_C}{V_S} \times \frac{\varepsilon_r'^2 + \varepsilon_r''^2}{\varepsilon_r'} + \frac{\varepsilon_s' \varepsilon_r''}{\varepsilon_r'} \tag{2}$$

$$\tan \delta_s = \varepsilon_s'' / \varepsilon_s' \tag{3}$$

where, ε_s' and ε_r' real part of permittivity of the sample and substrate, A and B are calibration constants determined by testing standard test sample, f_1 and f_2 are resonance frequencies before and after sample insertion, V_C and V_S are Volume of cavity and sample, ε_s'' and ε_r'' are imaginary part of permittivity of the sample and substrate, Q_2 and Q_1 are the unloaded quality factor with and without the test sample, respectively. The study demonstrates the sensors ability to detect loss tangent as low as 0.0003 and distinguishes between materials such as PVC ($\varepsilon_r = 2.65$) and plexiglass ($\varepsilon_r = 2.6$) which have nearly equal dielectric constant.

Table 2 Electrical properties of AM materials

Name	ε_r	loss tangent (tanδ)
Acrylonitrile butadiene styrene (ABS)	2.7–3.5	0.005–0.019
Metalfill	1.2–1.4	0.02–0.2
Hips (polystyrene)	2.4–4.8	0.0001–0.0003
Nylon (polyimide)	3.1–3.55	0.001–0.009
Polylactic acid (PLA)	2.5–3.2	0.005–0.008
Polycarbonate	2.55–3.4	0.009–0.01
Polyetherimide	2.6–3.2	0.0006–0.01
Polyethylene	2.2–2.4	0.0002–0.0004
Polyethethylene terephthalate (PET)	2.5–4.0	0.003–0.025
Polyethylene terephthalate glycol (PETG)	3.0–4.0	0.012–0.017
Polypropylene	2.3–2.6	0.0003–0.0005
Thermoplastic polyurethane (TPU)	3.1–4.2	0.0015–0.0024

Therefore, based on the operating frequency band (wide, narrow, or multi-band) and required accuracy level, it is necessary to choose an appropriate method for EM characterization of the material.

Electrical Properties of AM Materials

An estimate of electrical properties of some commonly used AM materials is summarized (Kronberger and Soboll 2016; Booth et al. 2017; Meyer 2015; Chen et al. 2019; Arbaoui et al. 2015) in Table 2. It should be noted that the values are mentioned in ranges as they are not constant but differ with respect to material manufacturing process, frequency of operation, infill%, and operating temperature.

Frequency Selective Surfaces

Frequency selective surfaces (FSS) are periodic surfaces capable of passing or stopping a certain range of frequencies based on their geometrical and material properties (Munk 2005). These surfaces are assembled using identical elements arranged at regular intervals over a one- or two-dimensional array. When an FSS is excited by an incident plane wave E_i, a part of it is transmitted (E_t) in the forward direction, while the other is reflected (E_r) in the specular direction. The ratio between the reflected and incident wave is determined by the reflection coefficient,

$$\Gamma = \frac{E_r}{E_i}. \tag{4}$$

Similarly, the transmission coefficient τ is the ratio between the transmitted and incident waves,

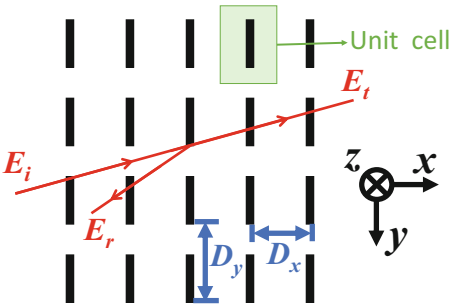

Fig. 4 FSS excited by an incident plane wave E_i

$$\tau = \frac{E_t}{E_i} \tag{5}$$

The part of the surface which repeats periodically along the x and y direction to form the periodic surface is shown in Fig. 4. This repeating unit is termed as the unit cell of the FSS. When a plane wave is incident on the FSS, every unit cell of FSS acquires the same current and fields, except for a phase shift. This property of the periodic surface allows us to estimate the response of the FSS by solving a single unit cell with Floquet boundary conditions.

