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Advances and Applications of Metamaterials and Metasurfaces Through RF to Optic Spectrum

Edited by Mohammad Alibakhshikenari



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IntechOpen Book Series

Materials Science

Volume 24

Aims and Scope of the Series

Materials science has always occupied an extremely high position in the human development process. As we explore the oceans of stars, various industries have put forward more stringent requirements for the performance of materials, forcing us to pay more and more attention to the development of new materials. At the same time, the formation of a data-driven scientific paradigm is dramatically shortening the development cycle of new materials. The huge data generated by synergistically combining theories, high-throughput experiments, high-throughput computation, and artificial intelligence is greatly contributing to our ability to utilize materials science to solve real-world problems. The three topics of this book series - Metals and Nonmetals; Composite Materials; and Surface Science - will address important areas of advancement in materials science. There will be a range of interesting works published under these topics.

Meet the Series Editor



Prof. Chonghe Li received his Ph.D. from the Chinese Academy of Sciences in 1995. From 1995 to 2000, he worked as a researcher at the Shanghai Institute of Metallurgy, Chinese Academy of Sciences, where he also served as director of the research laboratory. In 2000, he was appointed Professor at the Institute of High-Performance Computing in Singapore, where he worked on computation and simulation in materials science until 2004. Since then, he has been a professor at the School of Materials Science and Engineering, Shanghai University, China, as well as the director of the Shanghai Specialty Casting Engineering and Technology Research Center. Prof. Li's research focuses on titanium alloys, titanium-aluminum single crystals, intermetallic compounds, theoretical calculations, alloy design, and special refractory materials. His broad scientific expertise is well recognized by the scientific community around the world. He is a member of the editorial board of the journal *Metals*. As an author, he has published more than 200 peer-reviewed papers, 2 books, and over 40 patents.

Meet the Volume Editor



Mohammad Alibakhshikenari received the Ph.D. degree with the European label from the University of Rome “Tor Vergata” (UNITOV), Italy, in February 2020. He is now a Marie Skłodowska-Curie, ROSETTA, and LERO (the Research Ireland Centre for Software) Fellow at the University of Galway, Ireland. From September 2024 to June 2025, he was a Senior Researcher at UNITOV. From July 2021 to August 2024, he was with the Department of Signal Theory and Communications at Universidad Carlos III de Madrid (UC3M), Spain, as a Marie Skłodowska-Curie and CONEX-Plus Fellow. For the 2021/2022 academic year, he received the “Teaching Excellence Acknowledgement” certificate for the course “electromagnetic fields” from the Vice-Rector of Studies at UC3M. In November 2019, he was awarded a two-year postdoctoral research fund by UNITOV. His research interests include “Electromagnetic fields, antennas and wave-propagations, metamaterials and metasurfaces, RF and microwave technologies, wireless communications, millimeter-waves and terahertz circuits”. According to Stanford University’s yearly analysis, he has been among the World’s Top 2% Scientists of highly cited scientific authors since 2020.

Contents

Preface	XIII
Chapter 1	1
Metasurfaces for Structured Terahertz Waves: Generation, Control and Applications <i>by Wei Chen, Zhong-Ze Peng, Ling-Ling Ma and Yan-Qing Lu</i>	
Chapter 2	31
Metasurface-Based Reconfigurable Antennas and Intelligent Reflecting Surfaces for B5G and 6G Applications <i>by Dilek Uzer and Ayşe Özge Çınar Ak</i>	
Chapter 3	69
Metasurfaces for Applications of Multi-Dimensional Orbital Angular Momentum Mode (De)multiplexing <i>by Qingji Zeng and Shuqing Chen</i>	
Chapter 4	89
Radar-Absorbing Materials, Metamaterials, and Metasurfaces: Design Methods, Considerations, and Best Practices <i>by Mohamed Edries, Hesham A. Mohamed, Dalia N. Elsheakh and Mohamed A. El-morsy</i>	
Chapter 5	131
Bandgap Structure in the Terahertz Region for Reflecting Electromagnetic Waves: A Theoretical Approach Considering Dispersion <i>by Jesús Valenzuela-Sau, Raúl García-Llamas and Jorge Gaspar-Armenta</i>	
Chapter 6	147
Suppress Mutual Coupling in MIMO Antennas Based on Artificial Electromagnetic Structures <i>by Ijaz Khan, Yizhi Shao, Mian Muhammad Kamal, Chaoyun Song, Tianjun Ma and Mohammad Alibakhshikenari</i>	

Preface

Metamaterials (MTMs) and Metasurfaces (MTSs) have become an inviting research area, widely used in a variety of high-frequency communications and other applications for manipulating electromagnetic waves. These structures, typically designed to be as thin as sub-wavelengths, offer advantages such as easy manufacturing with small volumes and the ability to control wave propagation around the structures on which they are mounted. From shaping light beams to producing ultrathin lenses and enabling future communication devices, MTSs fascinate both researchers and engineers with their potential to transform industries. In addition to massive wafers, MTSs must accommodate a wide range of surface options, meet diverse practical requirements, and be compatible with various devices. Consequently, creating a soft MTS that can bend or stretch is one of the hottest new research areas.

The book summarizes recent advances in THz metasurfaces for structured beams along four themes: high-efficiency beam-shaping mechanisms; control strategies using tunable materials and reconfigurable architectures; multifunctional integration incorporating frequency, polarization, and mode multiplexing; and applications in communication, imaging, and high-field emission. We discuss the future directions for scalable, tunable, and efficient platforms to facilitate chip-to-free-space interfaces, metrology, and manufacturable implementation. Two noteworthy areas of interest are ferroelectric nematic liquid-crystal platforms and spatiotemporally structured THz waves, which are also discussed.

Today's wireless communication systems increasingly require antennas with versatile functionality, adaptability, and flexibility. Unlike conventional antennas, reconfigurable antennas can instantly change their operating characteristics, such as frequency, radiation pattern, polarization, and their combinations, to meet specific needs. This feature provides a significant advantage for next-generation wireless communication applications. While data rates and communication capacity can be significantly increased in beyond-5G and 5G cellular communication systems, electromagnetic (EM) wave propagation faces greater atmospheric attenuation. Reconfigurable Intelligent Surfaces (RISs) play a crucial role in maintaining communication connectivity. This book discusses metasurface-based applications and Intelligent reflective surface (IRS) applications, also known as reflective metasurfaces (RMTSs).

The advancement of metasurface-enabled multi-dimensional orbital angular momentum (OAM) mode (de)multiplexing has emerged as a critical solution to address the soaring capacity demands of optical communication systems in the big data era. Metasurfaces, with their subwavelength configurability and multi-parameter manipulation capabilities, serve as powerful platforms for engineering multi-dimensional OAM modes. Hence, this book systematically studies three core metasurface-based architectures that capitalize on subwavelength programmability and multi-degree-of-freedom light manipulation to overcome the intrinsic limitations of

conventional optics – including single-mode operation, Bragg diffraction constraints, and the inability to jointly process OAM, polarization, and wavelength dimensions. For each scenario, the fundamental principles, design methodologies, optical demonstrations, and performance trade-offs are discussed. Accordingly, this book highlights the pivotal role that metasurfaces are poised to play in shaping the development of next-generation high-speed, large-capacity all-optical communication networks.

Radar-absorbing materials (RAMs) are crucial for reducing radar cross-section (RCS), suppressing electromagnetic interference (EMI), and improving the efficiency of antennas and wireless systems. Several design approaches have been established, ranging from conventional dielectric and magnetic composites to multilayer structures, frequency-selective surfaces (FSS), and metamaterial-based absorbers. Each method provides specific advantages, such as wide bandwidth, compactness, or tunability, but also poses challenges, including structural thickness, fabrication complexity, or polarization dependence. More recent advances involve graphene- and metasurface-based absorbers, which offer ultra-thin, lightweight, and reconfigurable solutions suitable for emerging 5G/6G communication platforms, stealth applications, and multifunctional devices. Hence, this book presents a comprehensive overview of RAM design methods, key considerations, and best practices, providing a structured reference for researchers and engineers aiming to develop next-generation RAMs for defense, aerospace, and modern communication systems.

The dielectric constant plays a crucial role in the study and development of technologies in which electromagnetic waves interact with matter during operation. Well-known examples of this are optical fibers, lenses, radars, and X-ray machines. More recently, metamaterials and metasurfaces have demonstrated innovative properties for possible applications across a wide range of the electromagnetic spectrum. This book provides a comprehensive study of a bandgap structure in the terahertz region, considering the universal dispersive nature of the dielectric constant as a function of electromagnetic wave frequency. In the theory section, a basic approach to the Plane Wave Expansion Method is presented. First, the electromagnetic wave equation is obtained; then mathematical functions for the dielectric constant and the electromagnetic field are defined; later, the general equations for one-, two-, and three-dimensional Photonic crystals (PCs) are derived. In the results section, the band structures from 0 to 30 THz of a honeycomb lattice of circular cross-section alumina rods in vacuum, considering experimental values for its dielectric constant, are presented, indicating notable behaviors due to unusual values of the dielectric constant. The behavior in both band structures is discussed in function of the dispersive dielectric constant.

As modern wireless platforms from sub-6 GHz to WLAN and mmWave 5G continue to demand higher data rates, greater spectral efficiency, and compact integration, the effective decoupling techniques have become essential. This book provides a comprehensive overview of mutual coupling suppression in multiple-input multiple-output (MIMO) antenna systems using artificial electromagnetic structures (AEMs). It establishes the electromagnetic and mathematical foundations for analyzing mutual coupling (MC), including impedance and scattering matrices, far-field correlation integrals, and surface-wave models. AEM structures such as frequency-

selective surfaces (FSS), metamaterials, parasitic resonators, neutralization lines, and defected ground structures (DGS) are shown to be highly effective at controlling these paths through engineered electromagnetic responses. They have demonstrated significant improvements in isolation and overall system performance across multiple bands and environments, supporting the continued evolution of next-generation wireless systems.

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Metasurfaces for Structured Terahertz Waves: Generation, Control and Applications

Wei Chen, Zhong-Ze Peng, Ling-Ling Ma and Yan-Qing Lu

Abstract

Terahertz (THz) structured fields, such as vortex beams carrying orbital angular momentum, Bessel beams, and vector beams, support mode multiplexing, extended depth of focus, and improved scattering tolerance in communication, sensing, and imaging. Metasurfaces, consisting of ultrathin subwavelength arrays, employ metallic or all-dielectric building blocks to enable wavefront manipulation through geometric-phase or resonance mechanisms. Building on these foundations, subsequent research has sought to further enhance functionality and efficiency by developing multilayer configurations that integrate multiple operations and support more complex modulation. In parallel, tunable materials – such as phase-change compounds, liquid crystals, and graphene, whose properties respond to external physical stimuli – have been incorporated to dynamically control optical responses. These approaches enable planar, programmable manipulation of phase, amplitude, and polarization in the THz regime while reducing losses and device footprint, thereby providing reconfigurability beyond that of conventional bulk optics. The chapter summarizes recent advances in THz metasurfaces for structured beams along four themes: high-efficiency beam-shaping mechanisms; control strategies using tunable materials and reconfigurable architectures; multifunctional integration incorporating frequency, polarization, and mode multiplexing; and applications in communication, imaging, and high-field emission. We discuss the future directions for scalable, tunable, and efficient platforms to facilitate chip-to-free-space interfaces, metrology, and manufacturable implementation. Two noteworthy areas of interest are ferroelectric nematic liquid-crystal platforms and spatiotemporally structured THz waves, which are also discussed.

Keywords: terahertz metasurfaces, structured beams, orbital angular momentum, tunable and programmable control, multifunction integration

1. Introduction

Terahertz (THz) radiation (0.1–10 THz) occupies the spectral gap between microwaves and the infrared. It combines an underexploited bandwidth, low photon energy, and strong penetration through most nonmetallic materials and is widely

viewed as a bridge between electronic and photonic technologies [1–3]. At the application level, THz waves address spectral fingerprints of molecular rotations and lattice vibrations, enabling chemically selective biosensing and nondestructive quality inspection [4–8]. At the information level, tens to hundreds of gigahertz of contiguous bandwidth provide the physical basis for Tb/s-class wireless links and short-reach backhaul in 6 G/6 G + [9, 10]. At the physics level, single-cycle or broadband THz pulses drive strong-field nonlinearities and ultrafast dynamics of electronic, lattice, and spin degrees of freedom, opening a key window between classical electromagnetism and quantum material responses [11]. However, conventional Gaussian beams limit capacity, robustness, and wave–matter interaction strength: limited spatial modes and polarization degrees of freedom, together with sensitivity to scattering and turbulence, hamper stable high-dimensional transport in complex links and heterogeneous samples.

Structured THz fields offer three complementary degrees of freedom. First, vortex beams carrying orbital angular momentum (OAM) form a near-orthogonal eigenmode set under suitable aperture and alignment; modes with distinct topological charge l can be multiplexed at a single frequency and polarization to extend channel count in spectrum-limited links [12, 13]. Second, nondiffracting and self-accelerating beams, such as Bessel and Airy, provide extended depth of field (DOF) and reduced sensitivity to partial occlusion and diffraction, aiding imaging and detection on curved or scattering surfaces [14–16]. Third, vector-polarized fields – radial, azimuthal, and generalized Poincaré states – enable tailored transverse and longitudinal components, improving coupling for chiral recognition, anisotropic material characterization, and micro/nano manipulation [17, 18]. These features address, respectively, mode capacity, propagation robustness, and field–matter coupling, and can be combined in system design as permitted by mode purity, alignment tolerance, scattering strength, and device bandwidth.

Realizing this potential at THz frequencies requires addressing gaps in material response. Natural media offer limited tunable dispersion at sub-millimeter wavelengths; bulk optics are large, lossy, and constrained in phase continuity and dispersion control; and tunable devices (phase shifters, spatial light modulators) remain immature in the THz range, making it difficult to co-optimize refresh rate, phase accuracy, and aperture. Atmospheric absorption and multipath scattering further degrade links, complicating the simultaneous pursuit of high efficiency, broad bandwidth, and programmability. Progress must proceed in concert across materials, structures, and systems: low-loss and tunable materials; subwavelength phase and mode engineering; and programmable, algorithm-assisted optimization at the system level.

Metasurfaces built from subwavelength meta-atoms have therefore emerged as a key platform for THz wavefront engineering [19, 20]. By co-designing Pancharatnam–Berry (PB) geometric phase, resonance-based phase, and higher-order multipolar interference, metasurfaces can simultaneously manipulate amplitude, phase, polarization, and OAM within subwavelength thickness, enabling planar, full-parameter control across free-space and surface waves, scalar and vector fields, and single-mode and multiplexed operation [21, 22]. Compared with bulk components, metasurfaces offer three system-level advantages: (i) ultrathin yet high-phase-coverage – achieving 2π and even multi-turn phase coverage with electromagnetic thickness far below the wavelength; (ii) compatibility with multiple actuation mechanisms – readily integrated with phase-change media, two-dimensional

materials, liquid crystals (LCs), and semiconductors for thermal, electrical, optical, and magnetic actuation; and (iii) programmable and manufacturable – supporting rapid wavefront customization via discrete/continuous geometric encoding, and scalable production and on-chip integration via 3D printing and CMOS-compatible processes [19, 20].

Building on this base, structured THz fields have moved from static to active control, with early signs of intelligent operation. Initial studies used passive geometric phase and metallic or dielectric resonators to impose fixed wavefronts, enabling efficient generation of vortices, Bessel beams, and vector fields [23–26]. Tunable media, such as graphene and liquid crystals (LCs), then introduced thermal, electrical, optical, and magnetic bias for phase reconfiguration, mode switching, and multi-channel multiplexing [27–30]. In parallel, nonlinear generation shifted wavefront design upstream to the near-infrared: optical rectification (OR) and difference-frequency generation (DFG) transfer near-infrared (NIR) phase and polarization coding to THz with high fidelity, linking emission and control [31]. More recently, topological photonics and non-Hermitian phase engineering have provided disturbance-tolerant control of chirality and phase in the THz, pointing to higher-purity modes, multimode transport, and more robust propagation [32].

In line with this logic, the chapter moves from underlying mechanisms to control approaches, then to system design, and finally to multifunctionality and applications. Section 2 addresses high-efficiency generation of structured THz waves under material and dispersion constraints. Section 3 covers on-demand control of phase, polarization, and modes using tunable materials and programmable architectures. Section 4 discusses multifunctional integration and system-level design spanning multi-channel and polarization multiplexing. Section 5 interfaces these capabilities with forefront applications in communications, imaging, and strong-field science. Section 6 outlines coordinated strategies and standardization toward the concurrent goals of high efficiency, broad bandwidth, and programmability. Our aim is to unify dispersed device-level advances into a clear mechanism–function–application map, providing actionable pathways from principle to engineering for structured THz fields using metasurfaces.

2. Generation mechanisms

Structured THz fields have progressed along two complementary tracks. The first is passive wavefront engineering, which achieves high efficiency and high mode purity without time-varying drives by using static metasurfaces or photonic crystals. The second is active and programmable generation, which introduces the time dimension and cross-band mapping into the “generation–shaping” pipeline via nonlinear frequency conversion or tunable materials. These routes are not oppositional; rather, they are cascaded or hybridized under different task constraints and system trade-offs, jointly serving communications, imaging, and strong-field control.

2.1 Passive wavefront engineering

The passive approach relies on PB geometric phase and resonance phase to deliver reproducible wavefronts in fixed devices. Li et al. theoretically analyzed and

numerically simulated a metasurface–dielectric–metal reflective PB metasurface with orthogonal I-shaped elements, converting circular polarization into multi-order OAM across 0.8–1.4 THz with simulated reflectance >90% and an approximately linear PB phase response (**Figure 1a**) [33]. Extending this idea, Guo et al. theoretically designed and numerically validated a reflective PB metasurface producing $l = 1$ and $l = 2$ vortices over 0.9–1.8 THz, with simulated efficiencies exceeding 90% [25]. In transmission, Niu et al. experimentally realized broadband THz vortex generation (0.5–1 THz) using a three-layer Fabry–Perot cavity comprising orthogonal gratings and split-ring resonators, achieving a peak efficiency of ~65% at 0.5 THz and demonstrating mechanical flexibility for conformal use (see **Figure 1b**) [23]. Fan et al. experimentally demonstrated quasi-perfect vortices at 0.75 THz using all-dielectric high-index silicon pillars, demonstrating multi-channel quasi-perfect vortices and spin-distinguished superposed polarization vortices (**Figure 1c**) [34]. Beyond geometric and resonant phase, Bai et al. theoretically employed bound states in the continuum (BIC) in photonic crystal slabs to access high-Q modes and, in simulations, generate high-purity, higher-order THz vortex radiation (**Figure 1d**) [35].

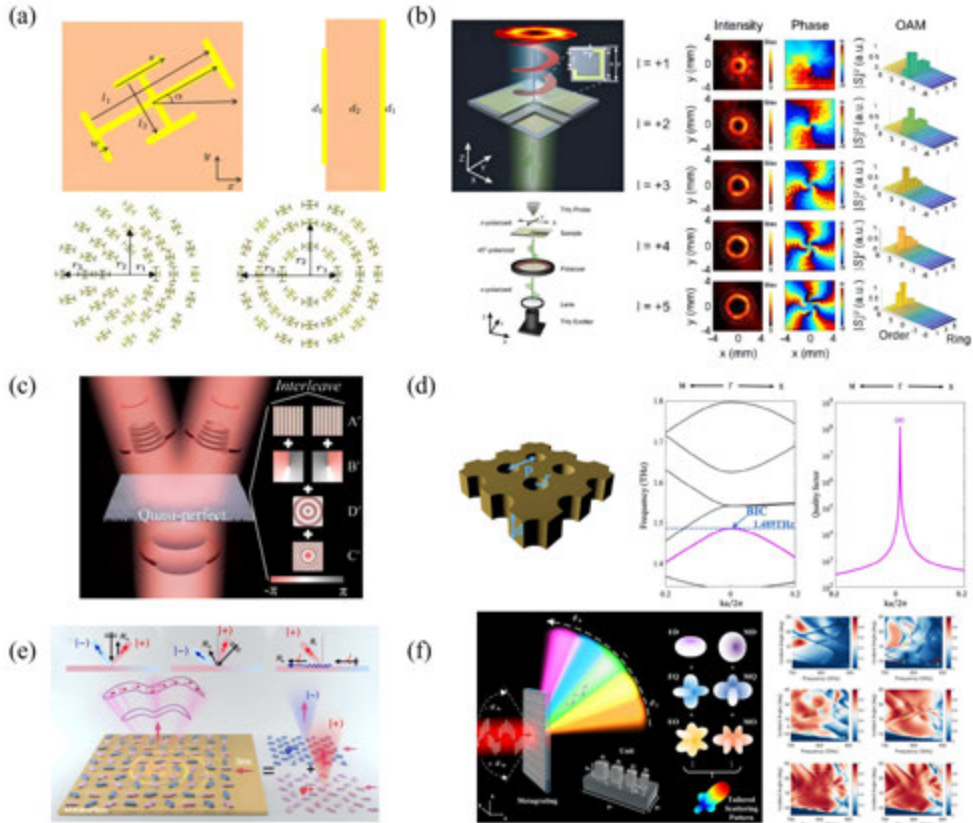


Figure 1.

Passive wavefront engineering of structured THz waves. (a) Metasurface unit and reflective PB design [33]. (b) THz vortex plates [23]. (c) Metasurface-based multi-channel quasi-perfect vortex beams [34]. (d) Photonic crystal with BIC for high-purity vortex radiation [35]. (e) SW-mediated vector-field generation [36]. (f) Asymmetric all-dielectric supergrating [37].

Balancing efficiency and directivity is important in passive devices. Wang et al. reported surface-wave (SW)-mediated, spin-selective on-chip-to-free-space launchers: experiments showed single-spin emission with $\sim 90\text{--}95\%$ relative efficiency (absorption excluded) at ~ 0.4 THz, and a dual-channel implementation driven by a common SW generated a radially polarized Bessel beam with high directivity and near-nondiffracting behavior (**Figure 1e**) [36]. Using an all-dielectric asymmetric supergrating, Peng et al. exploited octupole-assisted multipolar interference to achieve bi-asymmetric scattering and suppress parasitic channels, directing $>80\%$ efficiency into the -1 diffraction order and realizing wide-angle, broadband steering across $\sim 0.70\text{--}0.85$ THz (**Figure 1f**) [37]. In general, passive platforms offer low insertion loss, wafer-scale apertures, and high mode purity, which make them well-suited as fixed front- or back-end modules and as on-chip out-/in-coupling interfaces. Their limitations include a lack of temporal reconfigurability and limited programmability, which often motivates cascading with active stages for adaptive operation and fast multimode switching.

2.2 Active and programmable generation

Active approaches provide flexibility and cross-band capability primarily through optical-domain encoding followed by nonlinear conversion. Spatial phase and/or polarization patterns are first sculpted at near-infrared or visible wavelengths and then transferred to the THz regime via DFG or OR, enabling complex THz wavefronts under broadband pulse excitation. In a DFG scheme, Yuan and Xu' group used two pumps carrying independent phase/polarization profiles and, under type-II phase matching, realized arithmetic combinations of spatial wavefronts, reproducing vortices, Bessel beams, and hybrid states in the THz with high fidelity; the topological charge is continuously tunable while maintaining quasi-nondiffracting propagation (**Figure 2a**) [31]. In an OR scheme, Dhaybi et al. used vector pumps such as radial polarization and mapped their spatial phase/polarization to the THz via $\chi^{(2)}$ rectification in [110]-cut ZnTe, generating quasi-monochromatic THz vortex beams; by rotating a polarizer and filtering in the Fourier plane, they demonstrated switching between $l = +2$ and $l = -2$, indicating effective mapping from vector pumps to THz vortex and polarization states [38]. Moreover, Mou et al. used a liquid-crystal-polymer q-plate to imprint OAM on an 800-nm pump (e.g., $l_p = +1$) and achieved broadband ($\sim 0.1\text{--}3$ THz) $\chi^{(2)}$ rectification in a 111-cut ZnTe; measurements established $l_{\text{THz}} \approx -2l_p$, and rotating the pump polarization rotated the THz vector field, manifested as a coordinated rotation of the four-lobe transverse pattern [39].

Nonlinear metasurfaces integrate phase engineering at the emitter: McDonnell et al. employed PB phase in a C3 nano-antenna nonlinear array to emit THz vortex pulses with spin-orbit decoupling, enabling independent control over frequency, spin, and OAM (**Figure 2b**) [40]. Niu et al. programmed the local nonlinear PB phase via meta-cell orientation and combined it with pump-side DMD spatial gating to realize a 2-bit programmable THz phased array, experimentally demonstrating single/dual-beam synthesis, OAM generation, and scanning slit imaging across $0.8\text{--}1.4$ THz, while avoiding RF phase-shifter insertion loss and maintaining broadband pulse characteristics (**Figure 2c**) [41]. In the high-field

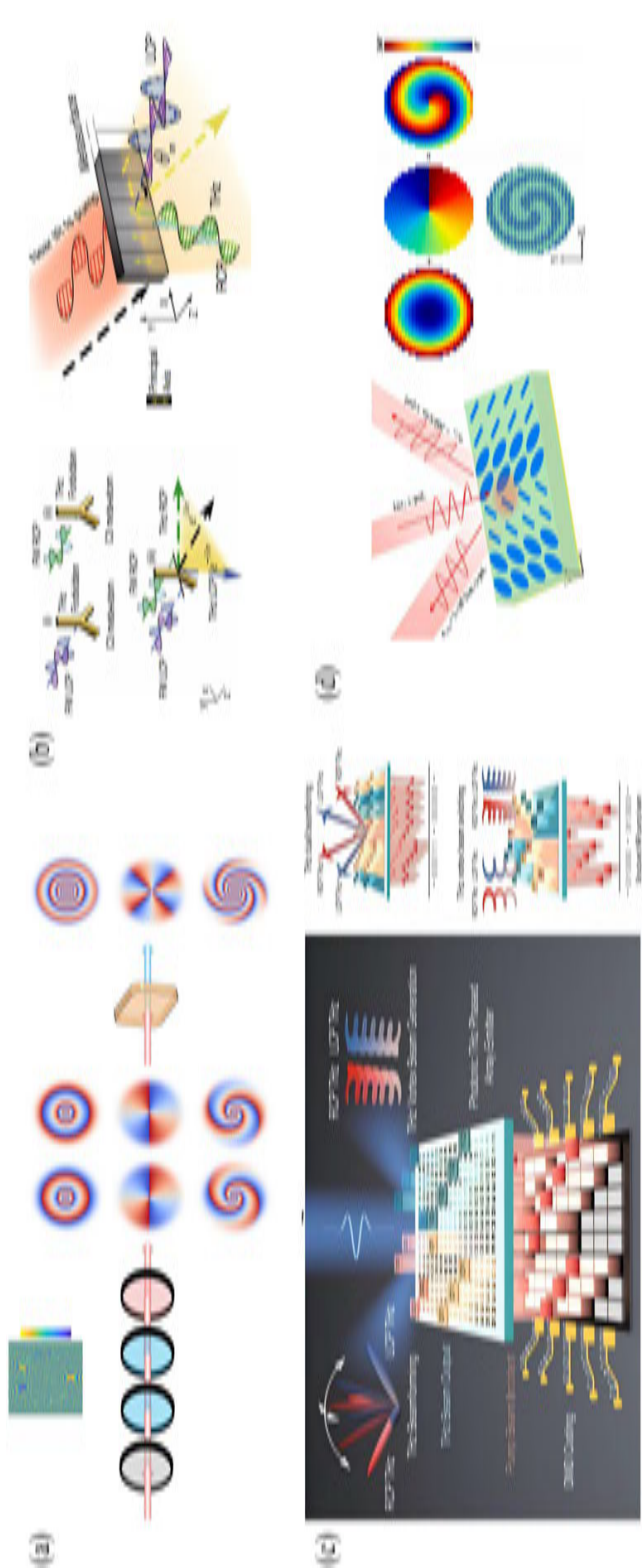


Figure 2. Active, programmable structured THz generation via optical encoding and nonlinear conversion. (a) DFG transfer of optical phase/polarization to THz [31]. (b) Nonlinear PB-phase metasurface emitting THz vortices [40]. (c) Programmable OAM and beam profiles enabled by meta-cell orientation plus DMD-gated pumping [41]. (d) High-field plasma-channel source producing tunable fractional-OAM THz pulses [42].

regime, Wang et al. theoretically and numerically demonstrated plasma-channel-mediated THz emission in which phase matching and plasma dispersion enable high-field vortex pulses with fractional/near-integer OAM; the topological charge was tuned continuously from about -2 to $+2$ by adjusting the laser injection (offset/incident angle), providing a high-field OAM source relevant to strong-nonlinear spectroscopy and field-driven matter studies (**Figure 2d**) [42].