FSS can be classified into four types based on their structure, namely: plate type, center-connected, loop type, and hybrid. A schematic of these four-unit cell types is shown in Fig. 5. FSS with plate type unit cell are conventional, their band-stop response is due to their patch type structure, while their polarization independence is due to their symmetry. However, plate type FSS suffer from angular sensitivity and early on-set of grating lobes due to the requirement of larger unit cell size. The center connected elements help in achieving frequency response with larger bandwidth and delayed onset of grating lobes. These advantages can be attributed to the geometric properties of the unit cell which permits a close packing resulting in reduced cell periodicity. Moreover, these type of FSS may be used as polarizers because of their polarization dependent response. The frequency response of a loop type FSS primarily depends on the loop length of unit cell. Unlike the other two structure types, it enjoys angular stability and polarization independence. An FSS with hybrid structure incorporates a combination of two or more properties of already discussed unit cell geometries. From this classification, it can be understood that the response of an FSS depends largely on its structure. Conventional FSS using any one of the first three-unit cell types as its repeating unit has a cell periodicity in the order of $\lambda/2$. FSS built using the fourth element type can have element dimensions and cell periodicities much smaller than $\lambda/2$, such surfaces with sub-wavelength unit cell dimensions can be termed as "metasurfaces."

The previous discussion bought out the importance of the unit cell structure in the response of FSS. The material properties like the permeability, permittivity, and conductivity of the dielectric and conductive components also impacts the

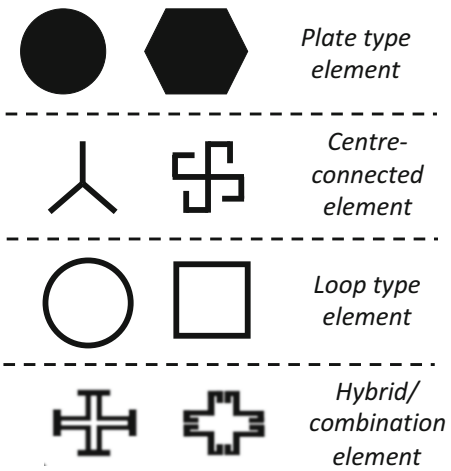

Fig. 5 Structural classification of FSS

frequency response of the FSS significantly. An efficient FSS is expected to have polarization independence, delayed onset of grating lobes, angular and conformal stability. Unit cell miniaturization which is an efficient solution to achieve these properties is possible in three ways: (i) designing planar hybrid elements/metasurfaces, (ii) designing 3D elements, and (iii) designing with left-handed metamaterial (MTM) approach. FSS with planar hybrid unit cell with meandered or convoluted or cascaded structures is an example of miniaturized unit cell, they have low cell periodicity. These miniaturized hybrid FSS are called metasurfaces, if their dielectric layer is very thin ($<<\lambda$). The second approach for cell miniaturization is obtained by transforming a planar FSS into a 3D structure, here the conductive traces have a dominant third dimension compared to the thin conductive traces of a planar FSS, which have a thickness in the order of μm. The third approach uses left-handed metamaterials, these are structures which exhibit negative-ε and μ in the designed frequency of operation. Manufacturing the above discussed FSS with a conventional PCB setup becomes difficult when opted for a cascaded or 3D or structures requiring unconventional dielectric layer properties. This difficulty can be overcome by migrating from standard PCB techniques to AM.

Application Based Design of FSS

When an EM wave interacts with a periodic surface, it can be reflected, transmitted, or absorbed. Based on the resultant behavior, the FSS can be designed to function as a reflector, band-pass filter, or absorber. The following subsections explain the design of FSS for these applications and its prototyping using AM technologies.

FSS as Reflector

An FSS with a patch type structure reflects a set of frequencies based on its geometrical and material properties. Consider an FSS reflector resonant at f_0, when a plane wave is incident on it, the transmitted wave contains all other frequencies except f_0. This functionality of an FSS as a reflector or bandstop filter is shown in Fig. 6. The electromagnetic properties of FSS are understood using its scattering parameters. The reflection bandwidth of the FSS is considered as the range of frequencies for which the transmission coefficient (S_{21}) is less than -10 dB and the reflection coefficient (S_{11}) lies between 0 to -3 dB.

Unlike 3D FSS reflectors, planar FSS reflectors can be manufactured with a PCB process. However, planar FSS reflectors prototyped using PCB process have dielectric layer with fixed structural and material properties. Using AM, dielectric layers with foldable, curved or arbitrary thickness can be manufactured. AM also allows changes in material composition, density, and print fill pattern which in turn changes the material properties of the dielectric layer. A common FSS reflector configuration is a periodic conductive layer etched on a planar dielectric surface, geometrical parameters of the conductive periodic structure and material properties of the dielectric layer determine the frequency to be reflected. These reflector configurations are often used for multi-band and wide-band operations. When used as a multi-band reflector, multiple resonant elements are combined in a single unit cell. For wide-band reflectors, multiple elements with close resonating frequencies are used in combination. Moreover, elements with higher conductive trace width or FSS with lossy or thicker dielectric layers may be used to achieve a wide-band response; these features are easier to manipulate using AM. For instance, a PCB-based FSS reflector has conductive patches etched on a flat dielectric substrate, neither is the dielectric layer bendable after manufacturing nor can copper etching be done on a curved dielectric surface.

Recently, an FSS reflector with curved dielectric layer along with its conductive traces were directly printed (Byford et al. 2018) using AM. The dielectric surface is printed with a high temperature resin using SLA, while the conductive traces were added with aerosol jet printing using a Silver-nanoparticle ink. Such a reflector was

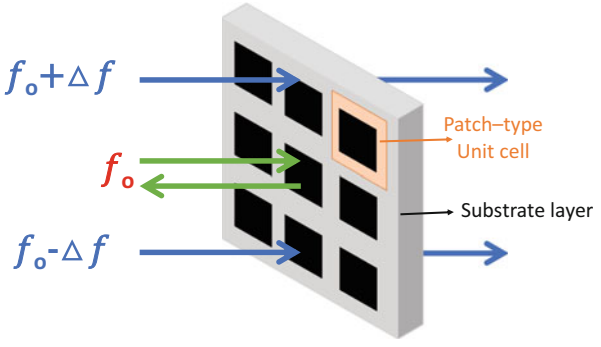

Fig. 6 Illustration of a patch type FSS showing a band-stop characteristic centered at f_0

designed to operate at a high frequency of about 150 GHz. At such high frequencies, minor discontinuities in the conductive traces leads to a large deviation from the predicted frequency response. Even though Byford et al. (2018) used a high resolution aerosol jet printing to realize the conductive traces, the measured frequency response deviated due to the discontinuities in the conductor caused by roughness in the curved dielectric surface. This problem could be avoided if a buffer layer had been added before printing the conductive traces. However, this problem can also be mitigated by adding an additional copper layer of 2.5 μm by electro-less plating.

Apart from wide-band and multi-band frequency response, band tunability and conformability are much needed features for the present generation RF and microwave systems. PCB-based tunable FSS generally achieve their tunability by adding additional surface mount components (SMC) on to the conductive traces. These FSS are generally prototyped on a rectilinear dielectric substrate. Such tunable FSS are planar and cannot be bent to conform a non-planar surface. Moreover, these FSS are heavy and require manufacturing with high cost and maintenance. On the other hand, an AM-based tunable FSS reflector can have a very thin dielectric layer. These dielectric layers are often manufactured using Polyethylene terephthalate (PET) which is flexible. The conductive traces are added on this flexible substrate using screen printing technique. Similar to a PCB-based tunable FSS, the flexible AM-based FSS also uses SMC to achieve tunability. Tian et al. (2020) demonstrated a flexible tunable FSS using a voltage-biased varactor by attaching it to the FSS with conductive adhesives. They also showed that by varying the bias volt age from 0 to 20 V, the frequency response of the FSS reflector can be tuned between the frequency range of 1 to 6 GHz. Such FSS reflectors also has a stable response when tested by conforming it to a curved surface from various angles of incidence. A novel approach to manufacture tunable FSS without SMC is to use origami-based structure. Prototyping such foldable structures is possible with AM but not with a standard PCB process. Unlike SMC-based reflectors, the foldable design gives a continuous tunable range. Such origami-based reflectors have flexible conductive and dielectric layer. The foldable dielectric layer can be printed with an elastomer using SLA technique and the conductor traces can be added to it with silver nanoparticle ink using inkjet printing. Discontinuities in the conductive layer due to surface folds can cause deviations from the predicted frequency response. These discontinuities arise due to the lower resolution of the dielectric layer compared to the nanoparticle conductive layer. A buffer layer printed between the dielectric and conductive layer eliminates these discontinuities. An origami FSS prototyped by Cui et al. (2019) had an angularly stable response in wide-band tuning frequency range of 21 to 26 GHz.