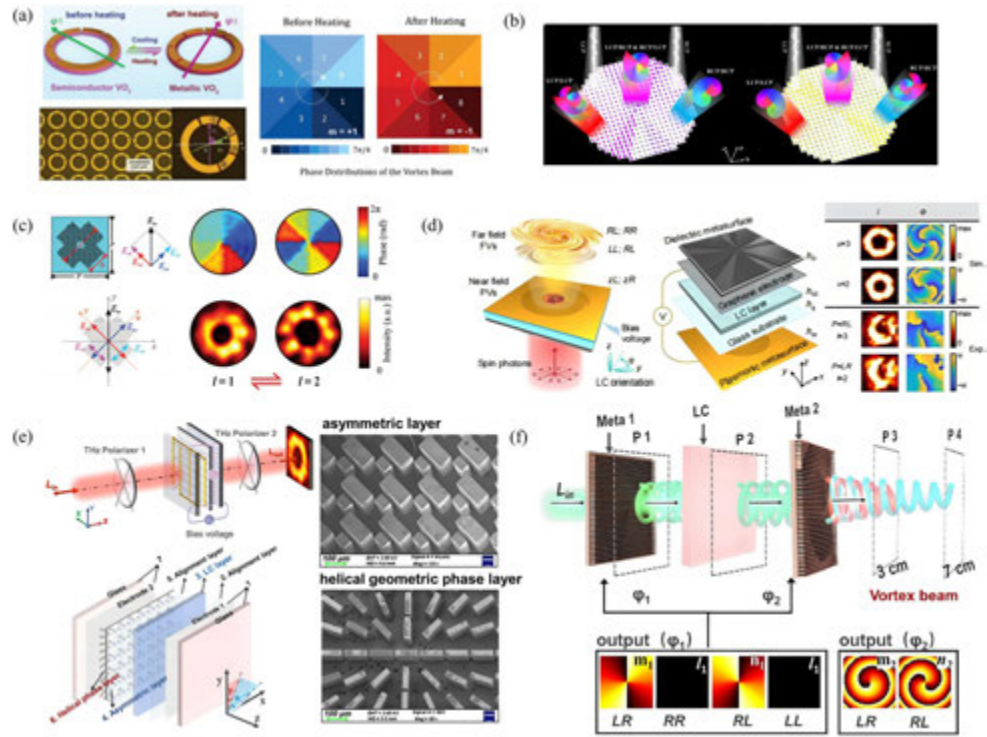
3. Control and reconfiguration strategies

Once structured THz fields can be generated reliably, a central question is how to reconfigure the wavefront on relevant time scales and across use cases. Progress broadly follows two directions: (i) material-centric tuning, which employs phase transitions, carrier modulation, and magneto-optic effects to add temporal control and multimodal multiplexing; and (ii) programmable architectures, which introduce real-time shaping and feedback through reconfigurable metasurfaces and phased arrays.

3.1 Tunable materials for phase, polarization, and mode control

Phase-change and carrier-tunable media offer response times ranging from milliseconds to sub-nanoseconds, with trade-offs among ON/OFF contrast, phase coverage/continuity, and energy cost. Using the VO₂ metal-insulator transition, Dong et al. realized thermally switchable PB-phase THz metasurfaces that toggle among anomalous refraction (± 1 diffraction orders), vortex beams with $l = \pm 1$, and holographic reconstructions near 0.6 THz (**Figure 3a**) [43]. Guo et al. demonstrated a VO₂-based, thermally driven THz metasurface that allocates OAM across six channels with sign inversion, achieving $l = -3/-2/-1$ (metallic state) and $l = +1/+2/+3$ (insulating state) over 1.35–1.39 THz (**Figure 3b**) [44]. With electrostatic Fermi-level tuning in graphene, Ma et al. [45] and Chen et al. [27] numerically demonstrated rapid switching of THz topological charge at 2–2.4 THz and 4.5–5.5 THz, respectively, while preserving linear polarization (**Figure 3c**). Wang et al. experimentally demonstrated a cascaded dielectric-LC-plasmonic metadvice that generates near-field (plasmonic) and far-field (free-space) multi-order vortices over 0.44–0.52 THz, and electrically selects among 12 vortex modes via LC reorientation, showcasing near/far-field integration with low-voltage control (**Figure 3d**) [46]. Zhao et al. experimentally introduced tunable chirality by coupling an LC-integrated asymmetric metasurface to a helical phase layer to control spin-orbit interaction, achieving active interconversion between $l = \pm 2$ vortices and cylindrical vector beams with excitation efficiencies up to $\sim 92\%$ and modulation depths up to $\sim 72\%$ (**Figure 3e**) [47]. Extending this stack, Zhao et al. built an electrically controlled spiral-phase/LC/focusing-lens assembly that independently addresses RR/RL/LL/LR spin channels (right-right, right-left, left-left, and left-right circular polarization, denoting input-output) over 0.4–0.6 THz, with 94% peak efficiency, $> 80\%$ purity for $l = 2, -1, 0, 1$, and dynamic dual-focal switching (**Figure 3f**) [48].

For thermal platforms, Yang et al. numerically leveraged the temperature-dependent plasma frequency and loss of InSb to implement broadband vortex phase engineering, thermally tuning operation from 1.8 to 4.5 THz while maintaining high


Figure 3.

Control and reconfiguration of structured THz waves with tunable materials and cascaded metasurfaces. (a) VO₂-integrated metasurface enabling thermal switching of $l = \pm 1$ vortices [43]. (b) Thermally reconfigurable multi-channel OAM metasurface [44]. (c) Graphene metasurface [27]. (d) LC-based cascaded device [46]. (e) LC-cascaded metadvice [47]. (f) Schematic of the LC-sandwiched dual-metasurface stack [48].

vortex mode purity [49]. He et al. numerically exploited InSb's thermoplasmonic response in a reflective PB metasurface to realize broadband THz vortex generation ($l = \pm 1, \pm 2, \pm 3$ from ~ 0.65 – 1.66 THz), with ON/OFF control via global or local heating, suggesting practical routes toward device implementation and scaling to larger apertures [50]. In addition, Nobahar et al. theoretically analyzed magnetically reconfigurable plasma–ferrite stacks with magneto-optic anisotropy, showing joint control of polarization, phase, and THz vortex beam profiles via external magnetic bias, thereby introducing magnetic bias as an additional degree of freedom for OAM wavefront engineering [51]. In practice, VO₂ is attractive for high-contrast switching and “full-space” phase shaping, LCs tend to support near-continuous phase and high vector-beam purity, graphene targets high-speed electrical control, and InSb enables fabrication-friendly thermal tuning. Coordinated use of these materials may approach a balanced compromise among speed, phase depth/continuity, and loss.

3.2 Reconfigurable and programmable architectures

Beyond material tuning, two complementary paradigms enable real-time wavefront control at the system level: digital coding and photonic phased arrays. First, bit-discrete programmable metasurfaces refresh codewords via low-loss

electrical or mechanical actuation. Zhou's group cascaded rotatable transmissive metasurfaces and, via relative rotation, tuned the effective Jones matrix in real time, enabling joint control of the deflection angle (polar/azimuthal steering) and polarization within a single THz device – linking structural reconfiguration to matrix-level programmability (**Figure 4a**) [52]. Second, optically driven photonic phased arrays excite nonlinear PB meta-arrays spatiotemporally using external spatial light modulators, avoiding the insertion loss and bandwidth limitations of RF phase-shifter networks. Using a DMD-programmed nonlinear PB metasurface, they demonstrated 2-bit beam scanning, target (slit) imaging, and OAM generation over 0.8–1.4 THz; the DMD affords kHz-class refresh in principle (the demo was system-limited), with pulse bandwidth and modal flexibility preserved (**Figure 4b**) [41]. Relative to material “gating,” these approaches emphasize array-codebook co-design: phase-bit depth, array fill factor, and aperture-normalized sidelobe level jointly determine modal crosstalk and effective isotropic radiated power (EIRP). In practice, 2–3 bits can support robust low-order OAM and multi-beam synthesis; higher bit depths help suppress sidelobes and inter-channel leakage for larger apertures and wider bandwidths [22].

In parallel, passive mechanical and geometric reconfiguration provide ultra-low-loss control on longer time scales. Zhang et al. theoretically proposed and numerically analyzed an all-dielectric metasurface embedding non-Hermitian exceptional points (EPs) to realize exceptional-topological-phase-based decoupling of phase and spin, enabling topologically robust wavefront control under perturbations (**Figure 4c**) [53]. Subsequent theory and experiments showed that, near EPs, spin and phase can be independently tuned, with reduced sensitivity to device deviations and external disturbances [54]. At the programmable-source level, Wang et al. demonstrated programmable exchange-biased spintronic emitters that flexibly generate structured THz waves via laser-assisted magnetization patterning while reducing system loss (**Figure 4d**) [55]. Cheng et al. demonstrated room-temperature, field-free, polarization-controllable ultrafast THz emission from an antiferromagnet/topological-insulator $\text{MnSe}/(\text{Bi,Sb})_2\text{Te}_3$ heterostructure. Circularly polarized THz pulses (1.0–3.1 THz) arise from the coexistence of magnetic-dipole radiation and spin-charge-conversion-driven electric-dipole radiation; the emitted polarization is tuned by the pump polarization [56]. The results support unified codebook design and coordinated control across the full stack, from on-chip implementations to free-space propagation.

Overall, materials and architectures are complementary across time, space, frequency, and spin-orbit degrees of freedom. Material tuning provides physical time variance and multimodal multiplexing – suited to moderate refresh with substantial phase depth and stable shaping. Programmable architectures deliver system-level real-time control and closed-loop optimization – suited to rapid codebook updates, beam/mode routing, and in-situ calibration. A hierarchical strategy commonly uses high-efficiency geometric/resonant phase as the primary energy path; overlays LCs, graphene, VO_2 , and ferroelectric LC media for near-continuous phase with coordinated polarization/amplitude control; and finally employs electronically or optically driven arrays for scene adaptation and residual correction. Employing closed-loop operation with in-situ sensing allows kHz-rate updates that sustain mode purity and link quality, effectively mapping discrete phase control to adaptive, task-oriented reconfiguration [22, 41].

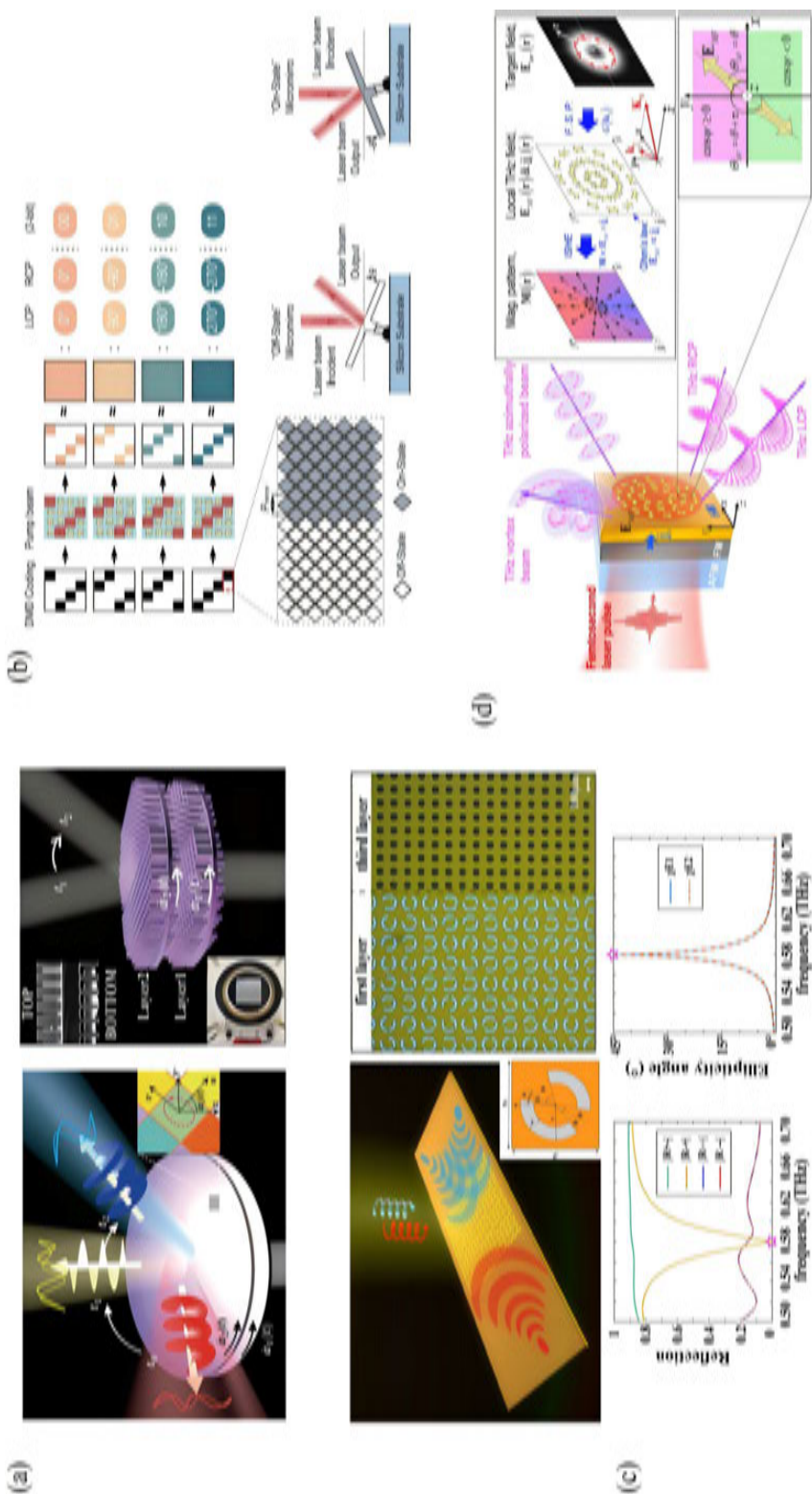


Figure 4. Reconfigurable and programmable architectures for structured THz control. (a) Cascaded, rotatable metasurfaces [52]. (b) Optically driven photonic THz phased array [41]. (c) Topological metasurface for THz surface-plasmon control [53]. (d) Programmed exchange-biased spintronic THz emitter [55].

4. Multifunctional integration

As generation and control mature, emphasis shifts from feasibility to integrating multiple tasks and channels within a single device or chip. Two directions are particularly relevant: (i) coordinated multiplexing across frequency, polarization, and spatial modes; and (ii) co-design of vector fields with topological properties. The goal is to increase functional density and reconfigurability under constraints of volume and loss.

4.1 Multichannel and polarization multiplexing

Combining geometric and resonant phase, and exploiting polarization and frequency as approximately orthogonal degrees of freedom, has enabled THz devices with increasing multichannel capability. Zhang et al. experimentally demonstrated a chip-integrated, all-silicon polarization-multiplexed metasurface at 1.3 THz that simultaneously generates four-channel orthogonal polarized vortex beams on a single aperture. Under x - and y -polarized incidence, channels 1 and 2 output linear-polarized fundamental vortex states, while channels 3 and 4 produce left- and right-handed circular polarization superposition vortex states. The far-field shows petal-like patterns whose lobe number equals the sum of the topological charges (**Figure 5a**), clearly indicating polarization–OAM multiplexing on a single aperture [57]. Cheng et al. theoretically designed a dual-band reflective metasurface that independently controls linear polarization (0.7 THz, via propagation phase) and circular polarization (1.4 THz, via geometric phase), enabling band-/spin-partitioned beam deflection, planar focusing, and vortex beam generation [58]. Liu et al. numerically demonstrated a transmissive tri-band geometric-phase metasurface that generates OAM modes $l = \pm 1, \pm 2, \pm 3$ at 0.6, 0.857, and 1.892 THz, with mode purity typically exceeding $\sim 60\%$, indicating cross-band modal consistency and multiplexing potential (**Figure 5b**) [59].

On material co-design, Liu et al. theoretically proposed and numerically validated a VO₂-integrated, full-space THz metasurface that thermally switches between transmission and reflection operation (**Figure 5c**) [60]. Li et al. theoretically designed a VO₂-based THz metasurface with anisotropic Z-shaped metal units that switch between transmission and reflection and use PB phase for broadband phase/polarization control [61]. Extending these ideas, Zhang et al. theoretically proposed omnidirectional devices that generate vortices and convert polarization at multiple frequencies in parallel, outlining a multidimensional multiplexing scheme spanning polarization, frequency, OAM, and space (**Figure 5d**) [62].

Overall, these examples reveal a practical breakdown into polarization selection, frequency partition, OAM multiplexing, and spatial routing. In such schemes, polarization separates channels or functions; frequency partitions tasks and links; OAM supplies parallel modal carriers; and spatial control sets beam direction and routing. This factorization can reduce inter-channel crosstalk and simplify separation/filtering along the link, providing a basis for mode-division multiplexing, frequency-division access, and hybrid on-chip/free-space routing. For fabrication and integration, all-dielectric (e.g., silicon) platforms tend to offer high efficiency and bandwidth, whereas phase-change VO₂ provides switchability; their complementarity offers a route to balance efficiency, bandwidth, and reconfigurability within one system.

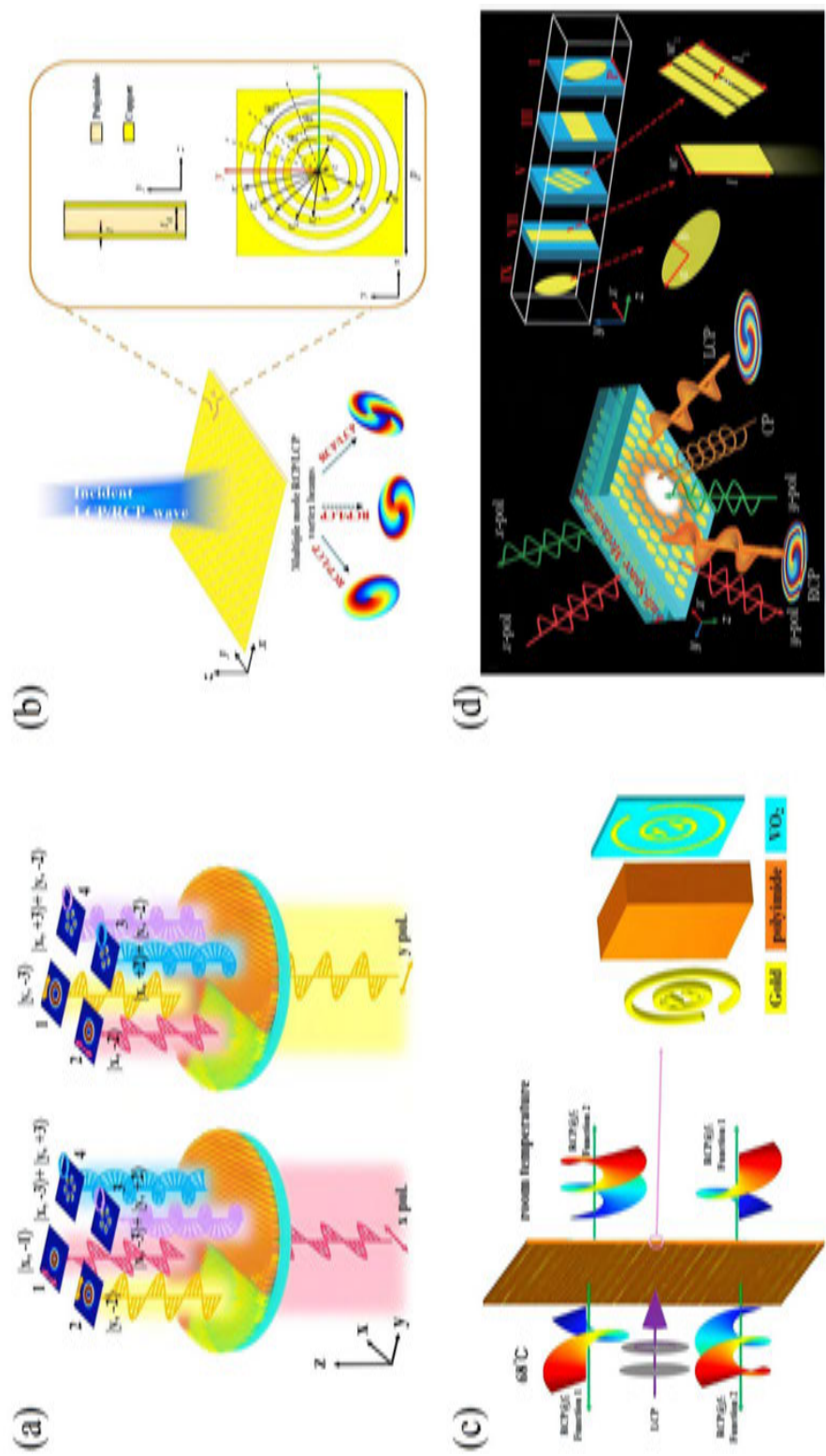


Figure 5. (a) Four-channel polarization-multiplexed THz generation. (b) Multispectral OAM generator [59]. (c) Dual-frequency bidirectional OAM emitter [60]. (d) Omnidirectional bifunctional THz metasurface [62].

4.2 Hybrid vector–topological design

A second route to multifunctionality is the co-design of vector fields and topological properties, enabling controlled coupling among polarization, OAM, and field profiles during propagation. On low-loss all-dielectric platforms, Yao and co-workers designed anisotropic silicon pillars to independently phase the two circular components and generate THz vector vortex beams with longitudinally varying polarization. The same group demonstrated spin-dependent longitudinal evolution of OAM with an all-dielectric THz metasurface: topological charge transitions from $l = +2$ to -2 under left-circular polarization (LCP) and from $l = +1$ to -1 under right-circular polarization (RCP), along with polarization-state transitions from radial to second-order radial and from azimuthal to radial (**Figure 6a**) [63]. Using an all-silicon metasurface and coherent superposition of orthogonal circularly polarized THz waves via a long focal-depth metalens, the same group realized

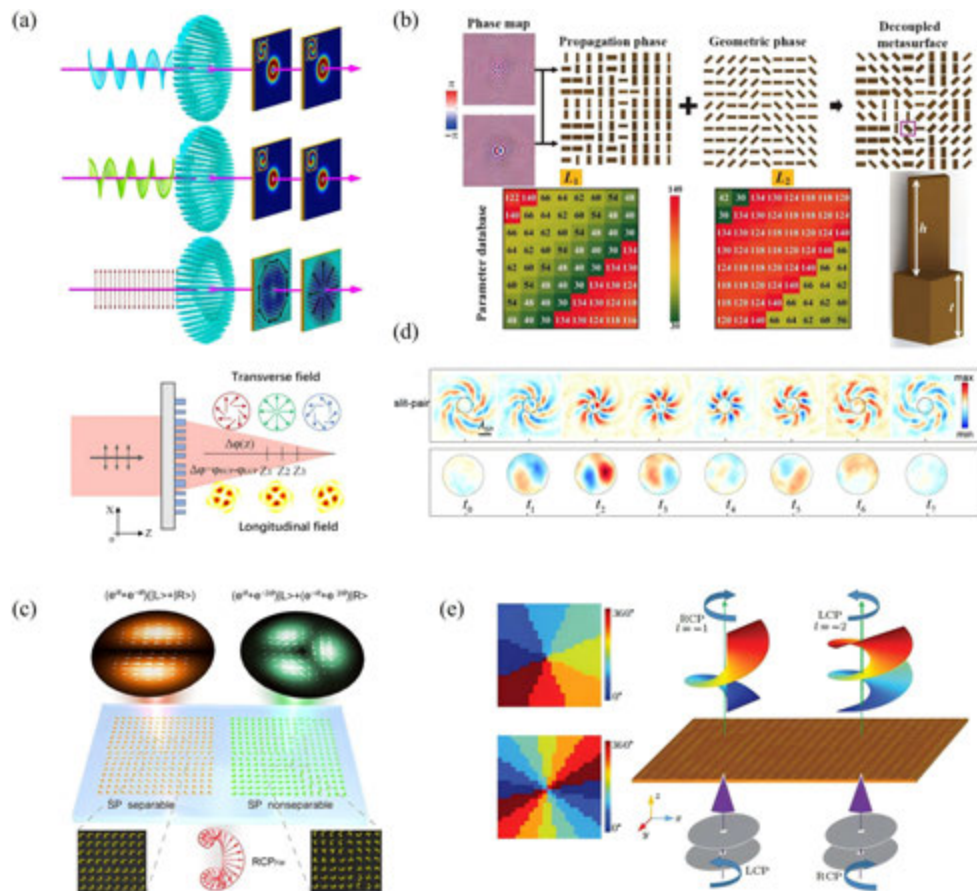


Figure 6. Hybrid vector–topological control of structured THz fields. (a) All-dielectric metasurfaces generating longitudinally varying vector vortex beams [63]. (b) Single-layer decoupled metasurface integrating propagation and geometric phase [65]. (c) Separation and non-separation schemes enabling near-independent polarization/amplitude control of $l = \pm 1$ [66]. (d) Time-resolved snapshots of surface-plasmon fields [67]. (e) The phase distribution of vortex beams with $l = -1$ and -2 under the incidence of LCP and RCP [68].

structured vector fields along the propagation direction, including continuous variation of the radial (transverse) component and spin–orbit coupling–induced manipulation of the longitudinal component [64]. For channel decoupling, Li et al. used a single-layer silicon metasurface to address LCP/RCP independently near 0.7 THz, encoding off-axis vector vortices with $l = \pm 1$ and ± 6 and enabling rotatable polarization control, indicating joint programmability across spin, orbit, and space (**Figure 6b**) [65]. On a nonlinear platform, Wang et al. developed a chiral nonlinear metasurface offering independent, near-continuous control of polarization and amplitude for vector vortices with $l = \pm 1$, suggesting device-level decoupling between amplitude–polarization parameters and OAM (**Figure 6c**) [66].

Temporal structuring extends control beyond the static phase. With a plasmonic metasurface, Yuan et al. tailored pulse-scale dispersion and phase to reveal transient vortex states and gate their appearance in time, enabling time-resolved handling of topological features (**Figure 6d**) [67]. At the device stack level, Zhong et al. theoretically proposed multilayer switching of focus, OAM order, and polarization on a common platform, suitable for imaging, focusing, and mode routing, with increased complexity balanced against insertion loss (**Figure 6e**) [68]. For cross-band transfer, two NIR phase profiles were programmed and then mapped via type-II DFG to THz, yielding tunable vortices, Bessel beams, and hybrid vortex–Bessel beams with broadband nondiffracting behavior [31]. At the source interface, Pettine et al. employed asymmetric Au nanoantennas on graphene to launch directional nanocurrents under ultrafast drive, coupling electronic degrees of freedom to optical spin and OAM for on-demand radial/azimuthal vector-vortex synthesis; achievable speed and efficiency are set by carrier dynamics and coupling losses [69].

Hybrid design allocates approximately orthogonal resources under finite aperture and loss constraints: geometric/resonant phase with engineered anisotropy for robust generation and interconversion of vector states and OAM; nonlinear and cross-band mapping for complex-field synthesis and nondiffracting transport; and temporal gating/transient dispersion for pulse-scale control of topological features. Multilayer, switchable stacks offer a route to combine static multiplexing with time-varying switching on a common platform. For system translation, open issues include establishing comparable metrology for inter-channel crosstalk and mode purity (harmonized across measurement setups) and tolerance control for broadband, large-aperture fabrication (phase error, sidewall angle, and thickness drift, which can accumulate for higher topological charge l).

5. Applications

5.1 Communications and information processing

For short- to mid-range high-speed interconnects and indoor backhaul, structured THz links are evolving from mode generation toward link-level design. For communications and information processing, such cascaded dielectric–LC–plasmonic modules, with electrically controlled spin–orbit coupling, enable 12-state THz routing and reconfigurable OAM multiplexing under low-voltage drive (**Figure 7a**) [46]. On an all-silicon dielectric metasurface, arranging polarization-maintaining and polarization-conversion meta-atoms enables simultaneous polarization–OAM control with four spatially decoupled linear/circular vortex channels at various charges,

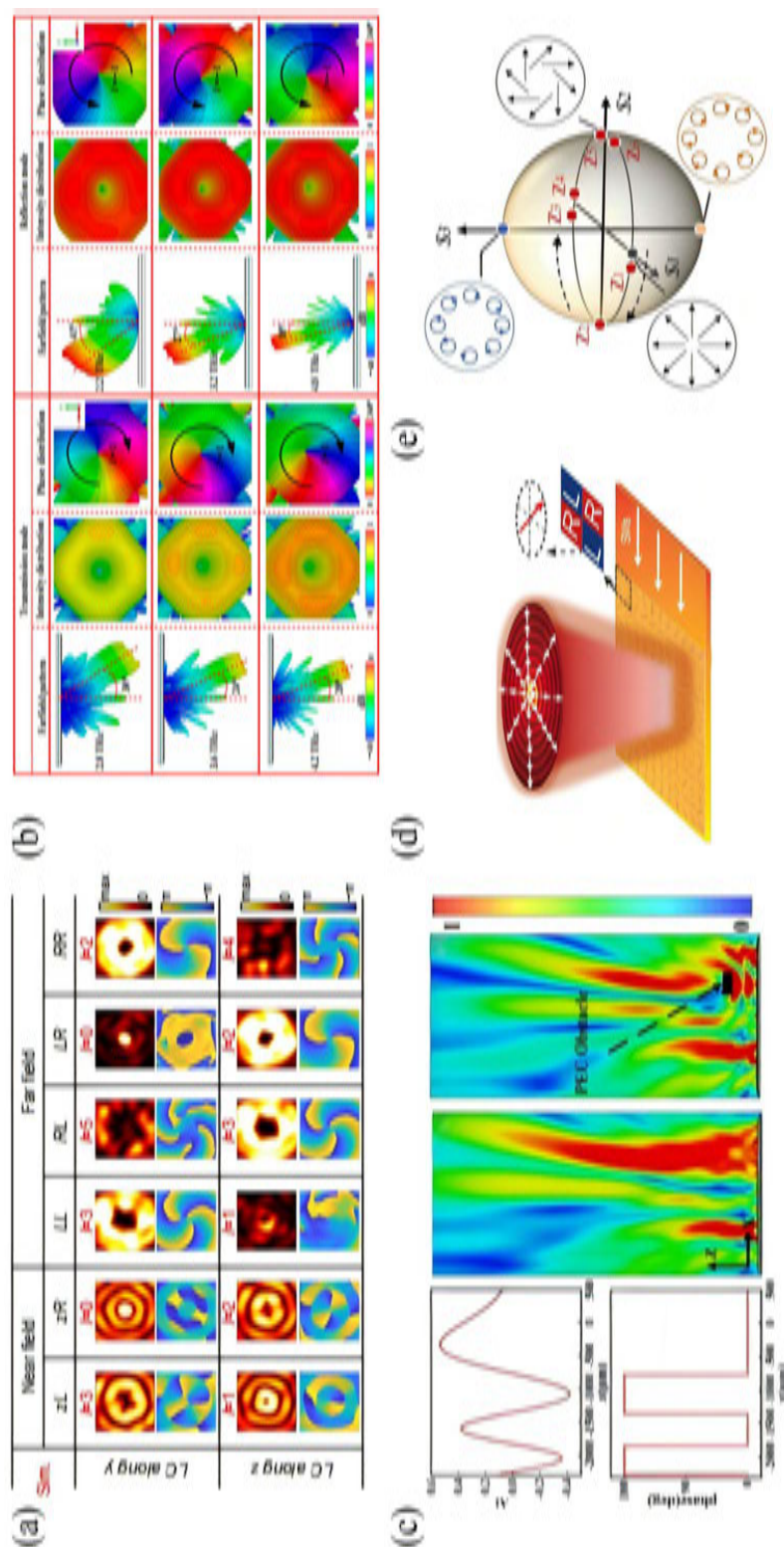


Figure 7. Communications and sensing using structured THz waves. (a) Cascaded metasurface for 12 OAM modes [46]. (b) Deflected vortex beams [61]. (c) Airy beam generation [16]. (d) SW-to-free-space coupling [36]. (e) Longitudinal evolution of E_x/E_z components in a transmitted vector field [64].

providing added degrees of freedom for parallel multimode THz transmission [57]. Optically driven nonlinear PB phase also enables a programmable THz phased array with 2-bit beam control, real-time imaging, and OAM encoding [41]. Moreover, by switching VO₂ between insulating and metallic states, the same device achieves full-space wavefront shaping – deflected, focused, and nondiffracting vortex beams – in transmission/reflection, enabling dynamically reconfigurable THz links (**Figure 7b**) [61].

Practical design is constrained by EIRP, effective aperture, beam divergence, and power budget [9, 70]; intermodal crosstalk and weather-induced fading can be mitigated by precoding/equalization, multiple input multiple output (MIMO), and data-driven calibration. Receiver design involves trading bit error rate (BER) against coding overhead to balance capacity and reliability [71]. Former studies also indicate that hybrid on-chip/free-space routing is a viable path for short- to mid-range backhaul and indoor high-speed interconnects: on-chip modules handle mode generation/multiplexing and out-coupling, while free-space links provide beam assignment and environmental adaptation. Unified codebooks and in-operation calibration support stable end-to-end links (**Figure 4b**) [41, 46].

5.2 Imaging and sensing

For imaging and sensing, two practical directions have emerged. Nondiffracting and self-accelerating beams (e.g., Bessel and Airy beams) provide an extended depth of field and some tolerance to occlusion and scattering, improving stability in heterogeneous or curved scenes at the cost of power in sidelobes. Wide-aperture vector fields and vector vortices enable selective coupling and multi-parameter interrogation of chirality and anisotropy, provided the generation and detection chains preserve the required polarization and phase structure.

Li et al. demonstrated a reflective Au/polyimide/Au metasurface that generates Airy, dual-Airy, and multi-order/focused vortex beams over 0.8–1.3 THz, exhibiting self-bending and extended depth of field, which can stabilize imaging under scattering; the efficiency is determined by metallic loss and bandwidth (**Figure 7c**) [16]. Nan et al. employed trajectory-phase encoding in a three-layer transmissive metasurface to launch Bessel-like beams on cylindrical helices with rotating polarization, yielding long-DOF, anti-scattering propagation with the expected sidelobe penalty [18]. SW-excited, radially polarized Bessel emission offers high directivity and efficiency, providing a compact, low-loss front end for imaging (**Figure 7d**) [36].

Vector control also supports material-selective probing. Independent LCP/RCP phasing drives longitudinal evolution of E_x/E_y in the transmitted field, enabling z -dependent transverse polarization (with coupled axial control) for anisotropic/chiral excitation; efficiency is limited by dielectric loss and the designed bandwidth (**Figure 7e**) [64]. Zhou et al. used a dual-metal metasurface at 0.14 THz to generate radially polarized vector vortices with selectable topological charge, furnishing polarization–phase templates for anisotropy sensing [72]. These exchange-biased programmable emitters cover the full Poincaré sphere and generate circularly polarized vortices, enabling reconfigurable polarization–OAM states for spatial imaging (**Figure 8c**) [55].

Overall, these studies indicate a system strategy in which, under dose and safety constraints, hollow-beam energy distributions and vector-polarization selectivity

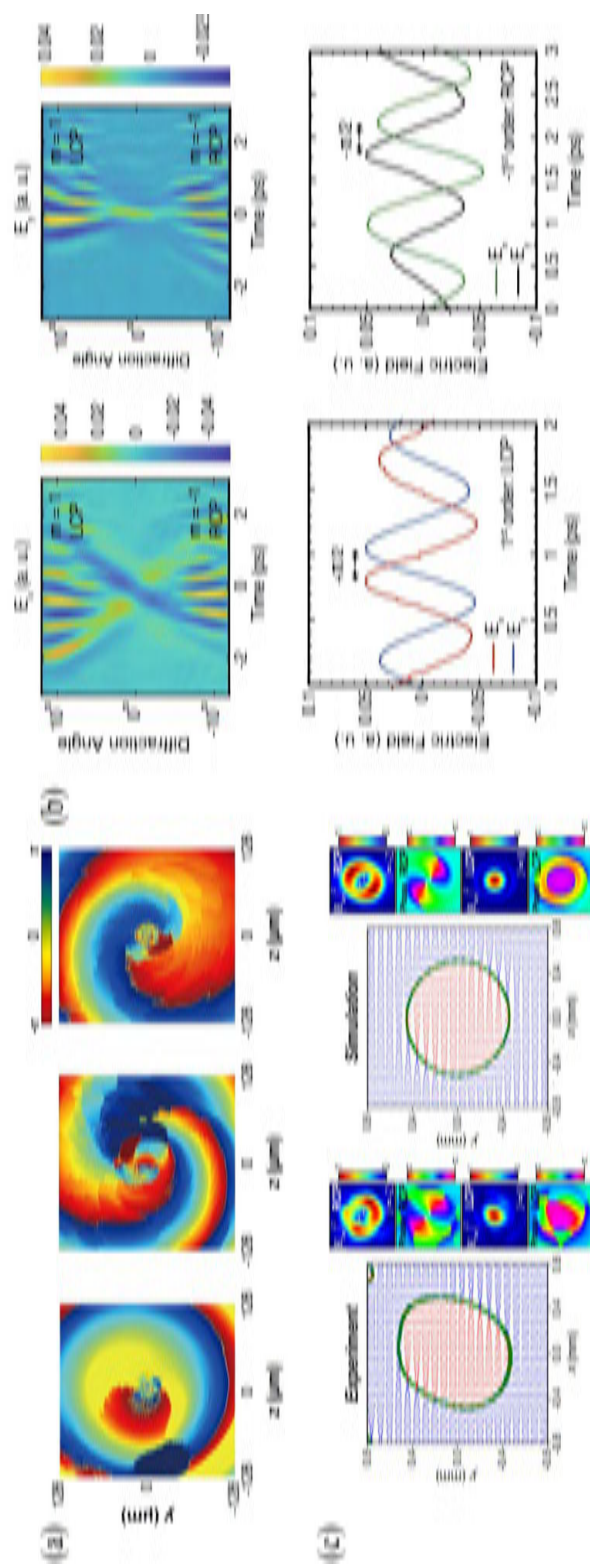


Figure 8. Strong-field and nonlinear/spintronic THz emitters. (a) Plasma-channel source generating near-GV/m fractional-OAM pulses [42]. (b) Nonlinear PB-metasurface emission [40]. (c) Programmable spintronic THz emitter covering the Poincaré sphere [55].

can enhance tomographic resolution and contrast without increasing total exposure. Coupled with on-chip/free-space bridging and programmable scanning, robust reconstruction can be maintained in complex media and in low-contrast or weak-scattering scenarios, subject to limits set by aperture, efficiency, calibration, and detector bandwidth.

5.3 Strong-field and nonlinear/spintronic emitters

Strong-field platforms are enabling structured THz at application-relevant intensities. In particular, oblique injection into a parabolic plasma channel produces near-GV/m THz vortices with continuously tunable (fractional/near-integer) OAM, offering a high-field, OAM-tunable source for strong-nonlinear spectroscopy and coherent control (**Figure 8a**) [42]. The PB-assisted $\chi^{(2)}$ processes yield broadband, single-cycle THz with spin-orbit decoupling and spatial/mode multiplexing, indicating an on-chip route where emission and wavefront shaping are co-designed (**Figure 8b**) [40]. In an AFM/TI MnSe/(Bi,Sb)₂Te₃ heterostructure, optically controllable circularly polarized ultrafast THz emission at room temperature and zero field is achieved via the coexistence of magnetic dipole radiation and spin-charge conversion, with polarization controlled by the excitation laser polarization [56]. The current trend is to move beyond field strength and bandwidth, with increasing emphasis on mode purity, EIRP, and stability. Co-design of emission and shaping in integrated nonlinear devices can reduce downstream loss and improve end-to-end efficiency. Spintronic and plasma sources, however, would benefit from systematic quantification of modal purity and pointing drift, as well as assessments of how these factors impact link performance and imaging stability.

5.4 On-chip coupling, manufacturing, and systemization

Translating structured THz from laboratory demonstrations to deployable platforms is often limited by outcoupling, manufacturability, and interface standardization. Momentum-space BIC provides a compact, alignment-tolerant route to vortex emission, offering robustness and mode purity (**Figure 9a**) [35]. Besides, a SW-excited metasurface forms a low-loss bridge between on-chip and free space: under single SW excitation, it selectively radiates a designated spin channel, and combining two spin channels yields a radially polarized Bessel beam with high directivity (**Figure 9b**) [36].

On fabrication and assembly, projection micro-stereolithography (PμSL), developed by Ke and colleagues, provides scalable, low-cost routes to multi-material, multi-scale meta-optics and lenses. At ~ 0.1 THz, tightly focused shaping of integer and fractional vortices has been demonstrated (**Figure 9c**) [73]. At the system level, optically driven programmable phased arrays have exhibited kHz-rate beam synthesis, scanning, and target recognition, indicating a path toward real-time operation [41]. Combining high-efficiency BIC outcoupling with PμSL/3D-printing scale-up may accelerate progress from controllable field patterns to integrated systems, while motivating harmonized metrology and interface standards for chip-to-free-space interconnects (**Figure 9d**) [74, 75].

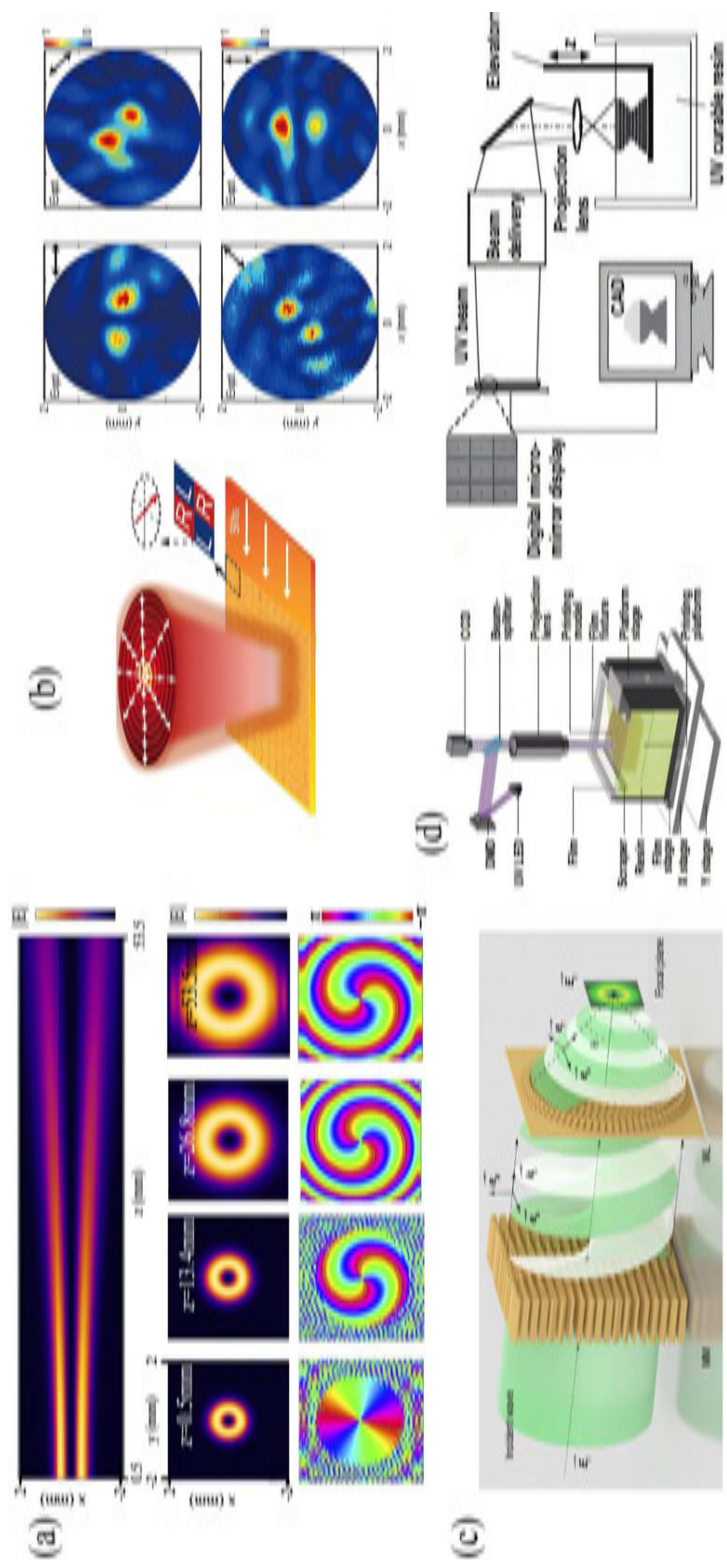


Figure 9. On-chip coupling, manufacturability, and systemization of structured THz. (a) Momentum-space BIC photonic-crystal emitter [35]. (b) SW-excited metasurface. (c) Schematic diagram of beam conversion by cascade modules [73]. (d) Projection micro-stereolithography (PμSL) [74, 75].

5.5 LC cascading and dynamic evolution

In recent years, significant progress has been made in THz-band photonic devices, particularly in OAM manipulation based on LC-integrated cascaded metasurfaces and research on non-Hermitian systems. In the generation and control of vortex beams, studies have focused on improving mode purity, expanding the degrees of freedom in manipulation, and achieving dynamic reconfiguration. For instance, Wang et al. and Zhao et al. realized multi-modal interference and dynamic encoding of plasmonic vortices and free-space vortices through LC integration and graphene electrodes, respectively, significantly enhancing information capacity and reconfigurability (**Figures 10a and b**) [76, 77]. In the same year, Chang's group constructed a multi-channel vortex generator using cascaded metasurfaces and an LC layer, successfully breaking the spin-orbit coupling limitation and achieving high-purity, low-crosstalk multi-mode multiplexing and dynamic switching in both near-field and far-field [46, 48]. In the field of non-Hermitian photonics, Wang et al. and Fu et al. explored the dynamic evolution paths from BICs to higher-order EPs and, for the first time, experimentally observed a third-order EP (EP3) in metasurfaces, providing new mechanisms for highly sensitive sensing and topological photonic control (**Figures 10c and d**) [78, 79].

Furthermore, Zhang et al. combined metasurfaces with deep learning to construct a high-dimensional photodetector capable of simultaneously perceiving intensity, polarization, and frequency, expanding the application prospects of photonic devices in intelligent perception and information security [80]. Collectively, these studies have propelled THz photonic devices toward multifunctionality, high dimensionality, programmability, and intelligence, laying a theoretical and experimental foundation for the next generation of integrated photonic chips, high-capacity communication systems, and advanced photoelectric sensing platforms.

6. Conclusion and outlook

THz structured wave now spans passive, active, nonlinear, and programmable layers within system-level architectures (as shown in **Table 1**). NIR pre-encoding mapped to THz by difference-frequency generation or optical rectification provides a practical “pump-in/THz-out” control path, with fidelity limited by phase matching, group-velocity mismatch, and walk-off [31, 41]. SW-to-free-space couplers and all-dielectric multipolar emitters improve extraction efficiency and enable broadband, wide-angle steering within aperture and radiation-loss constraints [36, 37]. Material co-tuning across VO₂, LCs, graphene, and InSb allows trade-offs among channel count, phase coverage/continuity, and speed, with insertion loss and drive power balanced against tuning range [27, 44, 48, 81]. Topological and non-Hermitian controls can add robustness of phase and chirality to fabrication and environmental perturbations [53, 82]. Representative systems report diffraction efficiencies exceeding ~80%, cross-polarization conversion near or above 0.9, mode/polarization purity on the order of 70%–80%, modulation depths above ~70%, and kHz-rate refresh, supporting OAM-multiplexed links, reconfigurable imaging, and strong-field drivers.

Looking toward application-driven deployment, emphasis shifts from maximizing single metrics to translating multi-degree-of-freedom advantages into measurable,

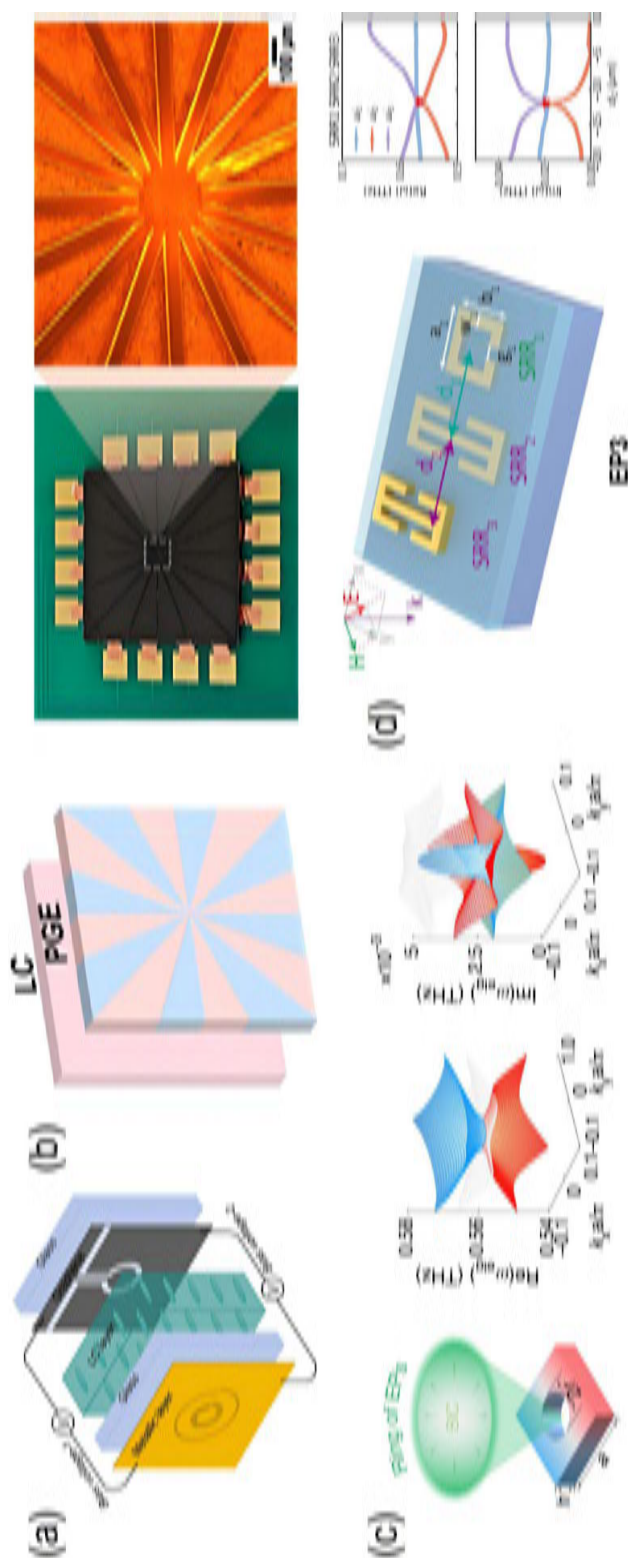


Figure 10. LC-cascaded THz metasurface and dynamic control of EP. (a) Structure of the THz LC coding PV interferometer [76]. (b) Structure of 16-pixel angular-patterned PGE [77]. (c) Transition from one bound singularity to a two-dimensional exceptional ring. A schematic of the transition of one BIC point to an exceptional unit cell [78]. (d) Schematic diagram of EP3 system [79].

Device name	Material	Control method	Structured wave type (with key performance)	Ref.
Orthogonal I-shaped Reflective PB Metasurface	Metal/ Dielectric	Static (Passive)	Multi-order OAM vortices; simulated reflectance > 90%, 0.8–1.4 THz.	[33]
Multilayer Fabry–Perot Vortex Plate	Metal & Dielectric (3-layer cavity)	Static (Passive)	Broadband THz vortex beams; peak efficiency ~ 65% at 0.5 THz, 0.5–1 THz.	[23]
All-Dielectric Quasi-Perfect Vortex Generator	Silicon pillar array	Static (Passive)	Multi-channel quasi-perfect vortices and spin-distinguished polarization vortices.	[34]
BIC Photonic Crystal Vortex Emitter	Photonic crystal slab (supporting BIC)	Static (Passive)	High-purity, higher-order THz vortex radiation (high-Q modes).	[35]
SW-Excited Vector Bessel Launcher	Metasurface (SW excitation)	Static (Passive)	Radially polarized Bessel beam; high directivity, ~ 90–95% relative efficiency at ~ 0.4 THz.	[36]
Nonlinear PB Vortex Emitting Array	C3 nano-antenna non-linear array	Optical pumping (Nonlinear)	THz vortex pulses with spin-orbit decoupling: Independent control of frequency, spin, and OAM.	[40]
VO ₂ -Thermo-Switchable OAM Metasurface	VO ₂ -integrated metasurface	Thermal (Metal-insulator transition)	Multi-channel OAM beams ($l = -3/-2/-1$ and $+1/+2/+3$); switching between states at 1.35–1.39 THz.	[44]
Graphene-Electrotunable Vortex Generator	Graphene metasurface	Electrostatic (Fermi-level tuning)	THz vortex beam with switched topological charge: 2–2.4 THz and 4.5–5.5 THz.	[27, 45]
Liquid-Crystal-Plasmonic Cascaded Metadevice	Dielectric-LC-Plasmonic stack	Electrical (LC reorientation)	12 selectable vortex modes (near- and far-field); low-voltage control, 0.44–0.52 THz.	[46]
LC-Based Vortex-Vector Beam Converter	LC-integrated asymmetric metasurface	Electrical (LC reorientation)	Active conversion between $l = \pm 2$ vortices and cylindrical vector beams; ~ 92% efficiency, ~ 72% modulation depth.	[47]
Nonlinear DFG Tunable Vortex-Bessel Source	Nonlinear crystal (e.g., ZnTe)	Optical encoding (NIR pump, DFG/OR)	Tunable vortices, Bessel beams, and hybrid vortex-Bessel beams; broadband, quasi-nondiffracting.	[31]
Plasma-Channel High-Field OAM Source	Plasma channel	Laser injection (Offset/angle)	High-field (near-GV/m) vortex pulses with tunable fractional/integer OAM (from ~ -2 to + 2).	[42]

Table 1.
Different approaches for the of thz structured waves.

transferable, and maintainable system gains. Three directions appear promising. First, combine broadband nonlinear emission with low-bit discrete phase tuning using reconfigurable elements. Use nonlinear emitters as low-insertion-loss transmitters/front ends; overlay amplitude or phase tuning via LCs, graphene, or VO₂; and apply in situ, data-assisted calibration to compensate for fabrication tolerances and thermal/environmental drift. This route aims to improve link capacity, robustness, and power efficiency in practical settings [31, 41, 61]. Second, pursue multiphysics coupling

and multidimensional modulation. Add magnetic, strain, and acoustic control channels to the electrical–thermal–optical stack, and co-design transient dispersion and phase in time to make strong-field control, nonlinear spectroscopy, and real-time sensing more reproducible and engineering-ready [40, 42, 82]. Third, develop chip-to-free-space interconnects and a standardized ecosystem. Use SW bridges and BIC-based outcoupling to reduce mode mismatch and divergence, and align efficiency, mode purity, EIRP/BER, and OAM metrology and interoperability. Standardized interfaces would support cross-material/process/platform replication and facilitate networked integration [35, 36, 71].

Furthermore, two directions deserve particular attention for closing the gap between laboratory demonstrations and practical THz systems: patterned ferroelectric nematic LC platforms and spatiotemporal structured beams. Ferroelectric nematic LC films combine low-voltage reorientation with large $\chi^{(2)}$ and a reconfigurable nonlinear PB phase, enabling a single pixelated stack to program phase, amplitude, polarization, and frequency-conversion pathways on-chip; when co-packaged with all-dielectric metasurfaces and SW bridges, they may form compact source-shaping-coupling front ends for adaptive links and mode routing [83–85]. In parallel, spatiotemporal THz beams – wave packets with tailored spatial and temporal characteristics – enable spatiotemporal vortices with transverse OAM, space-time vector beams, and time-varying OAM modes that add orthogonal channels and improve robustness to scattering; together with programmable phased arrays and optical-domain pump mapping, they may support adaptive spatiotemporal control [86–90]. These approaches highlight practical routes toward compact, reconfigurable THz front ends and adaptive spatiotemporal control, which enable measurable gains in capacity, robustness, and power efficiency. In summary, THz structured-beam metasurfaces are a rapidly developing field, with progress from device concepts to system-level demonstrations, and we believe continued co-design of sources, metasurfaces, and interconnects will enable scalable, application-ready systems.

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Metasurface-Based Reconfigurable Antennas and Intelligent Reflecting Surfaces for B5G and 6G Applications

Dilek Uzer and Ayşe Özge Çınar Ak

Abstract

Metasurfaces (MTSs) have become an inviting research area, widely used in a variety of high-frequency communications and other applications for manipulating electromagnetic waves. These structures, typically designed to be as thin as subwavelengths, offer advantages such as easy manufacturing with small volumes and the ability to control wave propagation around the structures on which they are mounted. Today's wireless communication systems increasingly require antennas with versatile functionality, adaptability, and flexibility. Unlike conventional antennas, reconfigurable antennas can instantly change their operating characteristics, such as frequency, radiation pattern, polarization, and their combinations, to meet specific needs. This feature provides a significant advantage for next-generation wireless communication applications. While data rates and communication capacity can be significantly increased in beyond-5G and 5G cellular communication systems, electromagnetic (EM) wave propagation faces greater atmospheric attenuation. In these higher frequency bands, EM waves can be significantly attenuated or completely blocked by physical obstacles. This creates blind spots where the signal cannot reach, leading to line-of-sight limitations, one of the main problems of B5G and 5G communication systems. Reconfigurable Intelligent Surfaces (RISs) play a crucial role in maintaining communication connectivity by creating an alternative LoS. There are multiple techniques for reconfigurable antennas, and in this section, metasurface-based applications and Intelligent reflective surface (IRS) applications, also known as reflective metasurfaces (RMTSs), will be discussed.

Keywords: metamaterial, metasurface, reconfigurable intelligent surfaces, intelligent reflective surfaces, reconfigurable antennas, beyond 5G, 6G

1. Introduction

The first study on metamaterials (MTMs) was carried out by Victor Veselago in the late 1960s, and it was proven that such materials can transmit light [1]. He made

a systematic study on the electromagnetic properties of negative parameter environments. Since previous studies were not systematic, Veselago's work is accepted as the first systematic study. In 1996, Pendry showed that negative permittivity could be obtained with an artificial structure, and in 1999, negative permeability could also be obtained [2]. In 2000, Smith and his colleagues succeeded in producing the first double negative (DNG) material using a split ring resonator (SRR) and wire, and they conducted experimental studies at microwave frequencies [3]. In 2001, Shelby studied negative diffraction at microwave frequencies [4]. MTMs are artificial materials consisting of subwavelength unit cells arranged periodically or quasi-periodically. Unlike natural materials, these structures directly determine their electromagnetic properties by their geometric design; by using the designed unit cells, materials with negative electrical permittivity and magnetic permeability values that do not exist in nature can be obtained [1–4].

Due to the bulkiness and difficulty of producing MTMs, a two-dimensional (2D) MTM has been proposed as the basis for MTMs [5]. MTSs can be defined as the 2D counterparts of MTMs and are unnatural materials formed by the periodic or aperiodic arrangement of meta-atoms at subwavelengths on a 2D surface. Significant progress has been made in this field in the last decade, and MTSs have become widely used in the development of numerous new devices and applications in the middle of the electromagnetic frequency spectrum. Snell's law, which traditionally describes the behavior of light in different environments, was made more general in 2011. It has been shown that MTSs with space gradients can impart an abrupt phase shift to EM waves by creating a phase discontinuity between the environments, thereby directing the wavefront in a desired manner. Today, these structures have evolved into smart platforms that can change phase, amplitude, and polarization in real time by equipping them with active diodes, varactors, or graphene sheets in the form of reflective metasurfaces (RMTSs), not just passive phase control [6]. Thus, the MTM adventure, which began with a theoretical prediction in 1968, has evolved into high-tech MTSs used in both reconfigurable antennas and smart reflective surfaces serving today's 5G and 6G communication systems.