The planar FSS reflectors discussed above are manufactured with a two-step process where the dielectric and conductive parts were built with two different processes. Although SLA support complex structures, smoother finish, and materials with large permittivity, FDM is preferred over SLA in terms of cost, material choices, and single-step manufacturing. A FDM printer with two nozzles can seamlessly print both the conductive and dielectric part of a reflector. Mirzavand et al. (2019) demonstrated the use of a single step FDM process in manufacturing an

FSS reflector. The dielectric layer was printed using nylon while the conductive layer was printed using conductive continuous carbon-fiber-reinforced thermoplastic (CFRTP). Unlike the metallic traces made of silver or copper, CFRTP is not a perfect conductor which results in a deviation in the frequency response. However, it was observed that this problem can be overcome by increasing the trace thickness which in turn improves the conductivity. This shows the design freedom available in AM to modify both the substrate and conductive layer in terms of geometry and material properties. Such design freedom and low-cost prototyping is not possible with the PCB process.

3D structured FSS are preferred over planar ones due to their high angular stability. Unlike PCB process, AM techniques can manufacture such 3D structures with ease. For 3D structured FSS reflectors SLS process can be preferred due to its ability to build very complex structures with precision, its large material library also include metals. The center-connected and loop type 2D FSS structures can be modified into a 3D FSS reflectors by extruding the planar conductive structure. These 3D conductive structures need not be fully metallic, a thin layer of metallic paint applied on a non-metallic 3D structure would produce the same frequency response as its fully metallic counterpart. Often two layers of silver paint is coated on the 3D structure to ensure conductivity. A good angular stability can be achieved by such 3D structures; however, it can be further improved by arranging these 3D structures in a triangular lattice (Sanz-Izquierdo and Parker 2013, 2014b). For low-cost rapid prototyping, FDM technique can be used (Sanz-Izquierdo and Parker 2014a, 2015), like their SLS counterparts, these low-cost prototypes are printed with PLA and coated with two layers of silver paint. These reflectors also show miniaturized response and improved angular stability.

Even though the above-mentioned 3D reflectors may be manufactured using AM techniques, they require a two-step process due to the presence of dielectric and conductive part. An FSS reflector made fully with dielectric material can eliminate this two-step manufacturing process. These all-dielectric reflectors work on the principle of dielectric resonator. An all-dielectric FSS reflector can be designed by extruding planar structure like circle, square, or hexagon. These FSS reflectors can be made conformal by prototyping them using a flexible material. A conformal all-dielectric FSS reflector reported by Mellita et al. (2021) uses FDM technique. The conformability of the reflector is achieved by prototyping it with TPU, a flexible material. Hexagonal protrusions arranged in the form of a hexagonal lattice is used to achieve a reflection bandwidth ranging from 14.7 to 15.4 GHz. These 3D FSS reflectors with protrusions and flexibility are impossible to manufacture using PCB process.