With the advent of the twenty-first century, MTMs and metastructures have been used to significantly optimize antennas' specifications. MTMs are artificially designed structures or materials that possess properties different from those found in nature. In conventional reconfigurable antennas (PIN/varactor switching, switchable feed/topology, etc.), additional feed networks, complex circuits, and additional losses can limit performance. MTSs, whether installed as a superstrate on top of the antenna, integrated into the bottom/ground, or directly attached to the radiation surface, enable more fine-tuned and low-profile modifications of the antenna's operating frequency, radiation pattern, gain, and polarization. In this approach, the metasurface (MTS) directs the waves propagating into the antenna's surrounding area with a phase gradient, shifting the resonance or transforming the polarization by engineering the local effective medium parameters [7].

In post-5G and 6G systems, metrics such as data rate, latency, reliability, capacity, and energy efficiency will become much more challenging. For example, mmWave and THz frequency bands are used; propagation losses, environmental obstructions, and LoS/NLoS situations create significant problems in these bands. In this context, IRSs come into play. IRSs consist of passive or semiactive elements that can reshape the signal path, controlling the reflection phase and sometimes the amplitude. Such systems have the potential to offer lower cost, lower power

consumption, lighter hardware, and easier scalability, beyond traditional techniques such as antenna steering (beamforming) and active relay systems [8].

This introductory chapter aims to present a holistic view of the role of MTS-based reconfigurable antennas and IRSs in B5G and 6G wireless communication applications. At the beginning, it is explained how MTMs and MTSs can manipulate EM waves. Then, the challenging requirements of the B5G and 6G era, such as high data rates, very wide bandwidth, ultralow latency, and attenuation and coverage issues in the mmWave/THz bands, are explained. Emphasizing that these requirements cannot be met solely by conventional antenna solutions, two primary application areas of MTSs are highlighted: (i) MTS-based reconfigurable antennas, low-profile, energy-efficient structures that can directly change the antenna's frequency, gain, radiation pattern, and polarization; and (ii) IRSs, wide-area solutions that passively or semiactively control the path and phase of electromagnetic waves in the communication environment, improving coverage and link quality. This approach, supported by current findings in the literature, demonstrates that MTSs, with their antenna-level reconfiguration and channel-level signal routing capabilities, are a critical technology for meeting the high-performance, flexible coverage, and energy-efficiency objectives of B5G/6G networks. Thus, the chapter provides the conceptual and technical foundation for the research and design work to be presented in this thesis by clarifying the complementary roles of MTS-based reconfigurable antennas and IRSs in B5G and 6G applications.

2. Fundamentals of metasurfaces

2.1 Metasurface concept and structural properties

A MTS is an array of MTMs that emit radiation, with dimensions and element spacing values smaller than the operating wavelength. MTSs can be thought of as 1D or 2D MTMs designed to modify the scattering properties of surface waves or the reflection/refraction properties of space waves. This means that an antenna does not need to have a negative refractive index to be called as a MTS [9]. The term “metasurface” is now widely used for homogenizable, electrically thin surfaces with superior properties and represents the 2D counterpart of MTMs [10]. The functionality and electromagnetic properties of MTSs can be modified or rearranged by adjusting the structure and dimensions of the MTS [11]. Recent technological and scientific advances have revealed the considerable utility of MTSs in controlling electromagnetic waves [12].

Studies on MTMs have highlighted several inherent disadvantages, including high losses, strong dispersion due to the resonance response that occurs when the natural frequency is reached, and the difficulty and cost of producing small (nano)-scale 3D particles. MTSs are used to overcome these similar limitations of classical MTMs. Another advantage MTSs have over MTMs is their easier fabrication. This advantage has been achieved through classical lithography and nanopress technology [13].

Under normal circumstances, MTSs are designed to perform specific electromagnetic and optical functions before fabrication. This normality stems from the constant alignment at the atomic level. However, scientists are searching to break this norm and fabricate MTSs that are dynamically aligned at the atomic level,

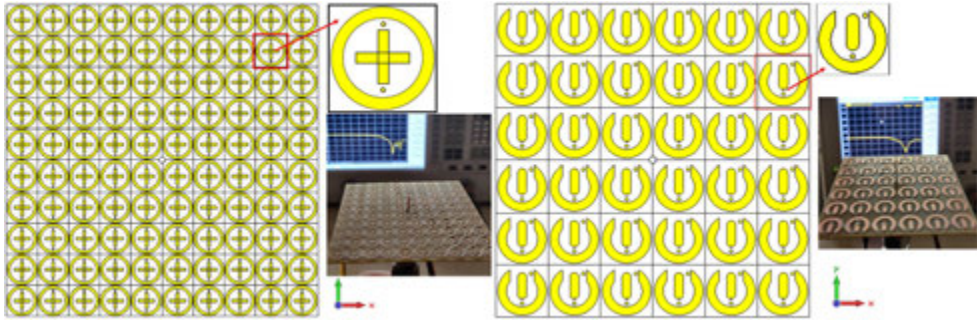


Figure 1.
Example of a metasurfaces with its unit cells image [11].

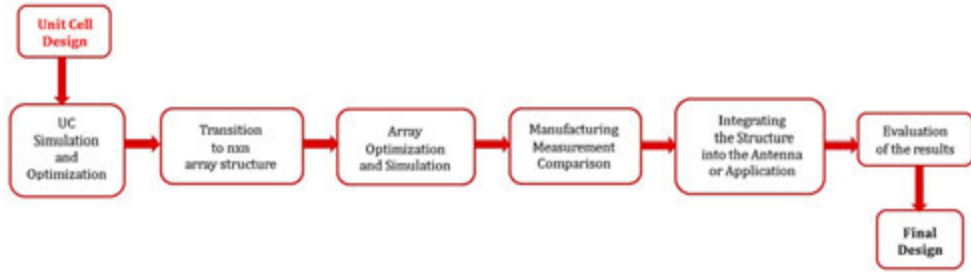


Figure 2.
A general metasurface-integrated structure design steps.

have variable dimensions, are flexible, and can be stimulated by external influences [14]. **Figure 1** shows two sample unit cells and a 10×10 MTS composed of these cells. The yellow parts represent the conductive material, while the white parts represent the dielectric material. As illustrated in **Figure 1**, the MTS is obtained by periodically arranging subwavelength-sized unit cells. The geometry of each unit cell affects the surface's electromagnetic behavior; for example, in the unit cell shown in **Figure 1**, the circular ring determines the resonant response, while the cross element at the center of the structure can provide a polarization-selective effect. When these cells are used in a larger array, the phase behavior of the incident waves can be changed, and the reflection or refraction angle can be controlled as desired within the framework of generalized Snell's law [15–17]. Thus, MTSs offer an effective platform for functions such as improving antenna performance and reconfiguring phase and polarization [6]. From **Figure 2**, generalized steps for designing and integrating MTSs into a structure can be seen.

Since the early 2010s, interest has shifted from passive structures to active, time-varying designs. Additionally, MTSs can be easily manufactured to be electrically thin, making them a favorite in today's communication applications. MTSs provide comprehensive control over reflection, transmission, and surface waves. Surface waves are transverse waves that propagate along an interface and decay exponentially away from it. A characterization matrix can be used to determine the surface impedance

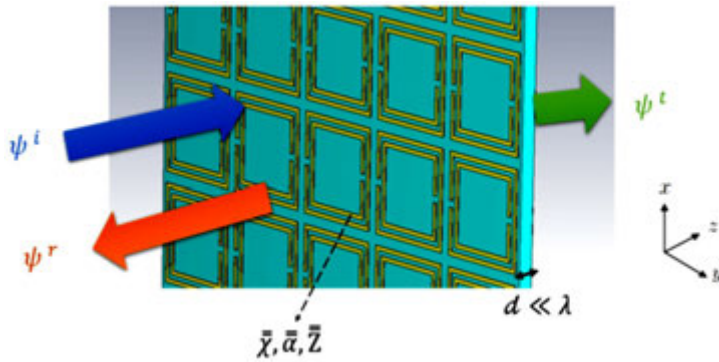


Figure 3.
 An illustrative example of a metasurface in free-space with characterization matrices [10].

required to achieve the desired field transformations on each side [10]. MTSs are also used to optimize the parameters of antennas or to control the properties of EM waves [18].

Figure 3 shows an example design of an electrically thin MTS for the condition $d \ll \lambda$. This surface can be electromagnetically characterized by models based on susceptibility (χ), polarizability (α), and impedance (Z). These parameters describe the MTS's response to electromagnetic waves from different analytical perspectives and can be transformed into one another. The areas indicated by the blue (ψ^i), red (ψ^r), and green (ψ^t) arrows represent incident, reflected, and transmitted waves, respectively [10]. In the most general bianisotropic case, since the electrical and magnetic responses can be cross-coupled, these parameters are expressed as tensors $\bar{\chi}$, $\bar{\alpha}$, $\bar{Z}$ and thus the MTS can respond asymmetrically to waves of different polarizations. MTSs are used to improve antenna performance, provide beam shaping, and add additional functionality by engineering the amplitude, phase, and polarization of these waves [17].

MTSs are characterized by sheet impedance and boundary conditions. MTSs are easier to fabricate by patterning periodic structures on planar substrates [19–22]. They can be single or multilayered, but the total thickness must be much smaller than the operating wavelength [23]. They have low loss in the propagation direction. Despite their planar structure, the interaction of MTSs with EM waves is quite strong, providing useful functionalities [24].

In the analysis and design of MTSs, an equivalent circuit model is used, which combines resistance, capacitance, and inductance according to transmission line theory. In this approach, the electromagnetic response of the subwavelength elements is represented by the resistance (R), inductance (L), and capacitance (C) elements. For example, currents induced in ring resonators or conductive strips create inductive effects, while gaps between adjacent elements create capacitive effects. The resistances used represent both ohmic losses and the loading added for active reconfiguration. This method allows complex electromagnetic interactions to be modeled with a simple series RLC circuit, allowing parameters such as resonant frequency, bandwidth, and losses to be easily predicted. **Figure 4** shows multiple equivalent circuit models of MTS unit cells developed according to the same electromagnetic principle [25–27], as comprehensively discussed in the literature [17]. The figure illustrates representative geometries such as ring, concentric-ring, and nested square-ring configurations together with their corresponding equivalent

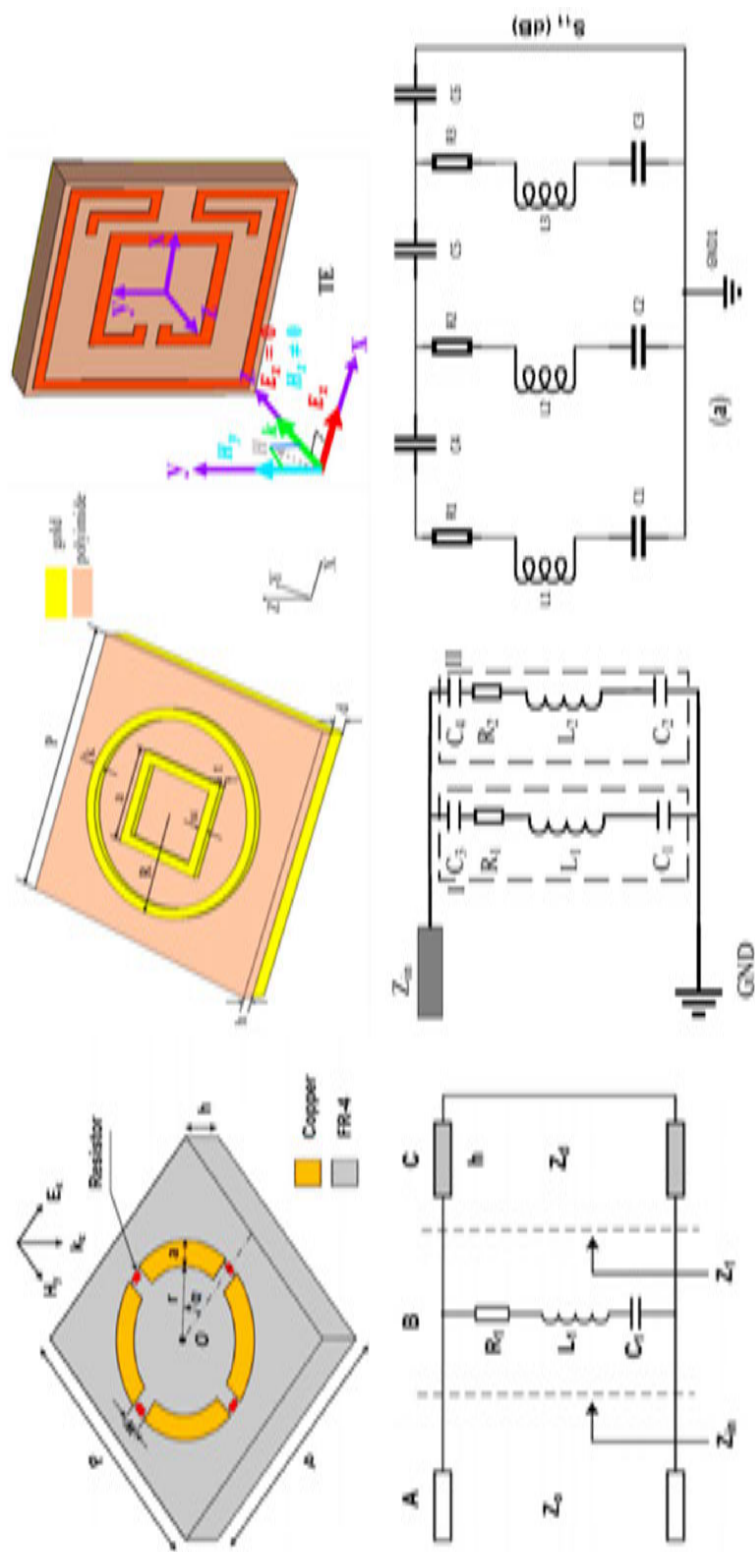


Figure 4. The MTS unit cell in 3D view and the equivalent circuit model reported in: (a) [25], (b) [26], and (c) [27].

circuit representations. In these models, the conductive traces of the MTS elements are represented by inductive components (L) due to the surface current flow, while the gaps or discontinuities between the traces are modeled as capacitive components (C). The resistive terms (R) account for the ohmic and dielectric losses. These equivalent-circuit formulations enable accurate prediction of the resonance frequencies, coupling effects, and bandwidth of different MTS configurations without requiring full-wave simulations.

Conventional Snell's law is used to relate the angle of incidence and refraction of EM waves as they pass through the interface between two isotropic media [24]. MTSs can also be designed to provide anomalous refraction of light according to Snell's law in transmission mode. Impedance matching between free space and the MTS input must be ensured to prevent reflection.

In order to study the physics behind the absorption mechanism, an equivalent circuit model of a unit cell of the metamaterial absorber (MMA) was established [28–32], using transmission line theory, as depicted in **Figure 4(a)**. The absorbance of the proposed MMA is given by Eq. (1).

$$A(w) = 1 - R(w) = 1 - \left| \frac{Z_{in}(w) - Z_0}{Z_{in}(w) + Z_0} \right|^2 \quad (1)$$

where

$$\frac{1}{Z_{in}(w)} = \frac{1}{Z_m(w)} + \frac{1}{Z_d(w)} \quad (2)$$

$$Z_m(w) = R + j\omega L + \frac{1}{j\omega C} \quad (3)$$

$$Z_d(w) = j\sqrt{\frac{\omega_r \omega_0}{\epsilon_r \epsilon_0}} \tan(k_r h) \quad (4)$$

$$k_r = k_0 / \sqrt{\epsilon_r \omega_r} \quad (5)$$

ϵ and ϵ_0 and ϵ_r are the relative permittivities, ω_0 and ω_r are the relative permeabilities, and k_0 and k_r are the wavenumbers of the free space and the dielectric substrate, respectively. From Eq. (1), $A(w)$ can be written as

$$A(w) = \frac{4Z_0 \left[\text{Re}(Z_{in}(w)) + i\text{Im}(Z_{in}(w)) \right]}{\left[\text{Re}(Z_{in}(w)) + i\text{Im}(Z_{in}(w)) \right]^2 + 2Z_0 \left[\text{Re}(Z_{in}(w)) + i\text{Im}(Z_{in}(w)) \right] + Z_0^2} \quad (6)$$

Eq. (6) can simply be rewritten as proposed by Li et al. [33] as follows:

$$A(w) = \frac{2Z_0}{\text{Re}(Z_{in}(w)) + i\text{Im}(Z_{in}(w)) + Z_0} \quad (7)$$

where $\text{Re}(Z_{in}(w))$ and $\text{Im}(Z_{in}(w))$ are the real part and the imaginary part $Z_{in}(w)$ respectively. According to Eq. (1), in order to obtain the perfect absorption [$A(w) = 1$] over all frequencies, the impedance matching condition

$Z_{in}(w) = Z_0$ should be satisfied. This condition can be achieved when $\text{Re}(Z_{in}(w)) = Z_0 = 377\Omega$ and $\text{Im}(Z_{in}(w)) = 0$, according to Eq. (7). (a) (b) (c)

2.2 Common applications of metasurfaces

MTSs are preferred in many applications in today's communication systems. The most common areas of use are listed below:

Precise scattering management: From microwave to optical frequencies, they can block, absorb, concentrate, scatter, or direct normal and oblique waves at the surface and in space. MTS unit cells can be designed with appropriate impedance to adjust the phase or group velocity of surface waves. For directing or splitting waves in specific directions, they are used in applications of scattering management [34].

For surface waveguides: MTSs, formed in compatible impedance unit cell designs, are used to adjust the phase or group velocity of surface waves. They can direct or split waves in predetermined directions through their patterns in scattering control applications. By modifying unit cell designs, they can achieve different surface refractive indices and serve different functions [34].

Controlling plane waves: MTSs can be used for controlling plane waves in free space. For use as ultrathin polarization transformers, the reflection phase of waves impinging on the MTS can be adjusted. Incoming waves with linear polarization can be converted to circularly polarized waves by rotating them 90° . MTSs are also being studied in depth as a potential tool for wavefront engineering, which involves beam shaping through defocusing, focusing, refraction, and reflection. Covering curved and planar surfaces with MTSs reduces scattering, which is useful for cloaking and other applications in the microwave and far-infrared frequency ranges. MTS absorbers can also be paired with active circuits or nonlinear devices, as they have proven to be superior to conventional absorbers [34].

Wave propagation: Directing the radiation field by converting surface waves into radiation modes and operating as a leaky wave antenna can be achieved through the use of MTSs. Traditional electronic components, such as lenses and waveguides, can be bulky and limit the miniaturization of devices. With their ability to manipulate light at the nanometer scale, MTSs offer a way to design compact and lightweight components. This is particularly valuable in the development of miniature sensors, display devices, and communication modules [34].

3. Metasurface-based reconfigurable intelligent surfaces

RISs constitute a class of programmable MTSs capable of manipulating the propagation of EM waves through dynamic modulation of their surface impedance properties. By altering the amplitude, phase, and polarization of incident EM waves, RISs enable the real-time reconfiguration of the wireless propagation environment, thereby enhancing network performance and spectral efficiency. Recent experimental studies and prototyping efforts have verified that RIS implementations exhibit low power consumption, minimal hardware complexity, and cost-effectiveness, making them highly suitable for deployment in 5G and emerging 6G communication networks [35]. The operational modalities of RISs are illustrated in **Figure 5**.



Figure 5.
Operation modes of RIS.

RISs are alternatively referred to as IRSs [36], software-controlled metasurfaces (SCMs) [37], or large intelligent surfaces (LISs) [38]. These surfaces achieve dynamic EM wave manipulation through programmable unit-cell architectures, allowing for adaptive phase, amplitude, and polarization control. Such capabilities facilitate precise shaping of the RF propagation environment, which is critical for improving link reliability, beamforming efficiency, and interference mitigation in next-generation wireless systems [35]. RISs represent a broader and more advanced form of IRSs; they can realize functions such as reflection, absorption, refraction, focusing, and polarization conversion. When used with active components (e.g., amplifiers, RF switches, PIN diodes, MEMS), they can perform amplification or digital processing. Typical uses of RISs include optimizing coverage, data rate, and energy usage in 5G/6G networks [39–41].

In *reflection* mode, as shown in **Figure 5**, RIS alters the direction of reflection by adjusting the phase of the incident wave on a cell-by-cell basis, enabling beam steering through phase gradient engineering, unlike conventional mirrors [42]. In *diffraction* mode, the phase profile can be adjusted to change the angle of refraction of the transmitted wave [43]. In *absorption* mode, unit cells are designed to resonate, converting the incident energy into heat, which is used in radar-absorbing applications [44]. In *backscattering* mode, RIS can direct the incident wave back toward the source, which is useful in RFID or passive sensing applications. In *transmitting* mode, RIS allows the wave to pass while simultaneously reshaping its phase. In *receiving* mode, RIS can collect incoming waves and focus them in specific directions by acting like antenna arrays [45]. In Section 5, under **Table 1**, literature examples of these RIS applications are also provided.

Future 5G wireless networks are expected to exploit the THz band as a core enabler for accommodating the rapidly increasing wireless traffic, thereby supporting ultrahigh data rates. Nevertheless, the intrinsic propagation characteristics of the THz spectrum, particularly its high attenuation, result in substantial path losses, which pose significant challenges for wireless system performance [46].

RISs have emerged as a viable solution to mitigate the limitations associated with THz propagation by dynamically controlling wavefront direction, thereby directing signals toward intended receivers. RIS technology holds considerable potential for enhancing wireless network performance, especially when integrated with UM-MIMO antenna systems. The deployment of UM-MIMO is essential for THz communications, as large antenna arrays can generate highly directive, pencil-like beams that efficiently transmit data between transmitters and receivers. To reduce

Reference	Design methods	Goals	FoM	Advantages	Disadvantages
[59]	MTS	Refraction and transmitting for indoor coverage enhancement	A receiver power increase of approximately + 20 dB, optical transmittance of over 74%		
[73]	MTS + SIW	Optimizing bandwidth, gain, and efficiency	Larger bandwidth for higher frequency	Enhancement of gain, radiation efficiency, and impedance bandwidth	Inherent dielectric losses of high-permittivity substrates and limited fabrication precision at THz frequencies
[75]	MTS Wall	Improving insulation between antenna arrays	An isolation improvement of more than 10 dB	Isolation level below – 63 dB Compact design	Fabrication tolerances and limited reconfigurability for adaptive array systems
[76]	MTS + SIW	Improved efficiency and gain with decoupling reduction	Improvement of 13.5, 6.5, and 10 dB for bandwidth, gain, and efficiency	High isolation and enhanced radiation performance Compact and symmetrical structure	Fabrication complexity at THz scale Substrate limitations
[77]	MTS + AoC	High-efficiency THz on-chip antenna design using the meta-surface concept	An average improvement of 10.8% and 39.2% in gain and radiation efficiency Min 10 dBi gain, 60% efficiency	Performance enhancement	Manufacturing is not easy and is expensive
[78]	MTS + AoC	Gain and efficiency optimization without changing the size and compatibility with integrated circuit structures	8.15 dBi radiation gain and 65.71% efficiency Simple structure	Enhancement of radiation efficiency Elimination of back-lobe radiation	Manufacturing is not easy and is expensive
[79]	PIN Diode + MTS-RIS	Reflection and experimental beam steering and mmWave coverage enhancement in the near/far field	Beam steering $\pm 60^\circ$, max 30.5 dBi gain, 1-bit phase control	Easy reconfiguration	Needs electronic control units and biasing circuits fast enough
[80]	Programmable MTS	Transmitting and reflecting for real-time ray tracking and direction control	Real-time beam steering 80° phase adjustment		Needs microcontroller units

Reference	Design methods	Goals	FoM	Advantages	Disadvantages
[81]	Multilayer MTS-absorber	Broadband absorber for EMC/RCS reduction	>90% absorption	Flexible and transparent structure	Multilayer manufacturing and higher dimensions
[82]	Hybrid RIS + reflective and sensing elements	Passive detection and channel reconfiguration	Hybrid structure		

Table 1.
Comparative summary of case studies in the literature.

implementation complexity in UM-MIMO architectures, hybrid precoding schemes are recommended [46].

Reconfiguration of free-space waves emitted from antennas or light sources is vital for the development of modern communication systems and remote energy-based applications [10]. Reconfigurable antennas are structures that can electronically or mechanically change their operating frequency, polarization, or radiation direction. MTSs have been extensively used in recent years to improve the performance of these antennas [39–41].

One of the main goals of adaptive antenna systems is to flexibly vary the amplitude, phase, and polarization properties of aperture fields. This is because the direction, shape, and polarization of the waves emitted by antennas are directly dependent on the phase and amplitude distribution of these aperture fields. Traditionally, in high-gain millimeter-wave and microwave antennas, these aperture fields have been created using lens antennas, dishes, transmitter arrays, and reflector arrays. However, these methods often require bulky and heavy structures.

Recently developed MTS technology has enabled the design of lighter, more economical, and more compact antenna systems by controlling the direction of electromagnetic waves through sharp phase transitions. MTSs integrated with antennas can transform input fields into aperture fields using leaky waves. This transformation utilizes the local properties of the MTS (homogeneity and anisotropy) to achieve desired amplitude and phase distributions in the aperture fields. This allows for precise control over the characteristics of guided waves [24].

MTS-based antennas, while offering small size, low cost, and flexible production capabilities, can be used in a lot of applications, from surveillance systems to radar applications, from airborne platforms to satellite communications. In summary, the integration of MTSs into antenna systems offers the following advantages:

- Beam steering:* MTSs can dynamically change the antenna’s radiation direction.
- Frequency tunability:* thanks to reconfigurable MTSs, antennas can operate at multiple frequencies or provide broadband support.

Compact design: high gain and efficiency can be achieved without increasing the antenna size.

5G and 6G applications: MTS antennas are gaining prominence for providing better coverage and directionality in mmWave and THz bands.

For these reasons, MTSs have become a key technology that increases the flexibility of reconfigurable antennas in next-generation wireless communication systems [12, 24, 39–41, 47].

4. Metasurface-based intelligent reflective surfaces

4.1 IRS concept and basic operating principles

IRSs have been identified as a cost-effective solution for mitigating signal attenuation caused by obstacles and enhancing coverage in wireless networks [48]. Unlike conventional antenna arrays and phased arrays, IRSs are typically composed of planar MTM elements, referred to as unit cells, whose physical dimensions and spacing are significantly smaller than the operating wavelength. Each IRS unit cell functions by passively directing incident signals toward the desired path, achieving reconfigurability through precise control of the reflected wave phase [47].

IRSs essentially operate on the principle of reflection. They can redirect incoming electromagnetic waves using adjustable phase-shifting elements. IRSs are passive or semipassive structures and do not generate signals themselves; they merely direct the existing signal, changing its phase and amplitude. Typical applications of IRSs include expanding coverage in enclosed spaces and transmitting signals to shadowed areas.

IRSs have emerged as a crucial technology for next-generation wireless communication systems, primarily due to their capability to dynamically manipulate EM waves in real time. These surfaces, which can be realized as passive, active, or hybrid multielement reflective arrays, offer a cost-effective means of enhancing signal quality and coverage when optimally deployed [49]. Passive IRSs exploit the geometric and material characteristics of their constituent elements to reconstruct wavefronts, proving particularly effective in complex urban environments where signals are frequently obstructed by buildings or other large structures [12]. In contrast, active IRSs employ tunable components such as varactor diodes, PIN diodes, or MEMS to modulate both the phase and amplitude of reflected waves, enabling the active suppression or cancellation of interference [50].

IRS technology has demonstrated versatility across various communication scenarios, including terrestrial cellular networks, satellite links, and vehicular communications [51]. Among these applications, cellular networks in densely built urban areas represent one of the most critical deployment scenarios, as physical obstructions can significantly impede signal propagation. IRSs can be used in cellular communication networks for purposes such as privacy enhancement, interference reduction, beamforming integration, phase shifts, and real-time adaptation [51].

Although the mmWave frequency band offers the advantages of high data transmission capacity and wide bandwidth, it has a significant limitation: substantial electromagnetic losses encountered during transmission and reception at high frequencies. These losses are beyond the reach of existing optimization algorithms, leading to a significant decrease in system performance. In this context, IRS stands out as a next-generation hardware technology with the potential to improve the poor propagation conditions encountered in the mmWave band [52].

IRS can transform a statistically uncertain environment into a more stable and controllable one by modifying the nature of wireless communication channels through passive and reconfigurable surfaces. IRS technology, used to reduce signal attenuation during propagation, particularly in the mmWave band, has attracted significant interest in both academic circles and industry in recent years, thanks to advances in MTM engineering [53].

IRS systems, which stand out with their advantages such as low installation cost, energy efficiency, and increased signal coverage, not only improve communication performance but also contribute to the creation of sustainable communication infrastructures by reducing energy consumption [54]. In the literature, this technology is also referred to by different terms, such as reconfigurable reflective surface, LIS, and SCM.

Innovations in artificial EM materials have enabled IRS to possess a structure that is digitally programmable and reconfigurable according to environmental conditions. This has allowed the communication environment to be managed digitally, rather than physically, and the adaptability of communication systems to their environment has been significantly enhanced [55, 56].

IRSs also provide an alternative approach for relay and backscatter communication systems. An IRS consists of numerous small, reconfigurable, passive, and low-cost reflective elements, each capable of independently imparting a controlled phase shift to incident electromagnetic waves. By modifying the propagation characteristics between the transmitter and receiver, these elements enhance signal strength while mitigating interference. Each unit reflects the incoming wave with a tunable phase, enabling precise control over the reflected signal direction.