FSS as Band-Pass Filter

An FSS with a slot type structure transmits/passes a set of frequencies depending on its geometrical and material parameters. A plane wave incident on a band-pass FSS resonant at f_0, transmits a wave with frequency f_0, while reflecting all other frequencies. This functionality of FSS as band-pass filter is shown in Fig. 7. The range of

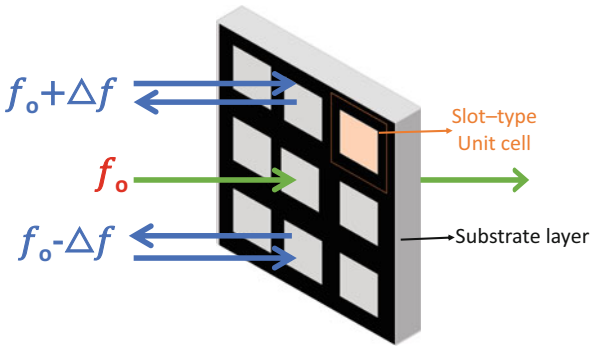

Fig. 7 Illustration of a slot type FSS showing a band-pass characteristic centered at f_0

frequencies for which the transmission coefficient (S_{21}) lies between 0 to −3 dB and the reflection coefficient (S_{11}) less than −10 dB is considered as pass-band.

As mentioned earlier, an FSS with miniaturized unit cell has higher angular stability and delayed onset of grating lobes. A conventional design approach for cell miniaturization of planar FSS is to use meander line structures. However, further miniaturization can be achieved by transforming such planar FSS into a 3D structured FSS. Extruding the conductive and dielectric layers of a planar FSS is a simple methodology to obtain 3D structured FSS.

A planar band-pass FSS proposed by Ghosh and Lim (2019) has a miniaturized unit cell with periodic conductive slots on either side of the dielectric layer. The conductive layer on the top has a cell-interconnected meander line structure while the bottom side has an inter-digital capacitor structure. This planar FSS extruded on either side by 0.6 mm achieves an additional miniaturization of 15%. The dielectric layer of this 3D structured FSS was printed with PLA using FDM technique while the conductive layer was added with a conductive spray using a negative mask. A band-pass FSS prototyped by Tehrani et al. (2017) demonstrates a proof-of-concept encapsulation for SoP-based wireless IC packages using AM. The reported band-pass FSS has a dielectric layer printed with an acrylate resin using SLA technique while the conductive layer was added with inkjet printing. The smooth finish of the planar substrate with SLA process eliminates the need for an intermediary passivization film before conductive coating. Thus, AM technologies with its vertically integrated fabrication approach for multi-material 3D structures and 2D topologies makes it an attractive option for prototyping SoP systems.

FSS as Absorbers

A conventional absorber is a Salisbury screen which has two layers, a resistive and metallic layer. The absorbing frequency f_0 is achieved by maintaining a separation of $\lambda/4$ between the two layers as shown in Fig. 8a (Fante and McCormack 1988). When an EM wave is incident on the lossy resistive layer, it passes through it and gets

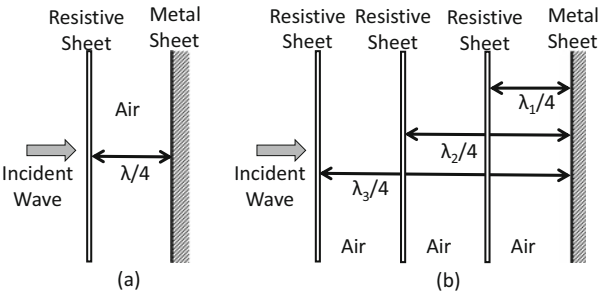

Fig. 8 Conventional absorbers (**a**) Salisbury screen (**b**) Jaumann absorber

reflected from the metallic layer. This reflected EM wave undergoes a phase shift of 180° as it reaches the lossy layer causing a destructive interference thus enabling absorption.

Jaumann absorber used for multi-band/wideband absorption uses the same principle of the Salisbury screen, whereas instead of one resistive layer, multiple resistive layers are added at different separations corresponding to the required absorption frequencies. A schematic of a Jaumann absorber is shown in Fig. 8b. These simplistic structures perform well for narrow band absorber but the requirement of $\lambda/4$ separation becomes a drawback for current applications which demand ultrathin wideband and multi-band absorbers. Achieving a wideband absorption without $\lambda/4$ separation is possible with planar FSS absorbers. These absorbers have an FSS etched on a grounded dielectric substrate. The FSS acts as a broadband impedance match between the grounded substrate layer and free space, enabling wideband absorption. The reactance offered by the FSS and grounded substrate layer reduces the required absorber thickness eliminating the need for $\lambda/4$ separation. An FSS absorber incident with a plane wave absorbs the wave corresponding to frequency of absorption f_0, while reflecting all the other frequencies. This FSS is designed to have an input impedance equal to free-space impedance (η_0) at the frequency of absorption. The functionality of an FSS absorber is illustrated in Fig. 9. However, due to the fully metallic ground there is zero transmission of EM waves, i.e., the transmission coefficient (S_{21}) is zero. The absorptivity ($A(\omega)$) and input impedance ($Z(\omega)$) of the FSS absorber can be estimated from the reflection coefficient (S_{11}) using the following equations (Smith et al. 2005),

$$A(\omega) = 1 - |S_{11}(\omega)|^2 \tag{6}$$

$$Z(\omega) = \eta_0 \frac{1 + S_{11}}{1 - S_{11}} \tag{7}$$

Requirements of a good EM absorber are wide angular stability, polarization independence, broader absorption bandwidth, thinner structure, and low density. However, designing an absorber with all these properties is a challenge. A

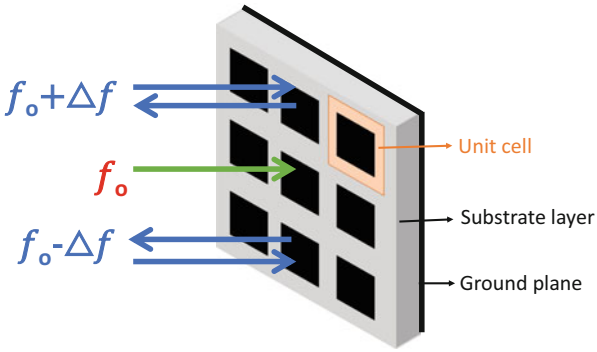

Fig. 9 Illustration of a grounded FSS showing an absorptive characteristic centered at f_0

metamaterial design approach can be used to build absorbers satisfying most of these requirements. The term metamaterial (MTM) was coined by Dr. Rodger Walser; these are manufactured materials with extraordinary electromagnetic properties which cannot be found naturally (Balanis 2012). This concept of MTM can be used in two ways to design FSS absorbers with better performance. The first approach is to modify the structural geometry of the FSS into a sub-wavelength 3D structure. While the second approach enhances the material properties of the absorber by introducing certain highly conductive or resistive impurities. The absorbers built using these approaches are generally termed as MTM-FSS absorbers. However, manufacturing such 3D structured or composite material absorbers is impossible with PCB process. With the advancement in AM technologies complex, FSS absorbers which were otherwise non-realizable with standard PCB techniques are now a possibility.

An FSS absorber with a conventional Jerusalem cross (JC) structure can be miniaturized by transforming it into a 3D structure to obtain an MTM-FSS absorber. The 3D structure is a conductive JC printed on a staircase-like substrate. The staircase projections of the substrate layer enable the conductive JC element to have a longer electrical length without changing the footprint of the unit cell, resulting in a miniaturized cell structure. Such a 3D structured absorber has been prototyped and tested by Lim et al. (2018). The substrate is built with PLA using FDM technique, the conductive JC elements were painted on it using an ELCOAT silver paste and the ground plane is realized by pasting a copper tape at the bottom. This 3D absorber can be further miniaturized by meandering the JC structure on the staircase-like substrate layer. The meandering causes further increase in electrical length of the JC without change in unit cell periodicity resulting in an improved cell miniaturization. This idea was prototyped by Kim et al. (2020) using the same manufacturing technique as the former. From this study, it can be concluded that the meandered absorber with its highly miniaturized unit cell has an improved angular stability. Absorbers prototyped by both Lim et al. (2018) and Kim et al. (2020) used manual painting to realize the conductive element. However, a single step manufacturing where both the conductive and dielectric layers use the same