As IRSs function passively, they can reflect incoming signals without the need for an external power source to perform RF processing, encoding/decoding, or retransmission. This passive operation makes IRSs a highly energy-efficient alternative to traditional amplify-and-forward (AF) relay systems, as they merely redirect the received signals without engaging in active modulation [56]. Additionally, by intelligently tuning the phase shifts through a centralized control mechanism, IRSs can execute dynamic beamforming techniques that improve signal quality, facilitate adaptive wireless connections, and mitigate interference among multiple users. These features establish IRS technology as a key enabler for next-generation wireless communication networks [54], which constitutes the main focus of this study [52].

4.2 IRS system architecture and design

The recent surge in EM-MMs can be attributed to their exceptional ability to manipulate EM waves in unconventional manners. Such capabilities enable functions including phase modulation, amplitude modulation (AM), and polarization conversion of EM waves [57]. Digital and programmable MTMs can be encoded through simplified designs with optimized phase reflection coefficients, providing a streamlined and effective approach for EM wave control across various applications, particularly in the mmWave and THz frequency bands. These functionalities can be independently governed via field-programmable gate array (FPGA) implementations, establishing a direct interface between physical EM phenomena and digital control systems [39, 52, 57].

Figure 6 depicts an MTS consisting of a large array of passive elements capable of dynamically steering incident EM waves in real time [49]. In scenarios such as that illustrated in **Figure 6(a)**, where a clear LoS exists between a base station (BS) and a user device, the deployment of an IRS enables the generation of a replicated signal path, thereby increasing system diversity. Conversely, as shown in **Figure 6(b)**, when the LoS is obstructed by a tall building, the IRS facilitates the establishment of

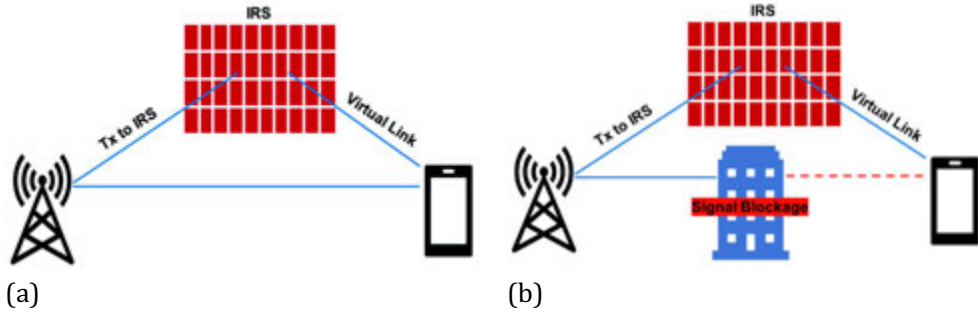


Figure 6.

IRS-aided wireless transmission system: (a) in the presence of LoS link; (b) in the absence of LoS link [52].

a virtual link, effectively bridging the communication path between the BS and the user device.

This virtual link will ensure that the receiver can receive the transmitted signal, improving the overall EE and SE of the system, which would not be possible without the IRS [52].

4.2.1 Layers of IRS

A conventional IRS is typically composed of three fundamental layers, as illustrated in **Figure 7**:

- Layer 1 comprises an extensive array of passive, reconfigurable patches fabricated on a dielectric substrate, designed to manipulate incident electromagnetic waves [49].
- Layer 2 consists of a copper sheet, which serves to minimize energy losses during the reflection process [49].
- Layer 3 includes a control circuit board responsible for dynamically adjusting both the phase and amplitude of the reflected signals. A FPGA can function as an intelligent controller, managing the reflection characteristics and system configuration while also acting as a communication interface between the BS and the target receiver [49].

In this section, examples of current studies containing the structures mentioned in the previous sections are given.

Traditional reflectarray antennas often suffer from inherent drawbacks such as fixed beam orientation, considerable insertion loss, and complicated phase-shifting circuitry, which limit their effectiveness in wideband wireless communication applications. To overcome these issues, Iqbal et al. [58] introduced an innovative broadband planar MTS designed as a polarization-rotation reflective metasurface (PRRM). This design integrates a 90° polarization rotation capability with 1-bit reconfigurable phase modulation. The MTS is built using a mirror-symmetric unit

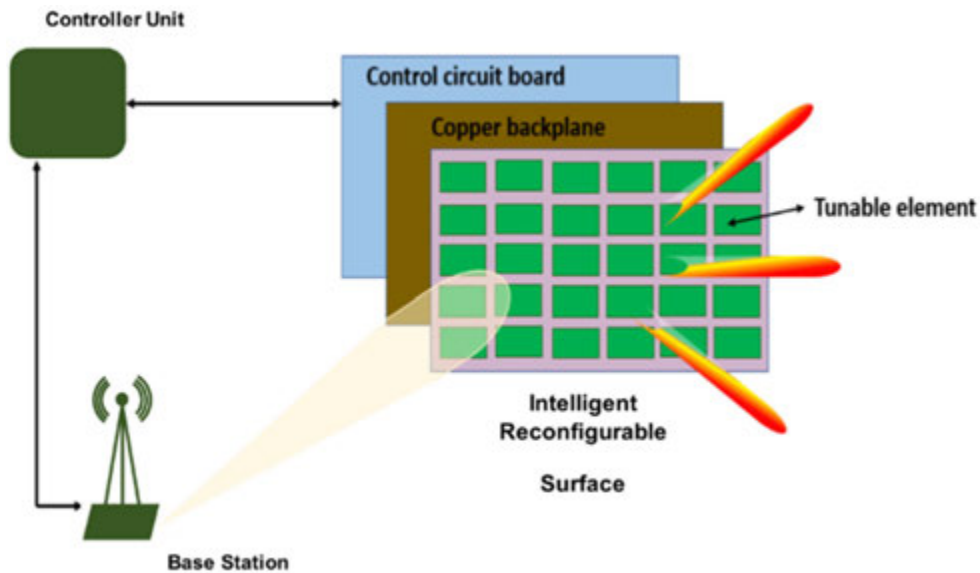


Figure 7.
 Architecture of an IRS [52].

cell that features a fan-shaped cruciform patch loaded with short-circuits and equipped with PIN diodes linked through vertical interconnect accesses (VIAs). As a result, it enables stable binary phase manipulation with minimal signal loss across a broad frequency spectrum. Simulation results verify that the proposed unit cell sustains consistent cross-polarized reflection and a smooth phase transition between 3.83 GHz and 15.05GHz, yielding an impressive fractional bandwidth of 118.89%. Moreover, full-wave simulations of a 16×16 array confirm effective 2D beam steering up to $\pm 60^\circ$ and a 3 dB gain bandwidth of 55.3%. These findings highlight the MTS's strong potential for advanced beamforming, making it a promising solution for compact, electronically reconfigurable antennas used in high-speed wireless communication, radar imaging, and sensing technologies. **Figure 8** illustrates the proposed structure and selected performance outcomes.

5. Current studies and literature on metasurfaces-based antennas, RISs, and IRS structures

Millimeter-wave signals experience substantial attenuation when passing through conventional building materials, resulting in poor indoor coverage from outdoor networks. To mitigate this issue, Yang et al. [59] proposed a 28 GHz “Meta-Window” – a transparent, mass-producible structure engineered to enhance mmWave signal focusing. The device is fabricated through metal sputtering and etching on a standard soda-lime glass substrate, featuring subwavelength metallic elements arranged in a rotational pattern following the Pancharatnam–Berry phase principle, enabling full $0\text{--}360^\circ$ phase modulation within the 25–32 GHz band. A prototype measuring 210 mm $\times$ 210 mm and composed of a 69×69 MTS unit cell

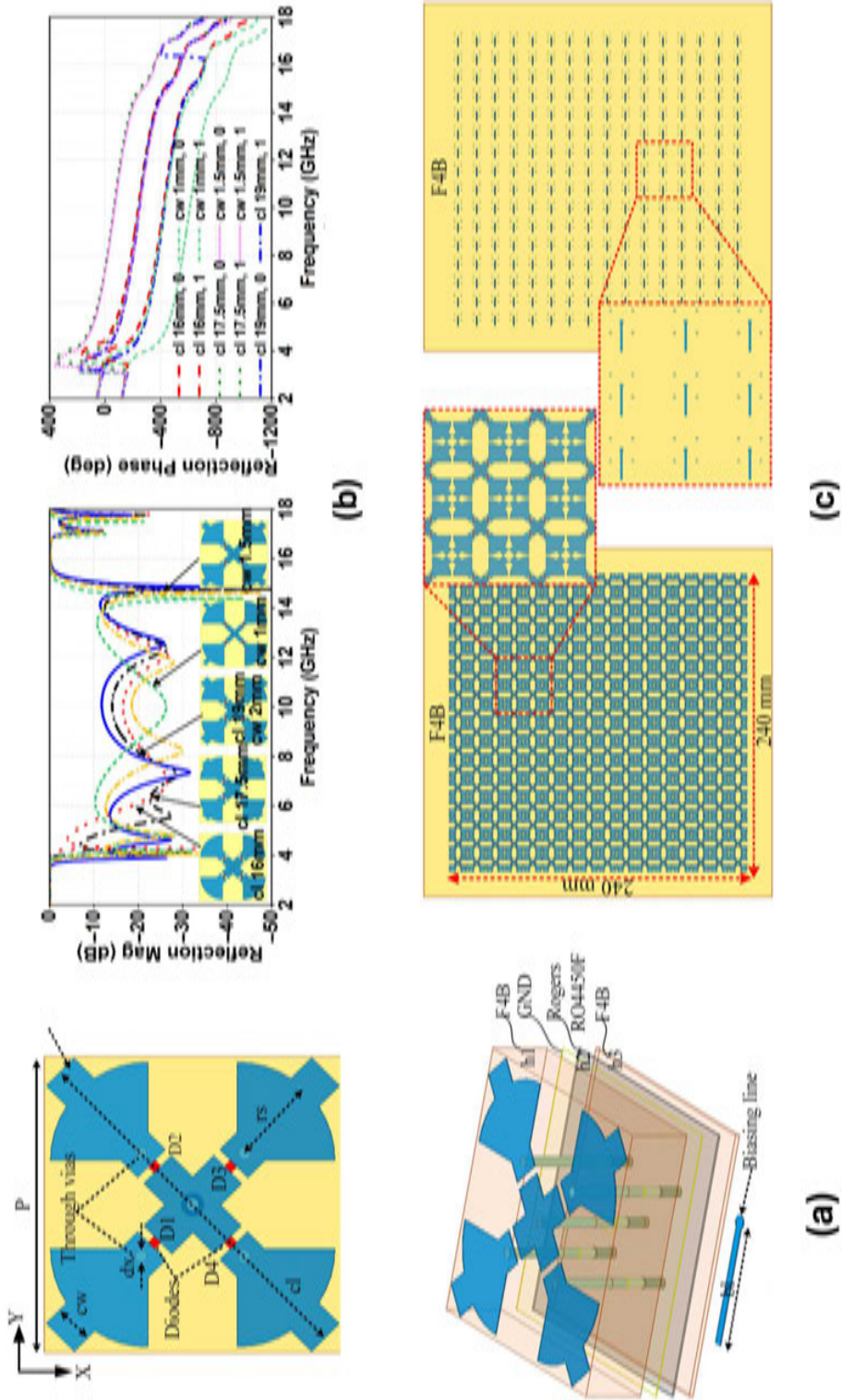


Figure 8. a) Configuration of the proposed RRM 1-bit element, b) the performance analysis, and c) array configuration of the design [58].

array was developed based on planar electromagnetic lens concepts. Measurements confirm that the Meta-Window provides over 20 dB of signal focusing gain from 26 to 30 GHz, aligning closely with full-wave electromagnetic simulations, while maintaining visible-light transmittance of up to 74.93%. Its dual transparency to both optical and mmWave frequencies was further verified through a communication system prototype, which achieved more than a 20 dB improvement in signal-to-noise ratio and successfully demodulated a 64-QAM single-carrier signal (1 GHz bandwidth at 28 GHz) with an error vector magnitude of 4.11%. In addition, combining the Meta-Window with a reconfigurable RMTS antenna array enabled wide-angle beam steering, maintaining reliable demodulation performance (error vector magnitude below 6.32%) across a $\pm 40^\circ$ range using the same transmission parameters. Unlike traditional MTSs, this approach utilizes well-established glass fabrication methods and offers strong potential for direct architectural integration, presenting a viable route to improving indoor penetration and coverage for mmWave communication systems.

Wang et al. [60] introduced a coding MTS optimized for 5G systems at 3.5GHz. The proposed design employs an innovative ring-shaped MTS unit that supports both single-beam and dual-beam functionalities. By applying convolution-based operations, the reflection angle can be precisely tuned within a range of 51.5° – 17.5° , with a fine angular resolution of 10° . This approach surpasses the limitations of conventional gradient coding techniques and establishes a robust framework for wide-angle reflection control in MTS engineering. Simulation and experimental results confirm that the MTS effectively manipulates the reflection direction of electromagnetic waves at 3.5 GHz, exhibits dual-polarization modulation behavior, and sustains stable performance under oblique incidences up to 20° . These findings pave the way for the development of reconfigurable and intelligent MTSs, with promising applicability not only in 5G but also in 5G communication systems and IoT technologies.

To overcome the inherent challenges of CubeSats – particularly the need for reliable and efficient communication within the dynamic and resource-limited conditions of low Earth orbit (LEO) – beam-steering antenna systems have emerged as a highly promising solution. These systems enable adaptive communication links with enhanced gain and coverage. Saeidi and Karamzaden [61] present a comprehensive overview of recent advances in CubeSat communication technologies, with a particular emphasis on beam-steered antenna developments. The paper synthesizes findings from current research, addressing critical issues such as power limitations, system miniaturization, and integration with compact CubeSat architectures. It further explores cutting-edge approaches, including phased array configurations, MTS-based antenna designs, and reconfigurable antenna technologies. This study serves as a valuable reference for researchers and engineers, offering insights into prevailing trends and identifying future directions for advancing CubeSat communication capabilities through innovative antenna solutions.

In a report by Boubekeur et al. [62], a compact on-chip MIMO antenna incorporating frequency-agile features is introduced, developed for next-generation communication systems. Enhanced isolation and a reduced footprint are achieved through the integration of an MTS layer, which enables effective frequency reconfigurability. The antenna operates in the mmWave band and is fabricated using microelectronic techniques on a low-loss Benzocyclobutene (BCB) polymer substrate, with overall dimensions of $7.54 \times 7.54 \times 0.055 \text{ mm}^3$. Both simulation and

measurement results confirm outstanding frequency agility around 60 GHz, with a realized gain ranging from 6.5 to 9 dBi. As an initial proof of concept, switching between open and short circuits was employed, while future implementations aim to incorporate diode-based elements for complete dynamic reconfiguration. This study underscores the promise of miniaturized, high-performance, and frequency-tunable on-chip antennas for emerging mmWave communication technologies. **Figure 9** illustrates the proposed design in both 2×1 and 4×4 array configurations.

Sheng et al. [63] reported a cost-efficient reconfigurable hybrid MTS-based antenna designed for multibeam switching in beamforming applications. The proposed architecture integrates a microstrip-fed antenna with a transmissive hybrid MTS comprising three distinct types of subwavelength unit cells, each exhibiting unique amplitude and phase responses to incident electromagnetic waves, thereby enabling precise manipulation of the feed antenna's radiation pattern. The design incorporates 16 PIN diodes controlled through a simplified biasing network. By applying two discrete DC bias voltages to the diodes, the antenna achieves dynamic switching between dual-beam and quad-beam radiation modes. A prototype was fabricated to experimentally validate the design, and the measured results demonstrate an operational bandwidth of 9.07–9.42 GHz ($|S_{11}| \leq -10$ dB). The antenna exhibits compact dimensions of $1.80\lambda_0 \times 1.52\lambda_0 \times 0.22\lambda_0$, where λ_0 denotes the free-space wavelength at 9.28 GHz. Owing to its compact structure, low fabrication cost, and superior radiation performance, the proposed design represents a promising candidate for integration into advanced beamforming and adaptive wireless communication systems. **Figure 10** illustrates the feed antenna configuration, the unit cell, and the overall structure of the proposed design.

The forthcoming 5G communication networks are expected to revolutionize modern applications and user experiences by enabling intelligent, reconfigurable systems with significantly higher data rates and broader bandwidths. Controllable THz waves have been identified as a key enabler for the rapid advancement of next-generation wireless communication technologies. Recently, RISs have demonstrated outstanding potential in manipulating EM waves for various THz applications,

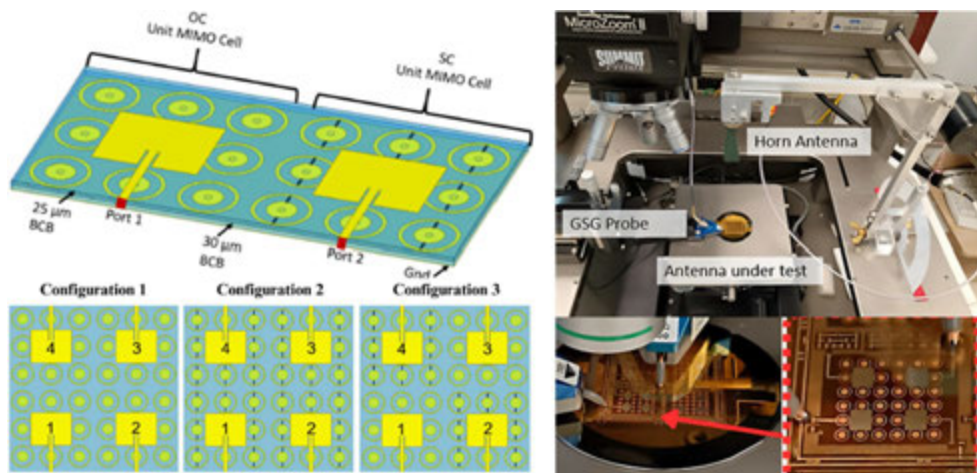


Figure 9.
The proposed 2×2 and 4×4 MIMO configurations [62].

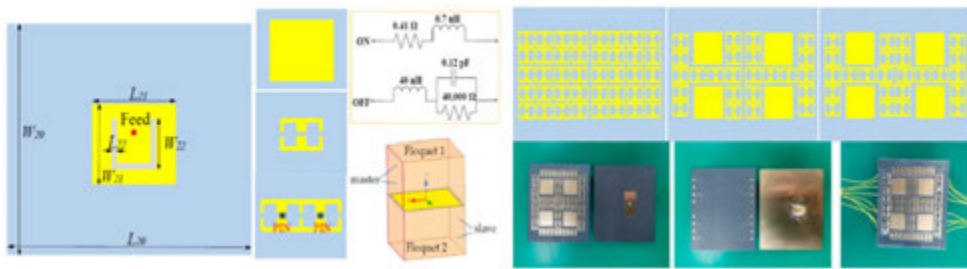


Figure 10.
 The feed antenna, unit cell designs, and hybrid MTS combinations [63].

including beam steering, multibeam scanning, polarization control, and beam focusing. In [64], a graphene-based plasmonic two-port MIMO microstrip patch antenna operating at a resonant frequency of 1.9 THz is proposed. To enhance the isolation between ports, an E-shaped MTM unit cell is introduced, improving isolation performance from -35 dB to -54 dB. Additionally, a controllable and reconfigurable graphene surface (G-RIS) is strategically positioned above the radiating elements to dynamically steer the emitted beam in multiple directions ($\pm 60^\circ$). The reconfigurable behavior is realized by toggling the meta-atom states (ON/OFF), thereby generating specific coding sequences that correspond to defined beam directions. Further gain enhancement is achieved using a graphene-based AMC layer placed beneath the combined MIMO antenna and G-RIS structure, increasing the overall gain from 4.5 dBi to 10 dBi. The proposed design is validated through both full-wave simulations using CST Microwave Studio and circuit-level modeling with ADS, demonstrating consistent and convergent results. As illustrated in **Figure 11**, the proposed MIMO antenna configuration – integrating G-RIS above and G-AMC below – exhibits superior reconfigurability and gain performance, making it a promising candidate for THz and IoT applications. A flowchart of the proposed design is given in **Figure 12**.

In a report by Ledimo et al. [65], an alternative approach utilizing MTSs for wireless communication is presented, where active reconfigurable antennas are proposed for next-generation transceiver architectures. A complete framework describing the design methodology of an MTS antenna is introduced and subsequently applied to the development of a 4×4 antenna array along with its reconfigurable version. Unlike conventional phased array antennas, the RMTS antenna eliminates the need for external phase shifters and amplifiers, thereby reducing system cost and complexity. Instead, each MTS element achieves dynamic reconfiguration by tuning its resonant frequency through semiconductor devices such as PIN diodes. The proposed antenna, illustrated in **Figure 13**, is designed to operate across the 28–40 GHz frequency band. Experimental and simulation results indicate an enhancement in both gain and directivity when the PIN diodes are integrated into the array. However, due to the electrical connection of the diodes with the metallic strips of the antenna surface, additional ohmic losses may occur, leading to a reduction in overall radiation efficiency.

From in **Figure 13**, L_{T1} and L_{T2} represent the total paths transversed by EM waves in the internal and external rings, and $\lambda_g/2$ is the fundamental mode length as in Eq. (8);

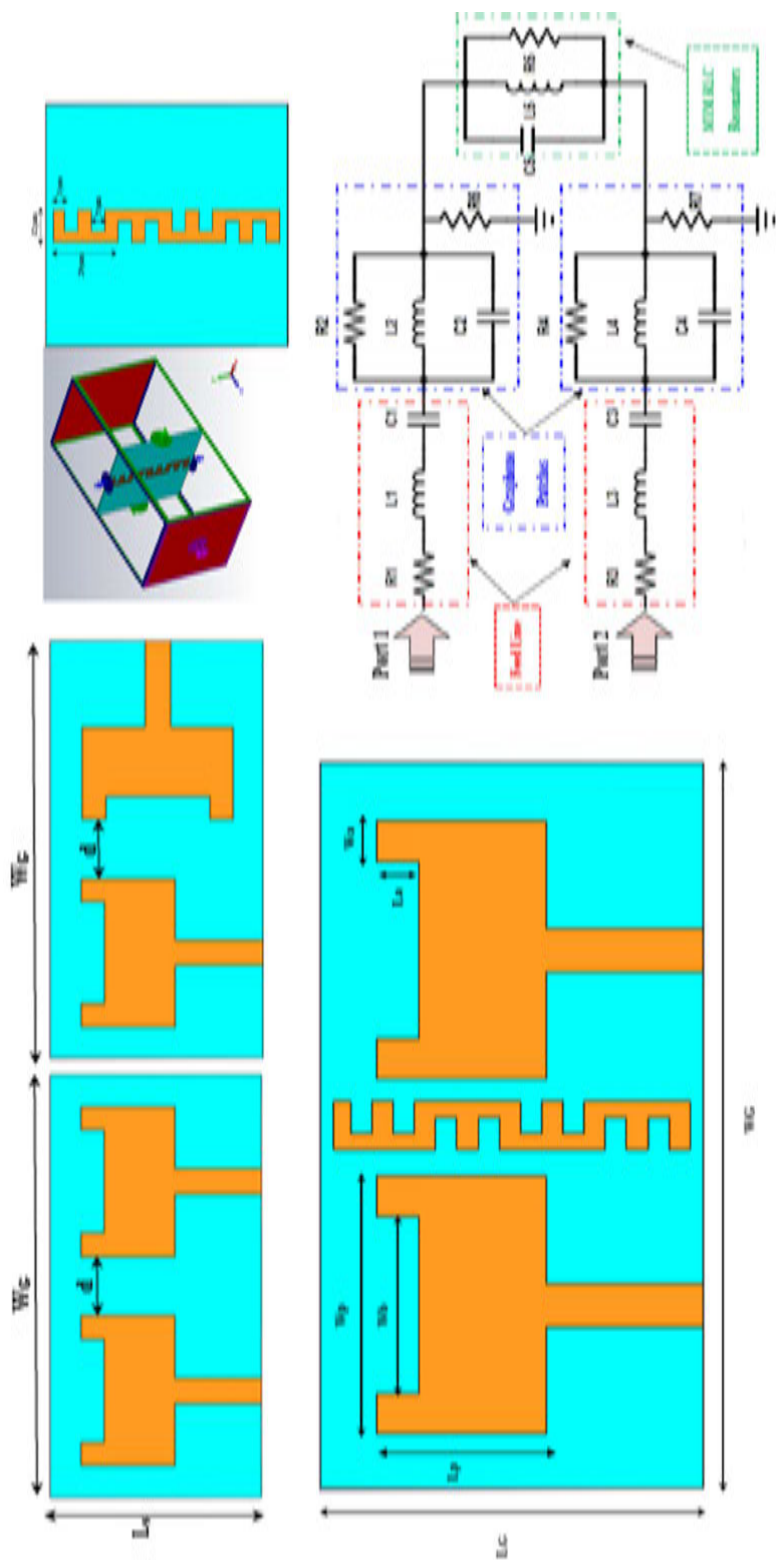


Figure 11. The proposed MIMO design with two combinations: the unit cell, MIMO with decoupling element, and the equivalent circuit model of the design [64].

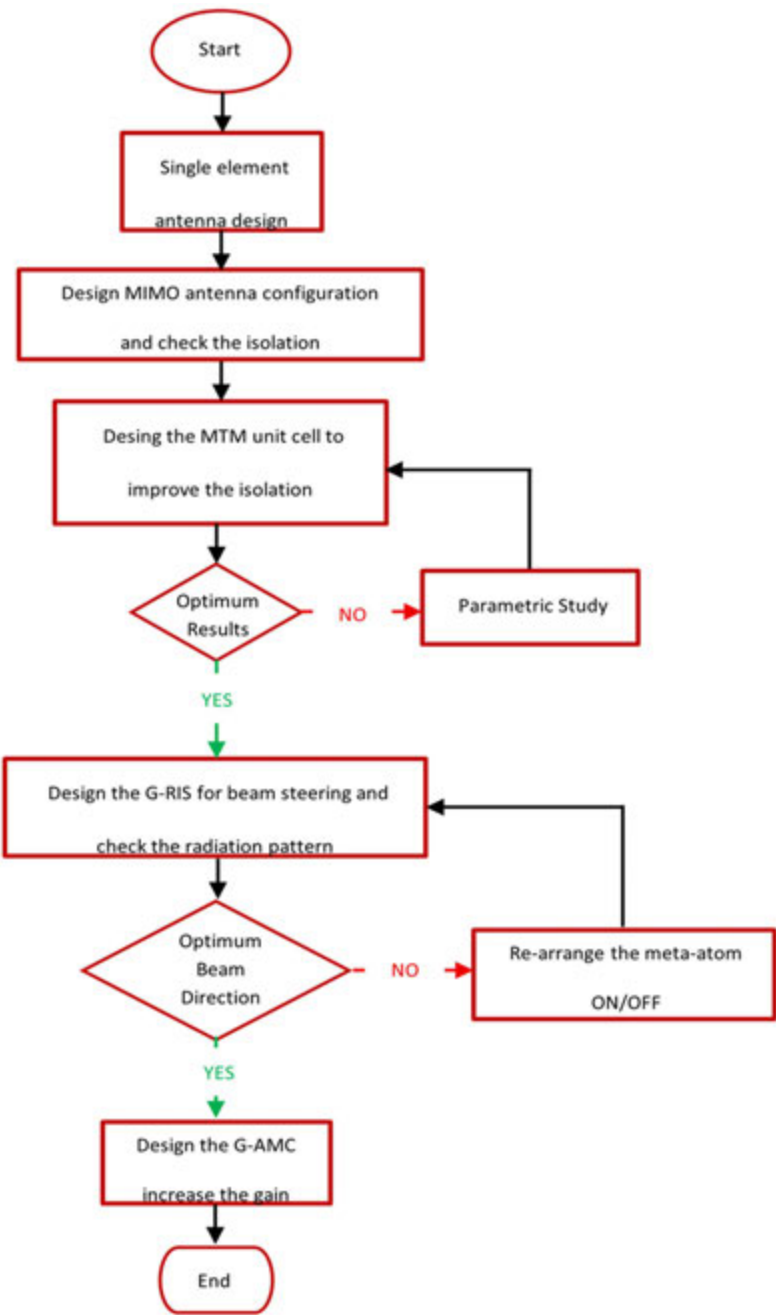


Figure 12.
Flowchart of the proposed design [64].

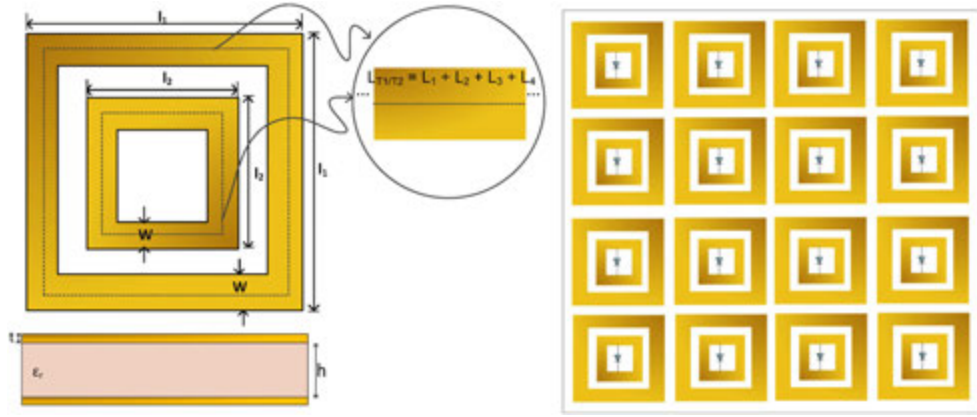


Figure 13.

The proposed unit cell and 4×4 RMS integrated with PIN diodes [65].

$$L_{T1} = L_{T2} = \frac{\lambda_g}{2} \quad (8)$$

The guided wavelength at resonant frequency f_0 is represented by Eq. (9) [66],

$$\lambda_g = \frac{\lambda_0}{\sqrt{\epsilon_{eff}}} = \frac{c}{f_0 \sqrt{\epsilon_{eff}}} \quad (9)$$

The rings' guided wavelengths λ_{g1} and λ_{g2} are given in Eq. (10),

$$\lambda_{g1} = \frac{c}{f_1 \sqrt{\epsilon_{eff}}}, \lambda_{g2} = \frac{c}{f_2 \sqrt{\epsilon_{eff}}} \quad (10)$$

where resonance frequencies of the internal and external rings are defined by f_1 and f_2 respectively. To attain the internal and external ring average length at any specified resonant frequency, Eq. (9) is substituted into Eq. (8) to give [66]

$$L_{T2} = \frac{c}{2f_2 \sqrt{\epsilon_{eff}}} \text{ and } L_{T1} = \frac{c}{2f_1 \sqrt{\epsilon_{eff}}} \quad (11)$$

Furthermore, since the resonators exhibit the same standing waves at the fundamental modes, Eq. (11) is used for both resonators [66]. From Eq. (11), the side lengths l_1 and l_2 of each ring can be calculated by Eq. (12),

$$l_1 = L_{T1} + \frac{W}{2}, l_2 = L_{T2} + \frac{W}{2} \quad (12)$$

The permittivity and permeability are calculated from the refractive index (h) and impedance values z using Eq. (13) and Eq. (14) in order to confirm if the designed SRR represents a DNG MTM.

$$\epsilon = \frac{\eta}{z} \quad (13)$$

$$\mu = \eta z \quad (14)$$

where η and z are denoted as [67, 68]

$$\eta = \frac{1}{kl} \Re \{ \ln(e^{jnl_{\text{out}}}) \} - \Im \{ \ln(e^{jnl_{\text{out}}}) \} \quad (16)$$

$$z = \frac{(1 + S_{11}^2)^2 - S_{21}^2}{(1 - S_{11}^2)^2 - S_{21}^2} \quad (15)$$

From Eq. (15) and Eq. (16), the reflection and transmission coefficients are represented by S_{11}^2 and S_{21}^2 , respectively, $\Re\{\cdot\}$ and $\Im\{\cdot\}$ denote the real and imaginary components, k represents the wave number, and $e^{jnl_{\text{out}}}$ is represented as [69]:

$$e^{jnl_{\text{out}}} = \frac{S_{21}^2}{1 - S_{11} \left(\frac{z-1}{z+1} \right)} \quad (17)$$

In a report by Zhang et al. [70], a hybrid dual-frequency polarized reconfigurable THz antenna is proposed and analyzed. The design employs graphene as a polarization-converting MTS and TOPAS as the dielectric substrate, enabling tunable polarization conversion and circular polarization. TOPAS is selected due to its low absorption characteristics, making it well-suited for broadband THz components. By varying the chemical potential of graphene from 0 eV to 0.5 eV, the polarization state can be dynamically controlled within the 1 THz (0.76–1.02 THz) and 2.5 THz (2.43–2.6 THz) bands. The multilayer graphene configuration, combined with the low-loss TOPAS substrate, yields wide operational bandwidths of 0.26 THz and 0.17 THz for the 1 THz and 2.5 THz bands, respectively. The antenna's polarization mode – circular or linear – can be selectively switched within the 1 THz (0.7–0.75 THz, 0.96–1.04 THz) and 2.5 THz (2.42–2.52 THz) frequency ranges without requiring any modification to the physical geometry. Furthermore, the individual components, including the graphene antenna, the MTS, and the hybrid structure, were evaluated independently to confirm that their performances remain mutually noninterfering. The hybrid antenna configuration, illustrated in **Figure 14**, demonstrates significant potential for reconfigurable THz devices and related high-frequency applications.

In a report by Qian [71], an RMTS antenna is introduced for versatile control of scattering fields through the use of liquid metal. The proposed design utilizes eutectic gallium-indium (EGaIn), whose melting point is approximately 30°C, enabling facile reconfiguration under moderate temperature adjustments. An element cavity is incorporated to contain the liquid metal, allowing flexible geometric reconfiguration. By melting and mechanically rotating the cavity structure, the shape of the liquid metal can be altered, producing distinct reflective phase responses. By varying the spatial distribution of these liquid metal elements across the MTS, different scattering field patterns – such as single-beam and dual-beam configurations – can be realized. This reconfigurable approach provides a promising pathway for advanced antenna designs in 5G systems and other intelligent wireless applications. The design can be seen in **Figure 15**.

In recent years, IRSs have emerged as a viable technology for dynamically shaping the radio environment through real-time control of signal reflection and refraction. In this thesis [72], the challenges of channel estimation and passive beamforming design in IRS-assisted high-mobility communication scenarios are investigated, considering both vehicle-mounted and roadside IRS deployment strategies. Several practical and efficient methodologies are proposed to address these challenges. The outcomes of this study provide valuable insights and contribute

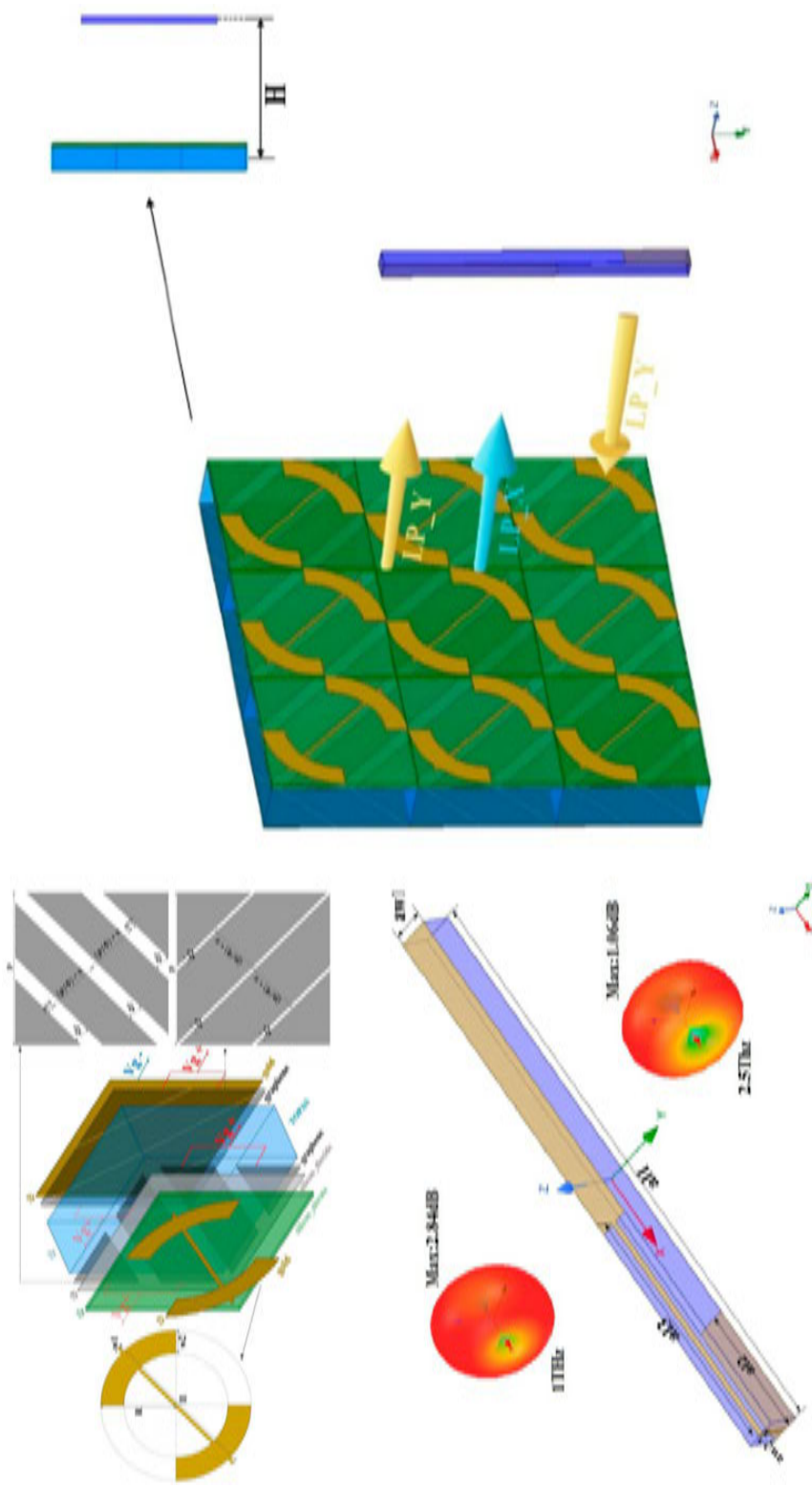


Figure 14. The reconfigurable terahertz antenna with MTS [70].

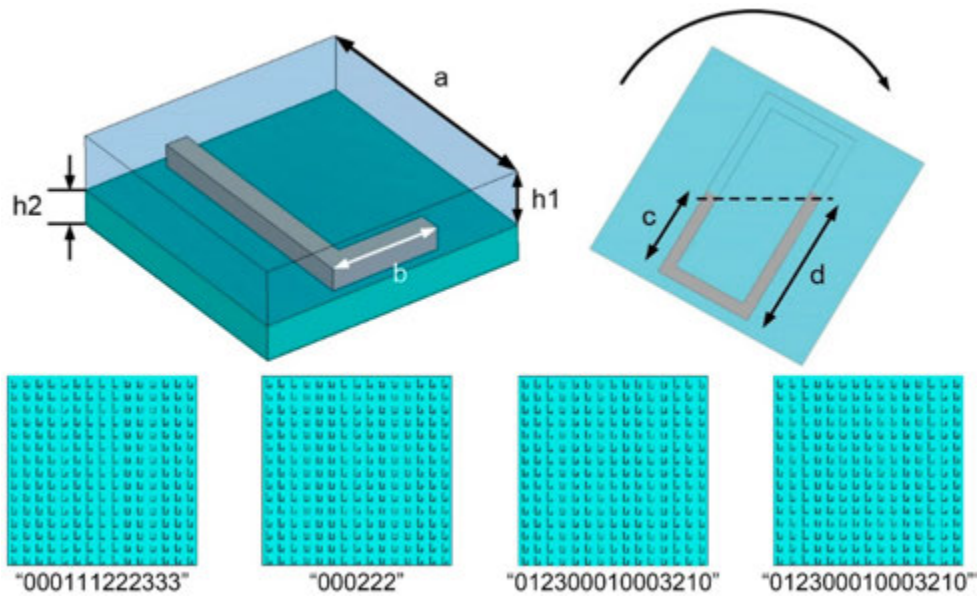


Figure 15.
 The reconfigurable MTS antenna using liquid metal and MTS coding [71].

to the advancement of IRS-based solutions aimed at enhancing communication performance in high-mobility environments.

This study investigates a novel IRS-assisted high-mobility communication system by exploiting the signal refraction capabilities of IRSs in conjunction with moving vehicles. A two-stage transmission protocol is introduced to enable efficient channel estimation, alongside an IRS refraction design aimed at enhancing communication speed and reliability. Initially, the LoS component of the BS-IRS-user cascaded channel is estimated using the quasi-static IRS-user channel, allowing the IRS to adjust its refraction to maximize the passive beamforming gain. Subsequently, the IRS-refraction and IRS-nonrefraction channels are coherently aligned through a common phase shift adjustment of all IRS refraction elements, further boosting the signal strength received by the user for data transmission. Simulation results indicate that the proposed scheme effectively converts the end-to-end BS-user channel from fast-fading to slow-fading, achieving substantial performance gains over conventional slow-fading IRS schemes designed for low-mobility users. Moreover, the vehicle-mounted IRS system demonstrates superior efficiency in enhancing communication performance for users in high-speed vehicles compared to the baseline roadside IRS system, which requires multiple IRS units deployed at fixed intervals along the roadside to sequentially serve passing high-mobility users.

Additionally, a roadside IRS-assisted high-mobility vehicular communication framework is examined, incorporating a low-complexity passive beamforming strategy and an efficient channel estimation method. These approaches do not necessitate modifications to the existing uplink transmission protocol or real-time feedback from the BS to individual IRS units, rendering the system practically feasible for high-speed vehicular communications. Simulation results confirm that the proposed designs significantly improve the user-to-BS channel gain distribution, thereby enhancing both throughput and reliability in high-mobility scenarios.

In the thesis by Koçer [20], designs of MTS microstrip antennas and 4×4 MIMO antennas operating in circular polarization in the 3.3–3.8 GHz frequency range, also known as the n78 band, for the sub-6 GHz 5G frequency spectrum were developed using HFSS software. To improve the performance of the microstrip antennas, a 4×4 MTS was integrated onto the antenna without an air gap. When the 4×4 MTS was placed on the antenna, the TM and TE surface waves generated by the MTS resonated at high frequencies, increasing the bandwidth, gain, and efficiency of the microstrip antenna. It was observed that creating a diagonal slot in the patch of the MTS microstrip antenna resulted in a lower reflection coefficient. The design can be seen in **Figure 16**.

In a report by Althuwayb et al. [73], a novel approach combining substrate-integrated waveguide (SIW) and MTS technologies is proposed to address the bandwidth, gain, and efficiency limitations commonly encountered in on-chip THz antennas. The integration of these techniques reduces substrate-induced losses and suppresses the detrimental effects of surface waves. It is demonstrated that incorporating MTSs with SIW structures significantly enhances impedance bandwidth, radiation gain, and efficiency, all without enlarging the antenna's aperture. The antenna is realized on a gallium arsenide (GaAs) substrate and consists of a 2×4 array of rectangular patches, each featuring a row of circular slots, and is excited via a microstrip coplanar waveguide feedline. Compared to prior works, the proposed design achieves a broader frequency bandwidth and operates at higher frequencies while maintaining comparable gain and radiation efficiency. Specifically, the structure provides a reflection coefficient better than $S_{11} \leq -15$ dB and achieves a radiation gain and impedance matching of approximately 30 dB over the 0.45–0.5 THz range – a 15 dB improvement relative to configurations without slots and VIAs. The antenna exhibits a fractional bandwidth of 10.52%. The inclusion of slots in the patch elements further enhances performance: the average radiation gain increases to 6.5 dBi, representing a 2.7 dBi improvement compared to designs without slots, while the corresponding radiation efficiency rises to approximately 65%, a 12% improvement. Incorporating metallic VIAs yields similar benefits, with the average gain reaching 6.5 dBi (a 4.4 dBi enhancement) and the radiation efficiency improving by 19% to approximately 65%. **Figure 16** illustrates the proposed antenna configuration and its key structural features.

In the review presented by Alibakhshikenari [74], it is demonstrated that mutual coupling reduction techniques inspired by MTM and MTS concepts can achieve enhanced isolation between adjacent radiating elements through simple and cost-effective decoupling configurations, while exerting minimal influence on array characteristics such as bandwidth, gain, radiation efficiency, and overall footprint. Various antenna configurations – including reconfigurable, etched, dielectric resonator, MTS, and MTM-based designs – have been widely employed to mitigate interference effects. Additionally, diverse methodologies for suppressing mutual coupling in antenna arrays intended for MIMO and synthetic aperture radar (SAR) applications have been comparatively examined, emphasizing the unique properties of MTM and MTS structures. As an illustrative example, designs aimed at enhancing isolation in MTS-wall-based antenna arrays operating in the terahertz frequency band are depicted in **Figure 17** [75, 76].

In a report by Alibakhshikenari et al. [75], a novel MTS wall was introduced to mitigate interference between antennas in arrays operating within the 139–141 GHz band, suitable for applications in security screening, medical imaging, and

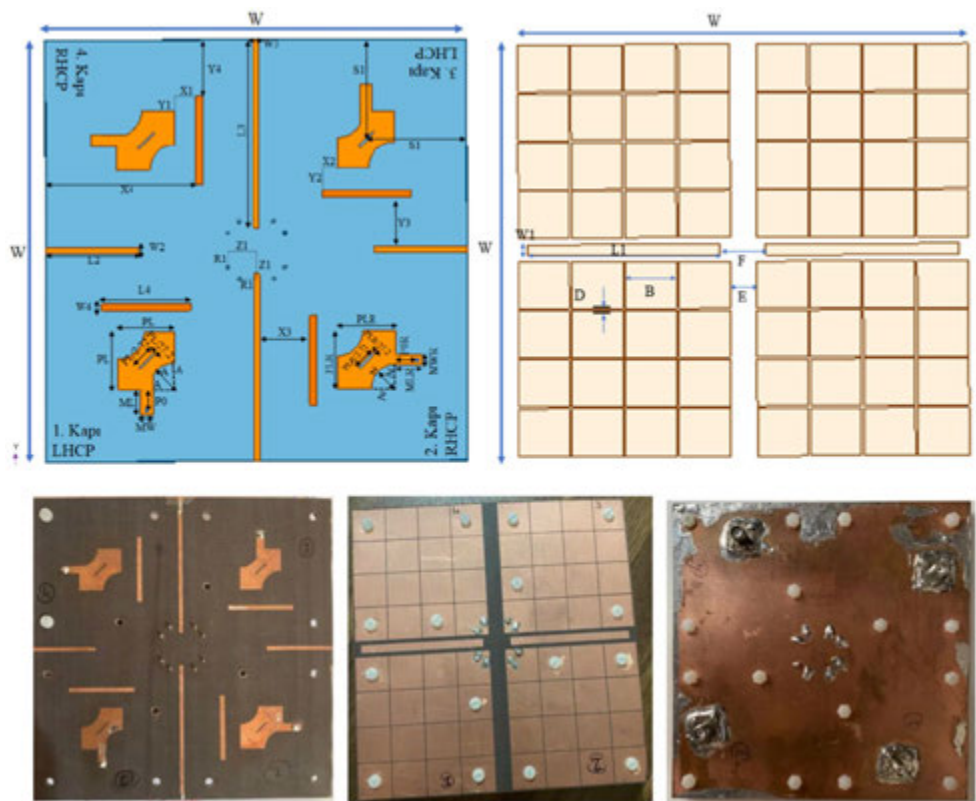


Figure 16.
The proposed 4×4 MIMO antenna design with fabricated structures [20].

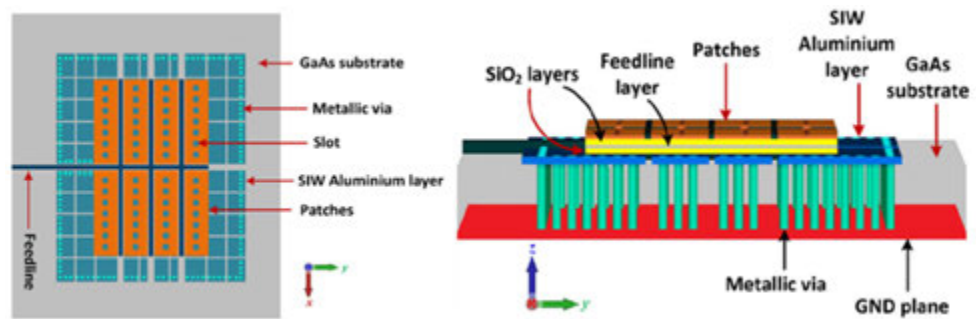
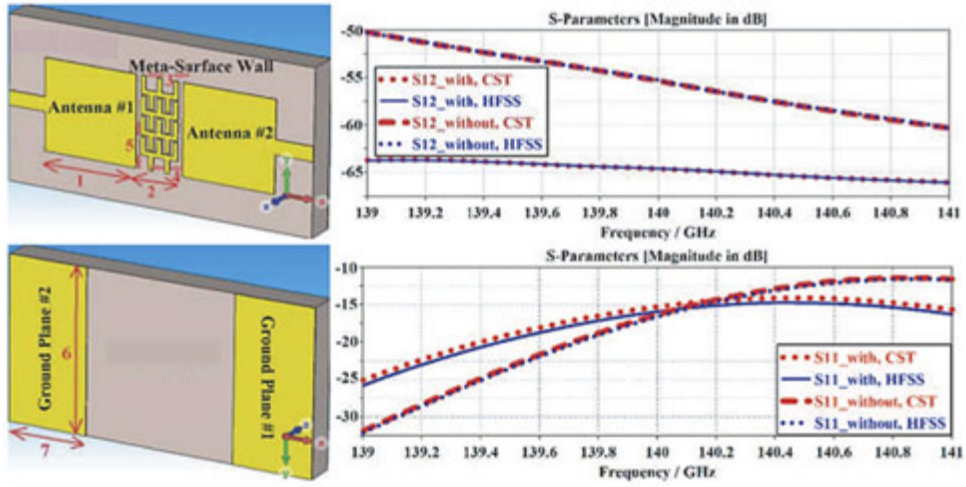


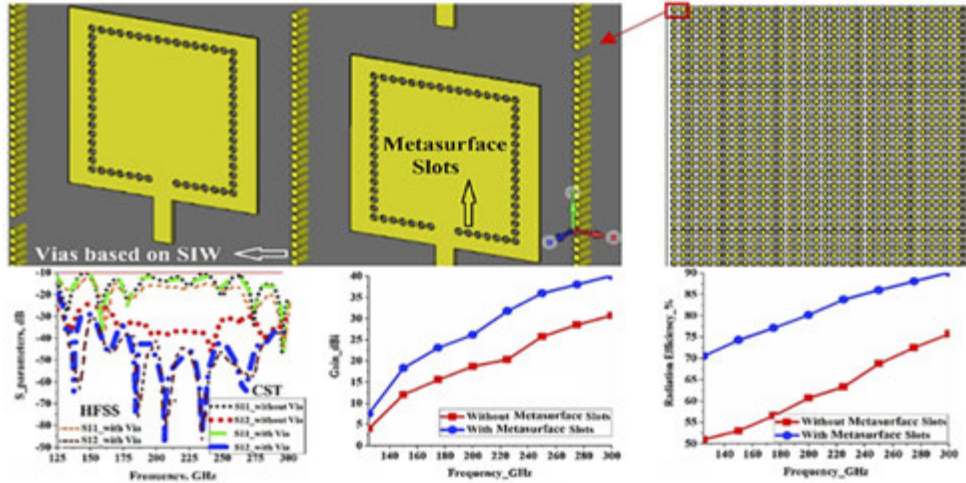
Figure 17.
Design with MTS + SIW technique recommended for THz chip antenna [73].

communication systems. The MTS unit cell consists of interconnected twin “Y-shaped” microstrip structures that are digitally coupled to form the complete MTS wall. The design eliminates through-holes and incorporates a shortened ground plane, simplifying the fabrication process.

As illustrated in **Figure 18a**, the MTS wall is densely positioned between antenna elements to enhance isolation and suppress surface waves. The wall is oriented perpendicular to the antennas to minimize mutual coupling. Across the THz band, the array antennas exhibit a gain below 9.0 dBi and isolation exceeding 63 dB. This approach achieves an isolation improvement of more than 10 dB over a 2 GHz bandwidth, surpassing previously reported results. When the MTS wall is implemented between elements, it effectively prevents electromagnetic signals from propagating from one element to adjacent elements. Utilizing this configuration allows the edge-to-edge separation between radiating patches to be reduced to 2.5 mm. The corresponding dimensions of the antenna and ground plane are $5 \times 5 \text{ mm}^2$ and $9 \times 4.25 \text{ mm}^2$, respectively, when fabricated on a conventional 1.6 mm-thick substrate.



(a)



(b)

Figure 18.

(A) The proposed two-dimensional MTS wall for suppressing interference [63]. (B) A 34x34 array with MTS and SIW for improving elements' isolation [76].

In a report by Alibakhshikenari et al. [76], a perceptual model of a 34-element antenna array operating within the 0.125–0.3 THz range, corresponding to an effective bandwidth of 82.35%, is presented. As depicted in **Figure 18b**, each radiating element consists of a square patch with physical dimensions of $2 \times 2 \text{ mm}^2$, excited via a matched microstrip line. Adjacent elements are separated using through-holes fabricated according to the SIW technique, which serves to enhance isolation and prevent degradation of radiation performance caused by closely spaced elements.

The perimeter of each patch incorporates circular dielectric slots based on the MTS concept to further improve radiation characteristics. Through the implementation of these design strategies, an average isolation improvement of 25 dB is achieved across the operational bandwidth, while the effective aperture of the array is increased without modifying its overall dimensions. The array demonstrates a gain ranging from 7.51 dBi to 40.08 dBi and radiation efficiency varying between 70.51% and 90.11%. Following the introduction of MTS slots, the results indicate gain and efficiency enhancements of approximately 60% and 30%, respectively.

In a report by Alibakhshikenari et al. [77], a feasibility analysis is conducted on the performance characteristics of a novel on-chip antenna employing MTS technology in the THz frequency range. The proposed MTS-based on-chip antenna is fabricated on a thin, high-permeability GaAs substrate layer. Measurements indicate that the antenna achieves an average gain exceeding 10 dBi and a radiation efficiency greater than 60% across a broad frequency range of 0.41–0.47 THz. The compact antenna structure, with dimensions of $8.6 \times 8.6 \times 0.0503 \text{ mm}^3$, demonstrates substantial performance improvements over previously reported on-chip antenna designs. Specifically, the MTS configuration enhances both gain and radiation efficiency while maintaining a minimal footprint. These results validate the suitability of the GaAs-based MTS on-chip antenna for integration into THz circuit applications. **Figure 19** illustrates the structure along with simulation and measurement outcomes.

In recent years, the development of MTM and MTS-based antennas has attracted considerable attention due to their exceptional ability to control the propagation, reflection, and radiation properties of EM waves. The integration of MTSs into antenna structures improves impedance matching, gain, radiation efficiency, and bandwidth, provides reconfigurability, and enables compact and low-profile designs suitable for modern communication systems [73, 75].

Various studies in the microwave, millimeter wave, and terahertz bands demonstrate that MTSs offer high gain, low isolation, and enhanced directivity in broadband [73, 75–78]. These developments, combined with recent RIS and hybrid MTS applications, have transformed MTSs from passive artificial materials into dynamic platforms capable of beam steering, sensing, and adaptive control functions [59, 79–82].

For the purpose of a collective comparison, the study examples selected from the literature and provided in the sections above are summarized in **Table 1**.

6. Conclusion and recommendations

Future 5G wireless networks are expected to leverage the THz band as a fundamental enabler for accommodating the rapidly increasing wireless traffic,

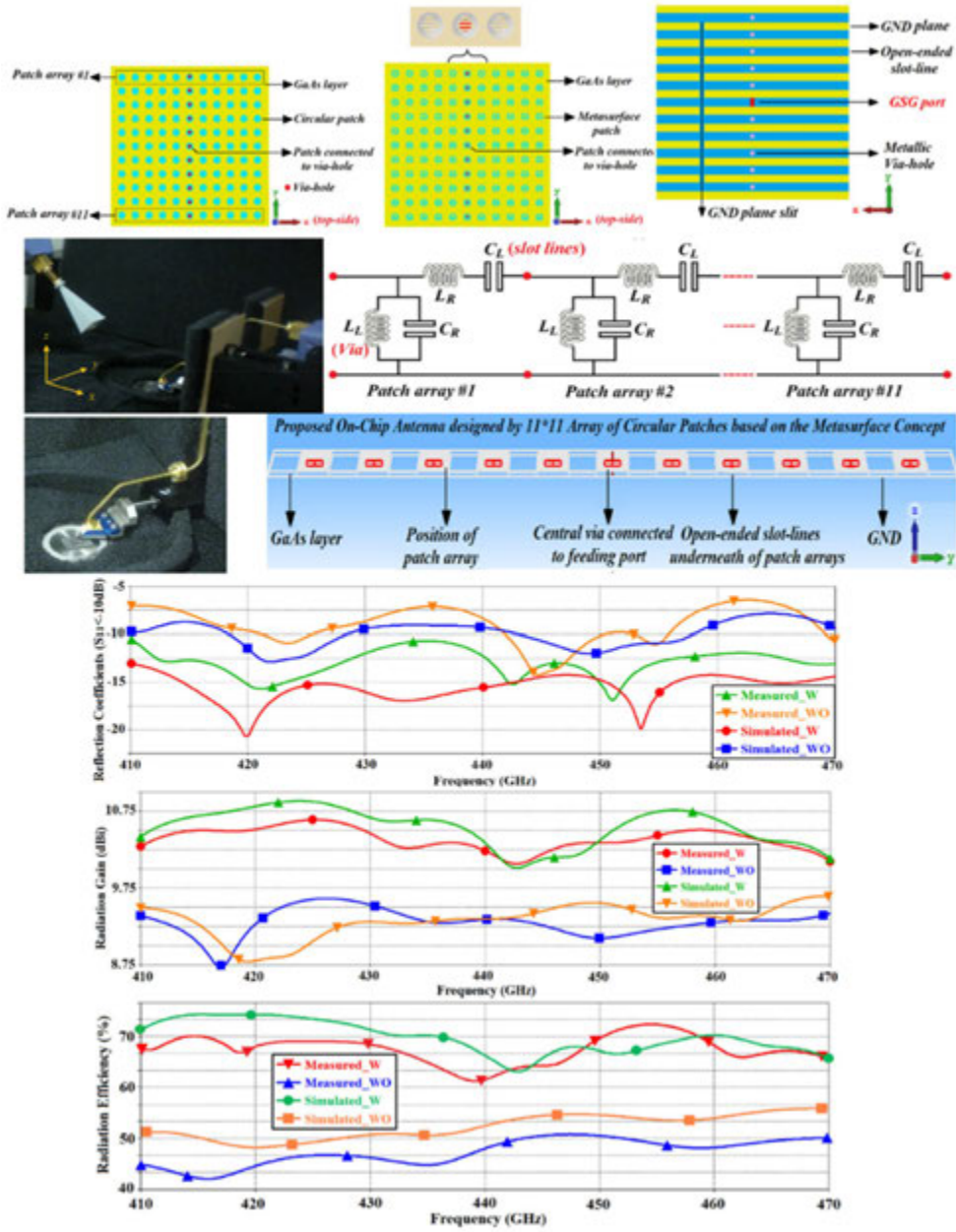


Figure 19. The proposed MTS structure on THz chip antenna with some of the results from simulations and measurements [77].

thereby supporting ultrahigh data transmission rates. Nevertheless, the unique propagation characteristics of the THz spectrum result in substantial signal attenuation, leading to considerable path losses in communication systems. RISs have emerged as a promising solution to mitigate these limitations by dynamically controlling wave propagation, directing signals toward desired targets. When integrated with UM-MIMO antenna systems, RIS technology can significantly

enhance network performance. In such configurations, the ability of multiple antennas to generate highly directive narrow beams is crucial for efficient signal propagation and the successful deployment of THz communications.