manufacturing process is preferable to avoid the strenuous and error prone manual painting. An FSS absorber can be built using a single AM technique by designing a cohesive structure using two different FDM materials. An FSS absorber proposed by Prince et al. (2021) uses single step FDM process to prototype its dielectric and conductive parts with solid PLA and graphite-PLA, respectively. PLA composite containing graphite has good electrical conductivity making it an ideal choice to manufacture the conductive layer. Inspite of the FDM machine tolerance, the performance of this absorber was unaffected due to its lower frequency of operation (about 13 GHz). Absorbers with higher frequency of operation should preferably use other AM techniques with higher resolution to have a stable frequency response. A multi-layered MTM-FSS absorber can also be prototyped using single step FDM process. Such an absorber proposed by Lleshi et al. (2019) has alternating metal and dielectric layers arranged in a truncated pyramidal fashion. The dielectric and conductive layers of this MTM-FSS absorber were realized with ABS and Electrifi, respectively. The Electrifi filament with its high conductivity of 17000S/m is a good substitute for the conventional copper conductive layer. Compared with a PCB manufactured multilayer absorber, the AM-based absorber exhibits better stability and a flat absorption response even in a wide frequency range (7.6 to 17.4 GHz). The unstable response of PCB-based prototype is due to its inability to manufacture a multi-layered structure with pyramidal cross-section. These absorbers prototyped with single-step FDM process were built by integrating two FDM materials. Nonetheless, due to low FDM machine tolerance such structures needing two materials are not preferable for high frequency ultra-wideband response. Designing a 3D absorber with a single material can eliminate the need of such two materials integrated printing. However, obtaining wideband absorption with such single material absorbers is a challenge. These absorbers can be modified into MTM-FSS absorber which have exceptional response by manufacturing them with composite materials with enhanced EM properties. An FSS absorber with simple cuboidal projections can be designed to have a wideband absorption by using a composite material. Kronberger and Soboll (2016) proposes the use of a composite material like Bronzefill to prototype such absorbers. Bronze-fill is an FDM composite which contains bronze and PLA in the ratio of 80:20. The bronze metal filler in the PLA material enhances the EM properties of the PLA composite enabling better absorptivity. Other PLA composite containing graphite can also be used to prototype such absorbers (Kjelgård et al. 2018). A study by Wu et al. (2020) aimed at developing an FDM composite to attain good absorptivity. They studied a composite with graphene, nano Fe_3O_4, and polylactic acid (PLA) as its constituents. After experimenting on various compositions, it was determined that an absorber made with a composite containing 5 wt% in graphene and 20 wt% in nanosize Fe_3O_4 had an average dielectric constant of 12 and produced a synergistic absorbing effect in a wide frequency range (about 8.2 to 18 GHz). Polymer composites with enhanced electrical conductivity can be prepared using a spatial confining forced network assembly method (Wu et al. 2017). A patterned conductive plastic (PCP) manufactured with this method can be used to prototype wideband MTM-FSS absorbers. One such absorber proposed by Lu et al. (2018) uses circular patches of

PCP. The absorber is manufactured by embedding these circular PCP patches within a PLA block while an adhesive copper tape is applied at the bottom to ensure zero transmission. This compact metamaterial absorber had an exceptionally large bandwidth (ranging from 16.3 to 54.3 GHz) with angular stability and polarization independence.

SLS technique is preferred over FDM for prototyping FSS absorbers with intricate 3D structures as it supports high resolution printing. An FSS absorber with an intricate 3D structure and cell periodicity as small as 5 mm can be printed with good resolution using SLS technique. One such absorber with a multi-layered staircase structure was prototyped by Zhou et al. (2016) with a powdered SLS composite containing nylon and CIP in the ratio of 30:70. Another SLS composite with nylon and graphite was studied by Abdullahi and Ali (2021) to build a three-layered metastructure absorber. Both the absorbers are angularly stable and polarization independent for the entire absorption bandwidth.