High-performance communication systems can be achieved by addressing the fundamental challenges of mmWave antenna design, such as high signal loss, poor penetration, difficult miniaturization, high heat generation at high frequencies, and material limitations. MTS-supported MIMO antenna designs, hybrid antennas with RIS, and IRS structures are used where necessary. mmWave antennas are utilized in diverse sectors beyond telecommunications, such as healthcare, defense, and the automotive industry. For 5G and 6G networks, mmWave technology enables high-speed wireless communication, offering faster data transfer, lower latency, and better efficiency.

MTSs are attracting significant attention due to their ability to perform hexagonal wavefront modulation. Composed of meticulously designed subwavelength elements, 2D MTSs rewrite large wavelengths with manipulative elements. From shaping light beams to producing ultrathin lenses and enabling future communication devices, MTSs fascinate both researchers and engineers with their potential to transform industries. In addition to massive wafers, MTSs must accommodate a wide variety of surface options, meet a wide range of practical requirements, and be compatible with a variety of devices. Consequently, creating a soft MTS that can bend or stretch is one of the hottest new research areas. For a flexible MTS structure, materials such as carbon nanofilms, metal-dielectric combined structures, or a geometry consisting of only the dielectric are deposited on substrates such as flexible polymers.

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Chapter 3

Metasurfaces for Applications of Multi-Dimensional Orbital Angular Momentum Mode (De)multiplexing

Qingji Zeng and Shuqing Chen

Abstract

Orbital angular momentum (OAM) mode, as an emerging physical dimension for optical (de)multiplexing, offers unprecedented potential to enhance communication capacity density by synergistically operating with conventional domains such as polarization and wavelength. Metasurfaces, with their subwavelength configurability and multi-parameter manipulation capabilities, serve as powerful platforms for engineering multi-dimensional OAM modes. They bring revolutionary advantages – including increased parallel processing dimensions, improved light-field modulation efficiency, and system miniaturization – beyond those achievable with conventional optical components in multi-dimensional OAM mode treatment. This chapter dissects metasurface-enabled strategies for hybrid multiplexing/demultiplexing in three multi-physical domain scenarios: OAM–polarization parallel processing using the expanded Pancharatnam–Berry phase for orthogonal linearly polarized Dammann-vortex interactions; OAM–wavelength parallel manipulation utilizing a dispersion-engineered multi-foci transformation strategy implemented in broadband meta-atoms; and tri-domain (OAM–polarization–wavelength) parallel (de)multiplexing realized through spin-decoupled cascaded phase modulation on a five-layer diffractive meta-platform. For each scenario, the fundamental principles, design methodologies, optical demonstrations, and performance trade-offs are discussed. The chapter highlights the pivotal role that metasurfaces are poised to play in shaping the development of next-generation high-speed, large-capacity all-optical communication networks.

Keywords: optical communication, metasurface, OAM mode, multi-dimensional (de)multiplexing, structured light manipulation

1. Introduction

Amid the big data epoch, global information volume is experiencing an unprecedented surge – driven by the deepening digital transformation across industries, the large-scale deployment of the Internet of Things, and the widespread

adoption of high-throughput data transmission technologies [1–3]. This exponential growth imposes formidable challenges on the performance and scalability of modern communication architectures, which are increasingly strained to keep pace with escalating data demands. Traditional single-channel communication systems are inherently constrained by three critical limitations: low throughput efficiency, suboptimal utilization of spectral bandwidth, and limited scalability of hardware equipment [4, 5]. Compounding this issue, annual global data generation has now approached the zettabyte scale – a threshold that far exceeds the processing and transmission capabilities of existing data infrastructure, rendering conventional systems inadequate for handling such massive data loads [6, 7].

Against this backdrop, optical multiplexing technology has emerged as a pivotal solution to bridge the looming gap in future data transmission capacity [8–10]. Its core advantage lies in enabling parallel data transmission across multiple independent physical dimensions through shared optical media, thereby drastically enhancing overall communication efficiency without requiring a proportional expansion of hardware. To date, researchers have made significant strides in advancing key optical multiplexing technologies, including wavelength division multiplexing (WDM), polarization division multiplexing (PDM), time division multiplexing, and code division multiplexing [11–18]. These technologies operate by encoding photonic signals onto distinct physical dimensions – specifically, the spectral domain, polarization states, time slots, and coding sequences, respectively – with the dual goals of maximizing the utilization of limited communication resources and safeguarding the integrity of transmitted data. Nevertheless, a critical bottleneck has emerged: The development of these conventional physical dimension resources is rapidly approaching fundamental limits imposed by the laws of physics. For instance, spectral bandwidth in WDM is constrained by the transparency window of optical fibers, while polarization states in PDM are limited by the intrinsic symmetry of electromagnetic waves [19, 20]. This physical ceiling means that further capacity gains from existing dimensions will be incremental at best, failing to meet the exponential growth of data traffic. Thus, breaking through the current constraints on channel capacity to substantially boost transmission capability has become an urgent and core task for the advancement of next-generation optical communication systems and networks. Exploration of novel multiplexing dimensions stands as a preeminent pathway for transformative breakthroughs.

Vortex beams, defined by their helical phase fronts that inherently carry orbital angular momentum (OAM) modes, introduce a novel physical dimension for channel multiplexing – rooted in the strict spatial orthogonality of distinct OAM modes, which prevents mutual interference between co-propagating signals [21–24]. Theoretically, OAM modes exhibit two key advantages for high-capacity communication: infinite scalability (in principle, an unbounded number of orthogonal modes can be generated) and intrinsically low inter-mode crosstalk – attributes that make them ideal carriers for mode division multiplexing (MDM) and demultiplexing in next-generation optical communication systems [25–27]. Leveraging this uniqueness, multiple independent data streams can be parallelly multiplexed within the same frequency band and single optical channel, directly translating to a proportional enhancement in signal transmission density without expanding spectral bandwidth [28, 29].

To realize practical OAM-based MDM, a suite of mode (de)multiplexing technologies has been developed and refined, including series-connected beam

splitter assemblies, multi-order diffractive gratings, optical coordinate transformation modules, angular dispersion lenses, and multi-layer cascaded modulation schemes [30–34]. These technologies have enabled precise generation, manipulation, and separation of OAM modes, driving significant advancements in OAM-MDM research. However, as global data traffic continues to grow exponentially, standalone OAM-MDM can no longer meet the escalating capacity demands; instead, integrating OAM modes with conventional multiplexing dimensions – specifically WDM and PDM – to form multi-dimensional hybrid multiplexing systems has become imperative to unlock the full capacity potential of optical networks [35–40]. This integration, in turn, necessitates the development of unified (de)multiplexing technologies capable of jointly processing OAM, wavelength, and polarization dimensions – an innovation widely recognized as the key to breaking current optical communication capacity bottlenecks and advancing multiplexing technology to its next evolutionary stage.

On the other hand, metasurfaces – engineered planar structures composed of periodically arranged subwavelength pixelated nanostructures – exhibit a unique capability to dynamically manipulate multiple fundamental physical properties of incident electromagnetic waves with high precision, including amplitude, phase, polarization states, and operating frequency [41–48]. This exceptional multi-dimensional control is rooted in the nanostructures' ability to tailor light-matter interactions at the subwavelength scale, allowing for simultaneous and independent modulation of distinct optical dimensions (e.g., OAM modes, wavelength bands, and polarization states) in a compact, integrated platform [49–54]. Unlike conventional bulk optical components that often require complex cascading to achieve multi-dimensional control (leading to increased system size, insertion loss, and alignment complexity), metasurfaces enable on-chip integration of multi-functional optical manipulation, making them a powerful and versatile platform for advancing multi-dimensional multiplexing in next-generation optical communication systems.

Accordingly, this chapter presents several metasurface-based technologies for multi-dimensional OAM mode (de)multiplexing. Specifically, Sections 2–4 are dedicated to three principal hybrid (de)multiplexing architectures: OAM–polarization parallel (de)multiplexing, leveraging an expanded Pancharatnam–Berry (PB) phase to manipulate orthogonal linearly polarized Dammann vortices; OAM–wavelength parallel (de)multiplexing via a dispersion-engineered multi-foci transformation strategy realized using broadband meta-atoms; and OAM–polarization–wavelength parallel (de)multiplexing achieved through spin-decoupled phase modulation in a five-layer diffractive meta-platform. For each configuration, the latest research progress is reviewed, competing design methodologies are compared, and critical performance trade-offs are highlighted. Insights into their current capabilities and limitations are included. Ultimately, the discussion underscores the transformative potential of metasurfaces in advancing next-generation, high-capacity, and high-speed all-optical communication infrastructures.

2. OAM–polarization parallel (de)multiplexing

Vortex beams carrying OAM modes have emerged as a key enabler for ultrahigh-speed optical communications, owing to their intrinsic orthogonality among

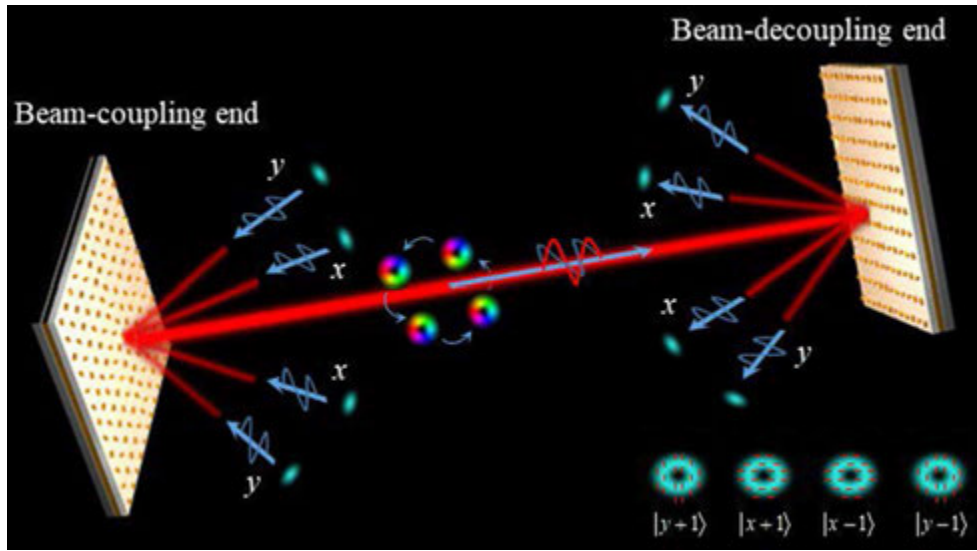


Figure 1. Schematic of parallel (de)multiplexing with OAM mode and polarization channels [36].

different modes, which allows multiplexing independent of the traditional polarization dimension. Integrating OAM with PDM and WDM has already pushed data transmission capacities to the petabit-per-second regime [55]; however, practical deployment remains constrained by the absence of compact (de) multiplexers capable of simultaneously handling OAM and polarization channels. Existing solutions suffer from fundamental limitations: spiral phase plates and Q-plates only manipulate single OAM modes, spatial light modulators (SLMs) lack the ability for polarization-selective phase control, and available metasurfaces often exhibit wavelength sensitivity or inadequate polarization operation – all of which lead to severe energy loss, increased system complexity, and barriers to all-optical integration [56, 57]. Addressing this gap is critical for unlocking the full potential of OAM–polarization multi-dimensional multiplexing.

To overcome this challenge, a metasurface-based approach for the parallel (de) multiplexing of OAM and polarization channels was presented, leveraging an expanded PB phase concept to independently manipulate the phases of two orthogonal linearly polarized light components [36]. The core (de)multiplexing mechanism is illustrated in **Figure 1**: At the beam-coupling end (realized by the left metasurface), four-channel orthogonal linearly polarized Gaussian beams are incident on the metasurface at distinct diffraction orders. Through an expanded PB phase mechanism implemented for a pair of binarized Dammann vortex gratings with different periodicities encoded on the metasurface, these Gaussian beams are converted into vortex beams carrying tailored OAM modes (topological charges $l = \pm 1$). This process generates a set of orthogonal linear polarization–OAM mode combinations, denoted as $|y + 1\rangle$, $|x + 1\rangle$, $|x - 1\rangle$, $|y - 1\rangle$, as shown in the lower right of **Figure 1**. Subsequently, the diffracted beams are coupled at the zeroth diffraction order and combined into a coaxial propagation path, thereby achieving parallel multiplexing across both OAM mode and polarization channels. At the beam-decoupling end after propagation (realized by the right metasurface), an

identical metasurface reverses this process based on optical path reversibility. The coaxial vortex beams are converted back into spatially separated Gaussian beams with specific linear polarizations (either x - or y -polarization), thus enabling parallel demultiplexing of the OAM mode and polarization channels. Overall, this scheme facilitates the transformation between orthogonal linear polarization channels and conjugate OAM modes, enabling parallel (de)multiplexing of polarization-encoded OAM modes – that is, the coupling and separation of multiple optical channels.

The implemented metasurface is constructed in a metal–insulator–metal configuration to attain high reflectivity and precise phase manipulation. It consists of rectangular gold (Au) meta-atoms (length $L = 380$ nm, width $W = 220$ nm, height $H = 50$ nm) positioned above a continuous 200-nm-thick Au reflector, separated by a 150-nm-thick SiO₂ dielectric spacer. The entire structure is fabricated on a silicon substrate with an 800-nm lattice constant to suppress high-order diffraction. Two binarized Dammann-vortex gratings (phase values: 0 and π) with orthogonal polarization responses are inscribed on the metasurface, each with distinct periodicities ($T_x \neq T_y$) to ensure spatial separation of polarization outputs. Specifically, the Dammann-vortex grating phases of paired linearly polarized polarizations are expressed as follows [36]:

$$\phi_{x,y} = \arg \left\{ \sum_{m=-\frac{N}{2}}^{\frac{N}{2}} C_m \exp \left[im \left(\frac{2\pi s}{T_{x,y}} + \Delta l \varphi \right) \right] \right\} \quad (1)$$

Here, N corresponds to the total number of diffraction orders, with m representing the diffraction order index spanning from $-N/2$ to $+N/2$. C_m (defined as $1/N$) denotes the normalized power distribution associated with the m -th diffraction order. The parameter s characterizes the phase transition location within the normalized grating period. $T_{x,y}$ signifies the fundamental period of the Dammann grating structure. Δl indicates the separation interval between consecutive topological charges, while φ stands for the azimuthal angular coordinate. Thereafter, by exploiting the expanded PB phase concept, the metasurface imparts binary phase modulation to x - and y -polarized light through azimuthal rotation (θ) of the meta-atoms. The geometric nature of the PB phase ensures minimal dispersion, yielding a broad operational bandwidth from 1535 to 1607 nm, which aligns well with WDM requirements.

Figure 2 illustrates the optical performance of the designed metasurface-enabled (de)multiplexer, which performs parallel processing of OAM mode and polarization channels. The device demonstrates robust performance under experimental validation. When illuminated by an x -polarized Gaussian beam at the zeroth diffraction order, four distinct polarization-dependent OAM modes ($|y+1\rangle$, $|x+1\rangle$, $|x-1\rangle$, $|y-1\rangle$) were generated at separate diffraction orders, exhibiting uniform intensity profiles and a total diffraction efficiency of $\sim 85\%$. The topological charges of the generated OAM modes were confirmed via cylindrical lens (C-lens) interference patterns, where the number of dark fringes corresponds to the modulus of the topological charge l , and their tilt direction indicates its sign (see the left panel of **Figure 2**). Notably, the device exhibits minimal crosstalk – approximately two orders of magnitude lower than the signal intensity – attributed to the strong orthogonality among OAM modes and polarization states (refer to the right panel of **Figure 2**).

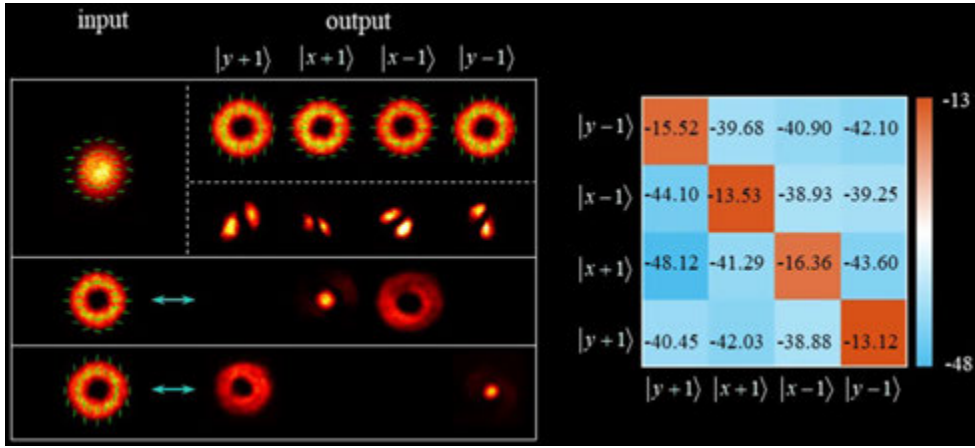


Figure 2. Optical characterization of the metasurface-based (de)multiplexer [36]. The left panel presents the experimentally measured output intensity profiles for the four-channel OAM–polarization (de)multiplexing operation, and the corresponding confusion matrix on the right quantitatively evaluates the channel isolation and inter-channel crosstalk.

The preceding section establishes a scalable and compact platform for multi-dimensional optical communication, paving the way for integrated, high-capacity data links that fully exploit the degrees of freedom offered by OAM mode and polarization. As analyzed, the key to achieving parallel OAM–polarization (de)multiplexing lies in enabling a polarization-decoupled phase modulation mechanism. Such a mechanism must facilitate both OAM mode conversion and beam gradient coupling/separation for multiple orthogonal polarization channels, thereby forming multi-dimensional links based on polarization-orthogonal OAM mode transmission. Several phase modulation designs in a polarization-decoupled configuration – such as those employing Dammann-vortex or grating-gradient interactions [58–60] – have been reported in the literature, yet their underlying principles align with the framework described above. It should be noted, however, that while grating-based multi-order diffraction enables multi-dimensional OAM–polarization (de)multiplexing, it does so at the expense of sorting efficiency, which is fundamentally limited to $1/N$. This leads to considerable energy loss, particularly as the number of channels increases. As a result, the practically achievable channel count is inversely constrained by power efficiency, presenting a critical trade-off that must be carefully balanced in real-world implementations.

3. OAM–wavelength parallel (de)multiplexing

The inherent independence between optical OAM and frequency makes mode and wavelength parallel demultiplexing feasible. When combined with WDM, OAM mode multiplexing systems can surpass the spectral efficiency of available bandwidth. Although methods such as composite vortex gratings and multi-layer diffraction modulation have made significant progress in multi-dimensional (de)multiplexing, on-demand parallel demultiplexing of OAM–wavelength channels

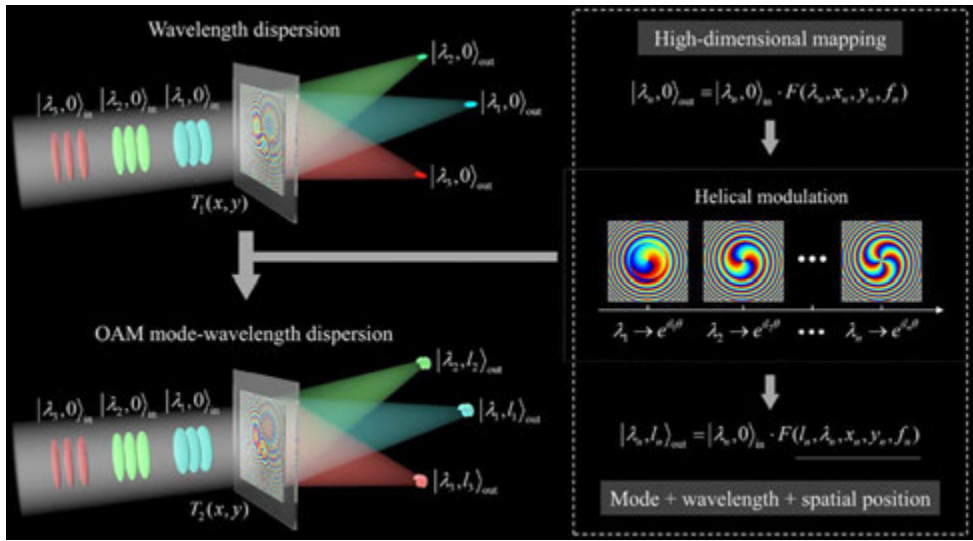


Figure 3.
 Mechanism of OAM–wavelength dispersion management [38]. The diagram depicts a coordinated mapping from OAM modes and wavelengths to distinct spatial foci, illustrating the underlying principle for the demultiplexing function.

remains challenging. Due to the Bragg diffraction effect of gratings, multi-order dispersion of modulated modes occurs, leading to low mode conversion efficiency [61]. Furthermore, while such diffraction facilitates wavelength separation, the usable angular deflection remains restricted to gradient-defined directions, thereby confining wavelength routing to a narrow spatial range and limiting channel allocation primarily by axial position [62]. Additionally, conventional OAM mode modulation devices, such as diffractive optical elements, SLMs, and digital micromirror devices, are typically limited to specific operating frequencies, rendering them incapable of spectrally adaptive control. As a result, the demultiplexing of jointly encoded OAM and wavelength channels remains an open challenge, constrained by inefficiencies in mode transformation and the lack of versatile spectral management.

To surmount this problem, a strategy introducing a wavelength-dependent multi-foci transformation and utilizing an off-axis dispersion-managed metasurface to parallelly control the OAM mode and wavelength dimensions was proposed [38]. As shown in **Figure 3**, by incorporating tunable wavelength and focal length variables into the composite lens phase, a linear mapping of multiple focal planes was spatially established, enabling efficient dispersion separation of multiple wavelength channels. Simultaneously, spiral modulation was introduced into the lens phase for mode control, where applying a conjugate spiral phase enabled the inverse conversion of target input OAM modes into fundamental Gaussian modes. To be specific, the dispersion-engineered multi-foci transformation phase profile is formulated as follows [38]:

$$\phi(x, y) = \arg \left\{ \sum_{n=1}^N \sum_{m=1}^M \exp[i\varphi_{n,m}(x, y)] \right\} \quad (2)$$

$$\varphi_{n,m}(x,y) = -\frac{2\pi}{\lambda_n} \left(\sqrt{(x-x_{n,m})^2 + (y-y_{n,m})^2 + f_n^2} - \sqrt{x_{n,m}^2 + y_{n,m}^2 + f_n^2} \right) + l_{n,m}\vartheta \quad (3)$$

Here, l denotes the topological charge of the OAM mode, and $\vartheta = \arctan(y/x)$ represents the azimuthal angle in the cylindrical coordinate system. As indicated in Eq. (3), each wavelength component λ_n at the matched spatial coordinates $(x_{n,m}, y_{n,m}, f_n)$ can be simultaneously imparted with multiple customizable helical phase profiles $\exp(il_{n,m}\vartheta)$. This enables multi-foci mode conversion, generating distinct phase-carrying foci at designated transverse positions $(x_{n,m}, y_{n,m})$. Notably, since the mode-wavelength dispersion relies on focused Fourier transform rather than periodic gradient modulation, this approach not only avoids multi-order mode conversion without excessive diffraction but also facilitates large-angle wavelength separation control, making OAM–wavelength parallel (de)multiplexing possible.

For physical demonstration, a dielectric transmission-type metasurface was designed, consisting of rectangular titanium dioxide (TiO₂) meta-atoms (length $L = 105$ nm, width $W = 300$ nm, height $H = 600$ nm) arranged periodically on a SiO₂ substrate with a 325-nm lattice constant. TiO₂ was chosen owing to its high refractive index, low optical loss, and broadband transparency across a 360-nm window in the visible spectrum. Phase modulation is achieved via the PB phase mechanism, which leverages in-plane rotation of the meta-atoms to provide wavelength-insensitive wavefront control. The resulting TiO₂–SiO₂ metasurface supports 20 parallel channels for OAM and wavelength multiplexing, incorporating five distinct OAM modes (topological charges $l = -2, -1, 0, +1, +2$) and four wavelengths (490 nm, 530 nm, 570 nm, 610 nm). The device exhibits a broadband response and features a compact footprint of $325 \times 325 \mu\text{m}^2$, highlighting its integration potential. In operation, the OAM–wavelength demultiplexer disperses different wavelength components to four distinct focal planes located at 0.45 mm, 0.55 mm, 0.65 mm, and 0.75 mm, while simultaneously converting each OAM mode into a fundamental Gaussian beam directed to predefined spatial positions. As illustrated in **Figure 4**, the output field distributions are examined under three illumination conditions: (a) a multi-wavelength Gaussian beam, showing wavelength separation at respective focal planes; (b) OAM-carrying beams transformed into Gaussian profiles via multi-focal conversion; and (c) 8-channel hybrid OAM–wavelength signals simultaneously demultiplexed in both spatial and spectral domains. These results confirm the device’s capability for parallel OAM and wavelength (de) multiplexing.

Owing to the absence of strictly orthogonal and independent wavelength control in the dispersion management mechanism, channel crosstalk inevitably arises during the demultiplexing of wavelength-dependent OAM modes. Therefore, an evaluation of wavelength and mode crosstalk in the 20-channel OAM–wavelength demultiplexing was conducted, calculating average wavelength crosstalk of -21.2 dB, -20.0 dB, -19.4 dB, -21.1 dB, and -22.0 dB, and average mode crosstalk of -15.2 dB, -18.1 dB, -20.3 dB, and -21.7 dB. These results confirm that the metasurface maintains low inter-channel interference in demultiplexing hybrid OAM–wavelength channels, reaching a level acceptable for practical implementations. The demultiplexing process follows an inverse operational principle, providing a generalized strategy for multi-dimensional OAM manipulation. By leveraging dispersion-sensitive properties, the metasurface assigns a dispersion-enhanced spiral-phase modulation function to each wavelength,

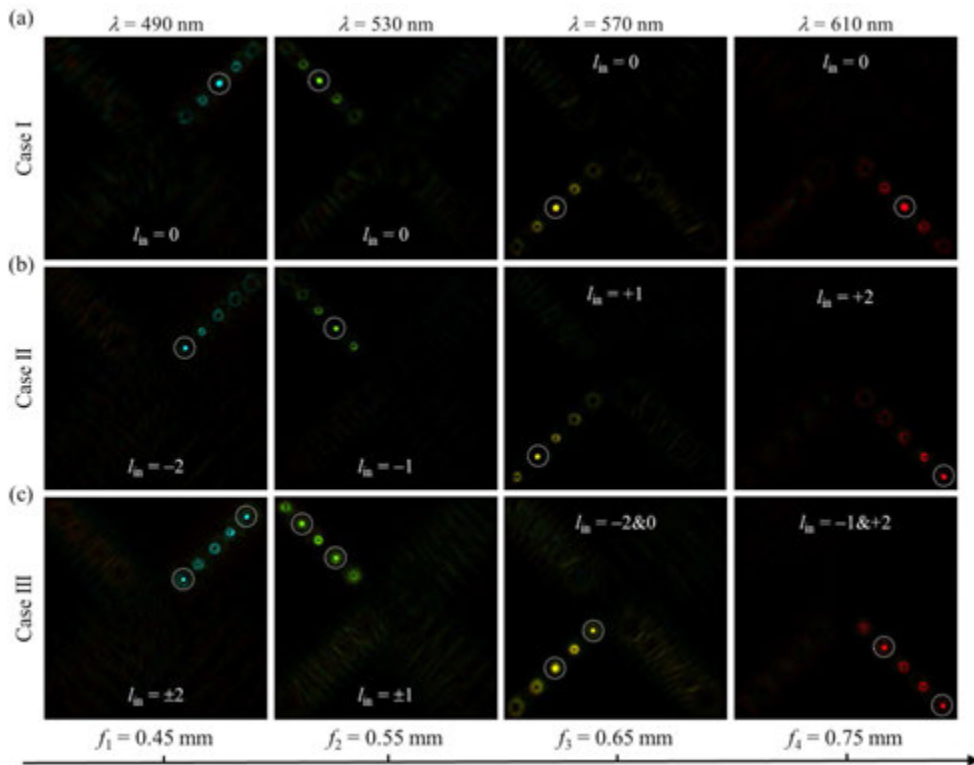


Figure 4.
 Light field characterization of OAM mode-wavelength parallel demultiplexing [38].

enabling differentiated OAM mode conversion and spatially resolved diffraction control. This approach has been effectively validated in both multiplexing and demultiplexing operations, establishing the spiral-phase methodology as a leading solution for parallel wavelength-mode control, as also documented in existing studies [61, 63]. Despite enabling independent and parallel processing of OAM and wavelength channels, the overall optical efficiency remains constrained by the uniform spiral-phase modulation applied across all channels, which fundamentally limits the power efficiency to $\sim 1/N$. Consequently, this strategy is best suited for OAM-based multi-dimensional communication systems with a moderate number of channels.