Dielectric resonator is another approach used to attain wide-band absorption; these absorbers use 3D resonator structures which are easier to build with AM technology. A cylindrical dielectric resonator for wideband applications was proposed by Ren and Yin (2018a). This resonator type absorber was printed with carbon loaded ABS (composite ABS) using FDM technique. Higher the dielectric loss of the dielectric resonator wider the absorption bandwidth. Water which has high dielectric loss in the microwave frequencies is suitable for design of wide-band EM absorbers. Though water has high dielectric loss, it cannot be used directly as an EM absorber due to the impedance mismatch which occurs at the water-air interface. The water-resonator based absorber proposed by Ren and Yin (2018b) overcomes this problem. The absorber has FDM printed PMMA layer which has a copper foil on one side and a resin cap with a cylindrical cavity on its other side. The cavity has microchannels which facilitates the supply of water into this resonator cavity. The cylindrical resin cap overcomes the impedance mismatch problem and the copper foil ensures zero transmission. When this absorber is injected with water, it has high absorptivity for a wide frequency range (about 5 to 25 GHz). FSS absorbers with perforated substrate layer can be used for applications which require light-weight absorbers. A honeycomb-structured substrate layer, though perforated, has optimal mechanical sturdity. A wideband absorber proposed by Jiang et al. (2018) uses a honeycomb structured substrate layer made of PLA using FDM technique. The walls of this substrate layer are covered with screen printed resistive patches for improved absorptivity, while a metal backboard is added to ensure zero transmission. Even though these absorbers are dielectric models, due to their structure, they have a much better mechanical performance compared to a metallic lattice design. A similar honeycomb structured layer made of PLA was proposed by Ghosh and Lim (2018) to design a broadband MTM absorber. Instead of covering the walls of the honeycomb structure with resistive patches, two square loops were added on the substrate for broadband absorptivity. The square loops were realized by painting silver ink via a negative mask. A study by Lai et al. (2020) compares the absorptivity of honeycomb and woodpile structured FSS absorber made from composite material containing carbon fiber, conductive material, and Acrylonitrile Butadiene Styrene

(ABS). They found that the woodpile structure had wider absorptivity, demonstrating the significance of micro-structures in improving EM absorbing properties.

As mentioned earlier, the properties of an EM absorber can be enhanced either by modifying its geometric configuration or changing the chemical composition of the material. An absorber where both the geometrical and material aspects were modified to attain enhanced absorptivity was proposed by Liu et al. (2019). The dielectric absorber had a tower-like unit cell geometry with a sharp tip. Such a structure provides the incoming EM wave with a gradual transition from free space into the lossy material which minimizes reflectivity and improves absorptivity over the length of the geometry. The proposed tower-like structure dissipates the incoming EM wave by multiple reflections and intrinsic loss to obtain optimal impedance matching, resulting in a better absorption performance. This absorber was prototyped with an FDM composite containing PLA and carbonyl iron particles (CIPs) in a ratio of 50:50. Similarly, a honeycomb-structured absorber with sharp projections was proposed by Quan et al. (2021). They used SLA technique for prototyping this absorber with a composite resin containing carbon nanotubes (CNTs). Both these absorbers have achieved enhanced absorptivity because of the combined microscopic dielectric and macroscopic geometry-induced losses.

Conclusion

This book chapter discusses the use of AM for prototyping better performing FSS which were designed using MTM approach. This approach included modifying the structural and material aspects of the FSS to obtain exceptional EM performance. The role of AM in manufacturing such FSS is reviewed with emphasis on its benefits over conventional PCB manufacturing process. Prominent AM techniques like SLA, SLS, and FDM were compared and the benefits provided by each of them for FSS prototyping were highlighted. The importance of material characterization and the methods to achieve them were discussed. The functionalities of FSS as a reflector, band-pass filter, and absorber were explored. AM-based FSS prototypes implementing these functionalities were reviewed, it was found that MTM type FSS combined with AM techniques had enabled the design and prototyping of several compact, wideband FSS with superior performance. Though AM techniques are advantageous and useful in prototyping complex structures with wide range of materials, they still suffer from a few drawbacks. Major challenges of AM in prototyping RF/microwave components is the availability of low-cost AM materials with suitable EM properties; these include materials with high conductivity and resistivity. However, various conductive AM materials have been introduced recently in the commercial market. For example, conductive filaments like electrifi, copperfill, and bronzefill have been used in FDM technology. Whereas various grades of metal alloy powders and polymer resins with conductive fillers have been used in SLS and SLA technologies, respectively. Such advancements in AM technologies project a promising future for prototyping low-cost, better performing RF/microwave components.

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