4. OAM–polarization-wavelength parallel (de)multiplexing

While parallel (de)multiplexing of OAM with polarization or wavelength has been demonstrated, the integration of OAM, wavelength, and polarization dimensions within a single device for full multi-dimensional parallel operation remains an outstanding challenge. Success in this endeavor requires independent control over mode conversion across both wavelength and polarization, coupled with efficient channel coupling and separation.

To address this need, Chen et al. proposed a spin-decoupled multi-layer phase modulation strategy for multi-dimensional hybrid (de)multiplexing, implemented via five-layer diffractive metasurfaces for multi-plane light conversion (MPLC) [40]. As illustrated in **Figure 5**, the system accepts input Gaussian beams at distinct wavelengths (λ_1, λ_2) with identical topological charge l_0 and deterministically transforms them into target OAM states via cascaded diffraction, establishing a bijective mode mapping that enables reversible (de)multiplexing. Crucially, orthogonal spin eigenstates are incorporated as an independent control dimension, enabling decoupled manipulation of OAM and wavelength channels. Wavelength-selective OAM transformation is achieved through a tailored MPLC scheme that exploits the inherent wavelength dependence of Fresnel diffraction. By alternating between free-space propagation and metasurface-imposed phase modulation at successive planes, distinct wavelengths undergo prescribed mode conversions according to precomputed phase profiles. These profiles are inversely designed based on target input–output field mappings, as depicted in **Figure 5**. For polarization diversity, the metasurfaces combine both propagation and geometric phases within

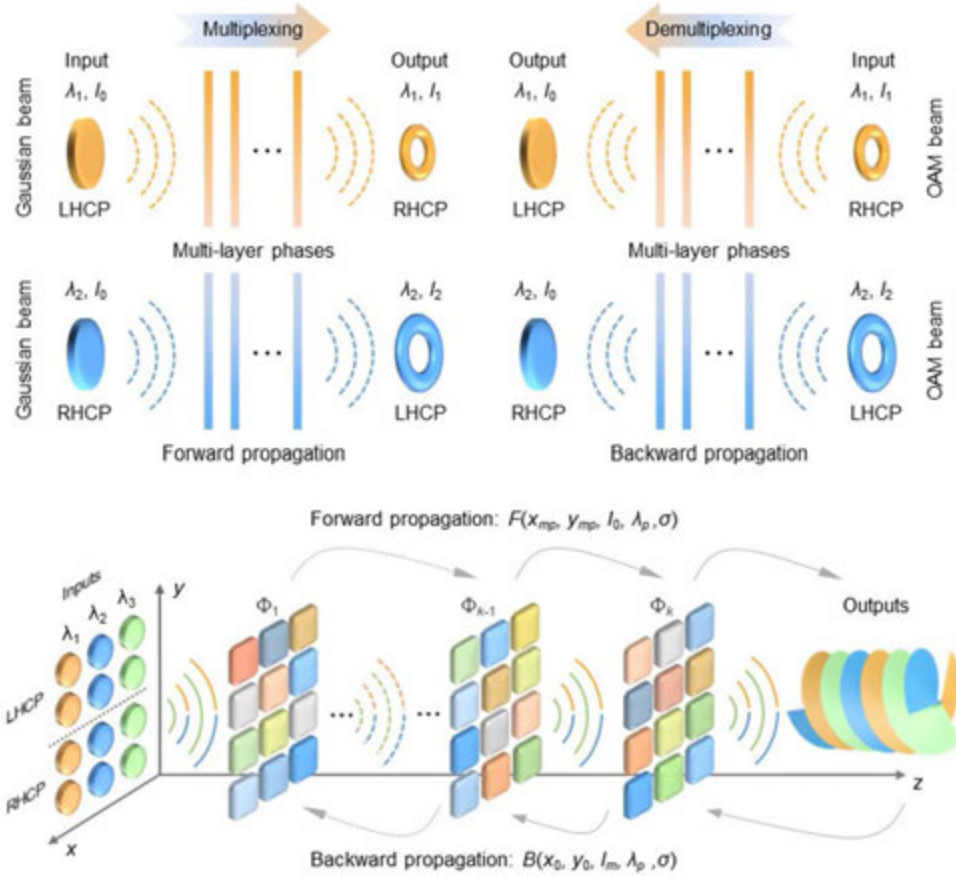


Figure 5. Conceptual principle of multi-dimensional hybrid (de)multiplexing in OAM modes, polarizations, and wavelengths [40].

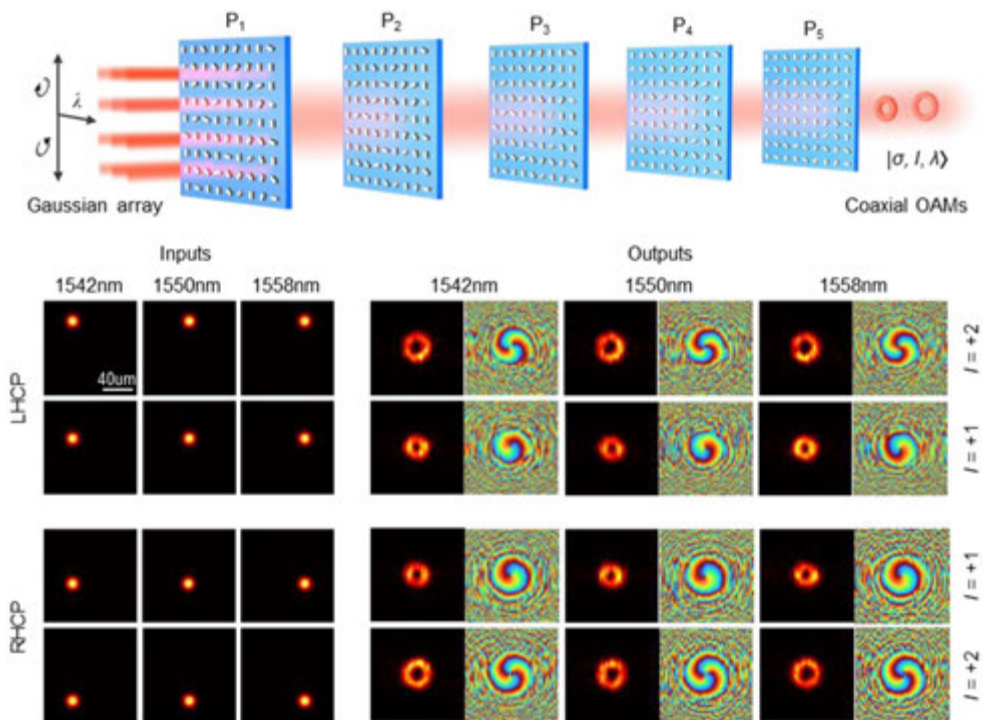


Figure 6. Optical characterization of multi-dimensional multiplexing on a five-layer diffractive metasurface platform, presenting the results of the device's performance in coupling simultaneous OAM, polarization, and wavelength channels [40].

a composite phase architecture. This spin-decoupled design enables independent wavefront control for two circular polarization states, thereby supporting full multi-dimensional channel separation within a compact multi-layer platform.

To verify the proposed scheme, a multi-dimensional OAM (de)multiplexer was fabricated using five cascaded diffractive metasurfaces, enabling simultaneous coupling and separation across 12 channels spanning OAM, wavelength, and polarization. The system encodes two OAM topological charges ($l = +1, +2$), three wavelength channels (1542 nm, 1550 nm, and 1558 nm), and two orthogonal circular polarization states. Each metasurface comprises 256×256 meta-atoms, yielding a compact overall footprint of $204.8 \times 204.8 \mu\text{m}^2$. As illustrated in **Figure 6**, the input–output optical field transformations confirm successful multi-channel operation. The first and fourth rows depict the conversion of fundamental Gaussian beams into OAM modes with charge +2, while the second and third rows show transformation into OAM modes with charge +1. The corresponding intensity profiles and phase distributions validate the device's ability to perform multi-mode conversion and multi-channel coupling in a hybrid multi-dimensional scheme. Demultiplexing follows the inverse operational principle: Coaxial OAM beams carrying distinct combinations of wavelength and polarization are incident on the cascaded metasurfaces. Through spin-decoupled multi-layer phase modulation, the input vortices are reconverted into fundamental Gaussian beams that are spatially separated according to their spectral and polarization signatures. Output beams are

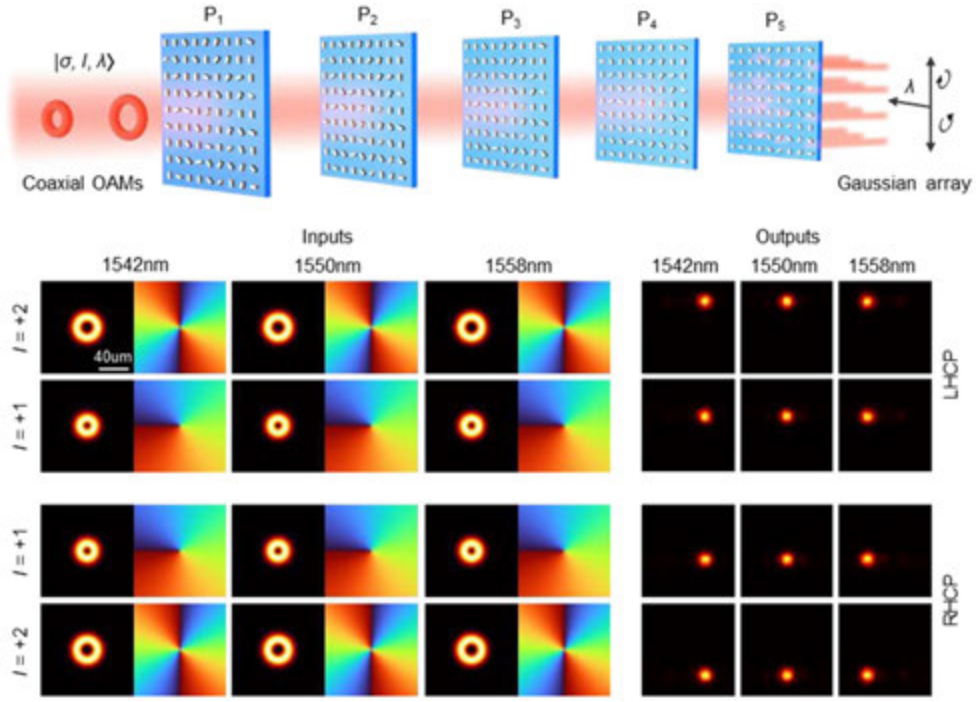


Figure 7. Optical characterization of multi-dimensional demultiplexing, displaying the results of the device's performance in separating OAM–polarization–wavelength channels in parallel [40].

directed to distinct ports, as demonstrated in **Figure 7**, confirming effective retrieval of the original Gaussian profiles at designated positions without crosstalk.

In contrast to the previously discussed (de)multiplexing schemes based on diffractive gratings and multi-foci phase plates, which are fundamentally limited by a $1/N$ power efficiency constraint, multi-layer cascaded modulation offers a highly flexible and efficient alternative. This methodology relies on performing independent unitary transformations on multiple input optical fields, a framework effectively implemented in MPLC and deep diffractive neural networks. Such a unified architecture has gained broad recognition in contemporary photonics research [64–67], enabling all-optical, high-efficiency mode conversion while supporting simultaneous control across multiple modes, spatial channels, and mode bases. Consequently, the multi-layer cascaded approach establishes a promising platform for high-capacity optical communications and advanced all-optical signal processing, offering new pathways for innovation in multi-dimensional photonic systems. Collectively, these multi-dimensional multiplexing and demultiplexing methodologies demonstrate distinct capabilities in cross-domain manipulation of OAM modes, thereby advancing the development of sophisticated multi-dimensional optical communication systems. A systematic comparison of their key performance metrics is provided in **Table 1**, offering a clear visualization of their respective advantages and operational characteristics.

Category	Mechanism	System integration level	Insertion loss	Mode capacity
OAM–polarization parallel (de)multiplexing	Dammann-vortex or grating-gradient interactions	High (single element)	High (1/N theoretical)	Medium (limited by diffraction orders)
OAM–wavelength parallel (de)multiplexing	Dispersion-engineered spiral-lens modulation	High (single element)	High (1/N theoretical)	Medium (limited by focusing efficiency)
OAM–polarization–wave-length parallel (de) multiplexing	Multi-layer phase diffractive modulation	Medium (multi-layer)	Low (multi-parameter optimization)	High (flexible spatial mapping)

Table 1.
Multi-dimensional OAM mode (de)multiplexing technologies using metasurfaces.

5. Conclusion

The advancement of metasurface-enabled multi-dimensional OAM mode (de) multiplexing has emerged as a critical solution to address the soaring capacity demands of optical communication systems in the big data era. This chapter has systematically reviewed three core metasurface-based architectures that capitalize on subwavelength programmability and multi-degree-of-freedom light manipulation to overcome intrinsic limitations of conventional optics – including single-mode operation, Bragg diffraction constraints, and the inability to jointly process OAM, polarization, and wavelength dimensions. The first approach adopts polarization-independent phase modulation, leveraging Dammann-vortex or grating-gradient interactions within orthogonal polarization channels to achieve polarization-insensitive OAM mode control for the coupling and symmetric separation of multiple beams. The second strategy employs dispersion-engineered multi-focal transformation, assigning a dispersion-enhanced spiral-lens phase function to wavelength channels for multi-wavelength-decoupled OAM spatial manipulation, facilitating multi-beam convergence and sorting. The third is based on a multi-layer phase-diffraction modulation architecture, which incorporates both polarization-dependent unitary transformations and wavelength-sensitive Fresnel diffraction propagation. This supports simultaneous spatial wavefront evolution controlled by polarization and wavelength, thereby enabling multi-dimensional (de)multiplexing across these parameters and OAM modes. Through representative case studies, this chapter has highlighted the operational principles and systematic categorization of these three mainstream methodologies. As optical communications progress toward ultra-high capacity and on-chip integration, metasurfaces are poised to play an indispensable role in bridging the gap between exponential data growth and the development of next-generation high-performance (de)multiplexing systems.

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Chapter 4

Radar-Absorbing Materials, Metamaterials, and Metasurfaces: Design Methods, Considerations, and Best Practices

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Abstract

Radar-absorbing materials (RAMs) are crucial for reducing radar cross-section (RCS), suppressing electromagnetic interference (EMI), and enhancing the efficiency of antennas and wireless systems. Several design approaches have been established, ranging from conventional dielectric and magnetic composites to multilayer structures, frequency-selective surfaces (FSS), and metamaterial-based absorbers. Each method provides specific advantages, such as wide bandwidth, compactness, or tunability, but also poses challenges, including structural thickness, fabrication complexity, or polarization dependence. More recent advances involve graphene- and metasurface-based absorbers, which offer ultra-thin, lightweight, and reconfigurable solutions suitable for emerging 5G/6G communication platforms, stealth applications, and multifunctional devices. This chapter presents a comprehensive overview of RAM design methods, key considerations, and best practices. The discussion emphasizes fundamental trade-offs between absorption depth, bandwidth, and thickness, as well as challenges related to angular stability, scalability, and integration with antenna systems. Practical case studies are included to illustrate the performance of multilayer absorbers, FSS and metamaterial-based designs, and hybrid configurations, highlighting their effectiveness in balancing design constraints with application demands. The chapter concludes by addressing future trends, such as machine learning-driven optimization, adaptive material platforms, and scalable fabrication techniques, which are expected to advance the development of broadband, lightweight, and highly efficient absorbers. By consolidating theoretical foundations, design strategies, and practical implementations, this work provides a structured reference for researchers and engineers aiming to develop next-generation RAMs for defense, aerospace, and modern communication systems.

Keywords: Radar-absorbing materials (RAMs), frequency-selective surfaces (FSS), metamaterial absorbers, electromagnetic wave absorption, and stealth technology

1. Introduction

The continuous development of advanced communication and sensing systems has intensified the demand for effective electromagnetic (EM) control technologies. Among these, radar-absorbing materials (RAMs) have become indispensable in both civilian and defense applications, playing a vital role in radar cross-section (RCS) reduction, EM interference (EMI) mitigation, and the improvement of antenna performance. Originally introduced during World War II for stealth purposes, RAMs have since evolved into multifunctional components integrated into a wide spectrum of platforms, including unmanned aerial vehicles (UAVs), airborne and naval stealth systems, and 5G/6G communication infrastructures. Recent antenna research has emphasized the importance of wideband, compact, and polarization-diverse designs for 5G, highlighting the need for absorber–antenna integration to enhance efficiency and reduce unwanted reflections [1–4]. The design of RAMs involves balancing multiple performance requirements, including broadband absorption, thin profiles, lightweight structures, and cost-effective fabrication. Achieving these characteristics is challenging because enhancements in one parameter often come at the expense of another; for example, bandwidth broadening typically increases thickness, while miniaturization may limit angular or polarization stability. To address these trade-offs, various design strategies have been developed, such as multilayer absorbers, frequency-selective surfaces (FSS), metamaterial-based absorbers, and, more recently, graphene- and metasurface-enabled structures. These approaches complement developments in millimeter-wave antennas designed for the 28/38 GHz band and higher frequencies, which demonstrate how absorbers can be used to suppress interference and improve gain [5–7]. Recent advances in machine learning (ML) and computational electromagnetics have opened new opportunities for RAM optimization by enabling rapid exploration of complex parameter spaces and automated performance enhancement. These innovations, together with scalable fabrication methods, are accelerating the transition of RAMs from specialized defense solutions to widespread applications in wireless communication, automotive radar, and EM compatibility (EMC) systems. For instance, compact and flexible multi-band antennas for WLAN, Wi-Fi, ISM, and sub-6 GHz applications [8, 9] further highlight the need for hybrid absorber solutions that can operate reliably across multiple frequency regimes. This chapter provides a detailed exploration of RAM design methods, considerations, and best practices, starting with fundamental theory, moving through practical design workflows and simulation tools, and culminating in advanced applications. Special attention is given to trade-offs between absorption depth, operational bandwidth, thickness, and polarization stability, supported by case studies and examples. The discussion concludes with perspectives on emerging technologies and research directions for the realization of next-generation broadband, lightweight absorbers for defense, aerospace, and modern communication platforms.

1.1 Historical perspective and motivation

The origins of RAM research date back to the early-to-mid-twentieth century, motivated primarily by military applications aimed at reducing the RCS of targets. The first generation, developed during World War II, employed ferrite and carbon-loaded composites to suppress radar reflections. Although these absorbers were relatively thick and narrowband, they laid the foundation of stealth technology.

The second stage expanded into EM interference (EMI) shielding and civilian uses, such as protecting sensitive electronics from EM pollution or in medical applications. The third stage involved the integration of absorbers with antennas, cloaking systems, and absorbing paints, as well as applications in anechoic chamber rooms. This stage also introduced FSS and multilayer absorbers, which enabled broader bandwidths and improved performance. The most recent stage is characterized by the development of metasurface and graphene-based absorbers, which provide ultra-thin, lightweight, and reconfigurable solutions suitable for modern technologies, including 5G/6G communications, UAV stealth, and adaptive multifunctional platforms. These absorbers operate by dissipating incident EM energy within the material instead of simply reflecting or scattering it. This is achieved by tailoring material parameters such as permittivity (ϵ), permeability (μ), and controlled loss, enabling impedance matching with free space and efficient energy dissipation [10]. Thus, as illustrated in **Figure 1**, RAM applications have evolved from simple stealth coatings to multifunctional EM solutions, addressing challenges across defense, communications, and electronic systems.

1.2 Recent developments and state of the art

The schematic in **Figure 2** illustrates how RAMs can be integrated with antennas to improve EM performance while reducing unwanted reflections. In conventional antenna systems, issues such as surface waves, mutual coupling, and back radiation degrade efficiency, bandwidth, and isolation. By incorporating RAM beneath or around the antenna, these parasitic effects can be dissipated, resulting in improved radiation efficiency, enhanced front-to-back ratio, and significantly reduced mutual coupling in array or MIMO systems. RAM-backed antennas also offer advantages in bandwidth broadening, where stabilizing impedance matching allows operation over wider frequency ranges. In stealth-related platforms, such integration further minimizes the RCS by absorbing incident signals rather than reflecting them, allowing antennas to remain electromagnetically “hidden” while fully functional. Over the last several years, several trends have emerged in RAM and antenna design that reshape how absorbers are conceived and deployed. Metamaterials and

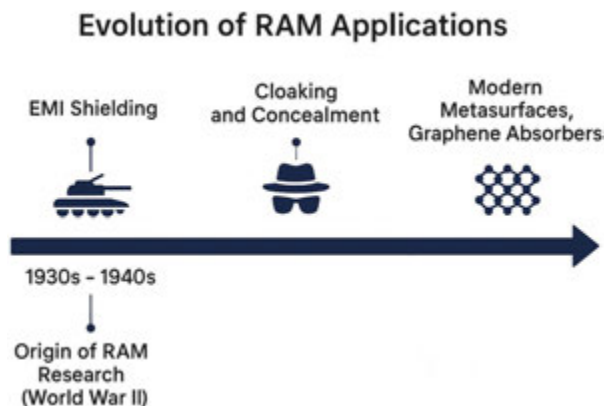


Figure 1.
Evolution of RAM applications from.

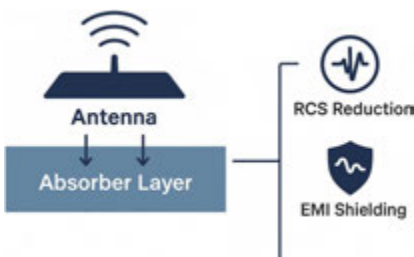


Figure 2.
Schematic representation of RAM integration with an antenna.

metasurfaces have enabled unprecedented control over surface impedance, supporting ultra-thin absorbers with high absorption across targeted bands. Deep learning and optimization techniques have been successfully applied to design metasurface-based antennas with enhanced gain, isolation, and bandwidth [11]. Another major trend is multiband and wideband antenna–absorber integration, where metamaterial or RAM layers are embedded within antenna systems to simultaneously improve antenna performance while reducing RCS. A recent study demonstrated a metamaterial-based tri-band compact MIMO antenna enhanced with MTM array elements for improved isolation across both microwave and millimeter-wave bands [12]. Metamaterials are also widely used for performance enhancement, including antenna size reduction, bandwidth extension, surface wave suppression, and in-phase reflection tuning [13]. Furthermore, RCS reduction in patch antennas has been achieved through metamaterial loading, polarization conversion layers, and fractal–metamaterial structures, which suppress reflections while maintaining key antenna parameters. In summary, RAM–antenna integration represents a synergistic design approach that not only mitigates EMI but also enables compact, efficient, and stealth-capable antenna systems – particularly relevant for 5G/6G communications, UAV stealth platforms, and adaptive multifunctional devices.

1.3 Challenges and design objectives

The key challenges in the design of RAMs are illustrated in **Figure 3**, highlighting the complex trade-offs among absorption depth, bandwidth, thickness, polarization

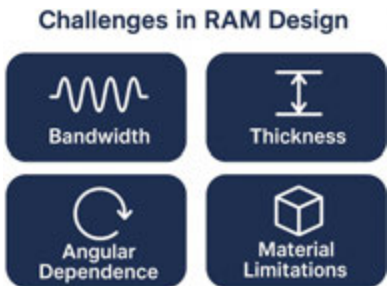


Figure 3.
Key challenges in the design of RAMs.

stability, and fabrication practicality. One of the most critical issues is the bandwidth–thickness trade-off, where achieving broadband absorption often requires thicker structures that contradict the demand for lightweight and conformal coatings. This limitation can be alleviated through multi-layer impedance-matching techniques or by employing metamaterial and metasurface absorbers that achieve wideband operation through multiple resonant modes within an ultra-thin profile. Another major challenge arises from polarization sensitivity – many RAMs perform effectively only for a single polarization state, leading to performance degradation under arbitrary field orientations. This can be mitigated using symmetric geometries, polarization-insensitive resonators, or three-dimensional (3D) metamaterial configurations that ensure uniform absorption for both TE and TM modes. Similarly, angular stability is crucial in real-world applications; as the incidence angle deviates, the effective impedance changes, resulting in poor matching and higher reflections. To overcome this, gradient-index designs, resistive sheets, and anisotropic layers can be utilized to maintain stable absorption over wide angular ranges. From a fabrication perspective, material losses and manufacturing tolerances pose additional constraints. Small deviations in permittivity (ϵ_r) and permeability (μ_r) can significantly affect impedance matching. These issues can be addressed through hybrid material compositions, additive manufacturing, and machine-learning-assisted optimization, which enable precise control of EM parameters.

1.4 Scope and contribution of this chapter

This chapter aims to provide a comprehensive set of design guidelines for RAMs, especially when they are to be integrated with antenna systems, covering: (1) fundamental EM concepts behind absorption, including key material parameters and wave interaction models; (2) input-output design parameters: how choices in material, geometry, and wave conditions map to performance metrics like reflection loss, absorption, RCS, and bandwidth; (3) tools and workflows: simulation platforms, optimization techniques, fabrication methods, and measurement procedures [14]; and (4) design cases and skill-building techniques: scaling laws, metamaterial absorbers (MMAs), layered structures, graphene, or other advanced materials.

2. Materials and methods

The design and optimization of RAMs require a systematic approach that combines EM theory, simulation workflows, fabrication techniques, and experimental validation. This section outlines the input and output parameters, simulation tools, fabrication methods, and measurement techniques used to develop high-performance RAM structures suitable for modern applications, including stealth technology and EMI mitigation.

2.1 Input and output design parameters

RAM design begins with defining the input materials and structural parameters that influence EM wave absorption, and identifying output performance metrics to guide the optimization process.

2.1.1 Input electrical parameters

The EM behavior of RAMs depends strongly on their constitutive parameters:

- **Permittivity (ϵ):** Determines how the material stores electrical energy and affects impedance matching.
- **Permeability (μ):** Governs magnetic energy storage and is crucial for low-frequency absorption [15].
- **Conductivity (σ):** Represents ohmic losses and is critical in resistive layers and graphene-based absorbers.
- **Loss Tangents ($\tan \delta$):** Represent dielectric and magnetic dissipation mechanisms that convert EM energy into heat.

The frequency-dependent nature of these parameters must be considered in both simulation and practical design.

2.1.2 Input physical parameters

- **Thickness of layers:** Directly influences bandwidth and peak absorption frequency as shown in **Figure 4**.
- **Geometry of patterns:** For FSS and metasurfaces, the shape and spacing of unit cells (e.g., circular, square, or fractal patterns) define the resonance characteristics.
- **Material type for conductors:** Copper, nickel, or graphene films may be used, depending on design goals.

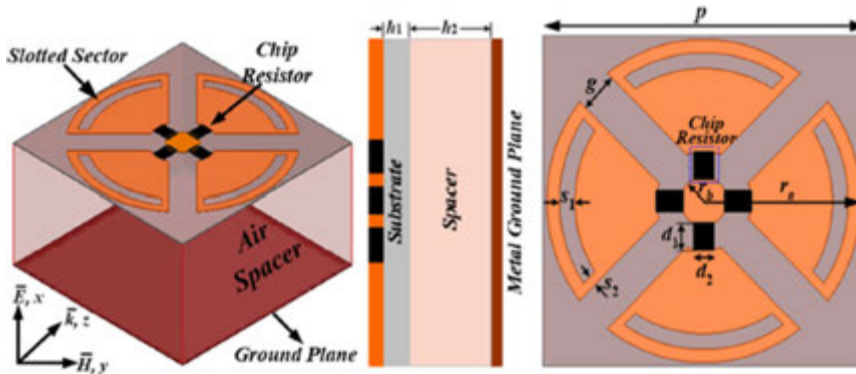


Figure 4. The basic physical parameters, including thickness, patch geometry, and incident wave characteristics: (a) unit cell geometry, (b) side view, and (c) top view [16].

2.1.3 Input wave parameters

Performance also depends on the properties of incident waves:

- **Frequency range** of operation.
- **Polarization** (linear, circular).
- **Incidence type**: Transverse electric (TE) or transverse magnetic (TM).
- **Angle of incidence**: Normal or oblique.

2.1.4 Output performance metrics

The target performance parameters for RAM design include:

- **Reflection coefficient (S_{11})**: Indicates the fraction of the incident energy that is reflected back.
- **Absorption (A)**: Calculated as shown in Eq. (1):
$$A = 1 - |S_{11}|^2 - |S_{21}|^2 \quad (1)$$
- **Bandwidth**: The frequency span where reflection loss remains below a specified threshold (e.g., -10 dB).
- **RCS reduction**: Essential for stealth and EMI control.

2.2 Simulation tools

Numerical simulation is essential for predicting absorber behavior before fabrication. Several specialized tools are used, as shown in **Figure 5**, such as high-frequency structure simulator (HFSS), computer simulation technology (CST), MATLAB, and FEKO, respectively.

2.3 Design-to-product workflow

The development of RAMs follows a design-to-product workflow, which integrates computational modeling, prototype fabrication, and performance evaluation. This iterative process ensures that the final product meets the targeted EM specifications and operational requirements.



Figure 5.
Various simulation tools.

The general flow of this process is illustrated in **Figure 6**, which consists of three major stages:

(1) Design through simulation:

- Initial design concepts are modeled using EM simulation tools such as CST Microwave Studio or HFSS.
- Key design variables include material properties (ϵ , μ , $\tan \delta$), physical geometry, and wave incidence conditions.
- Optimization techniques, including genetic algorithms and parametric sweeps, are applied to refine performance before fabrication.

(2) Prototype fabrication:

- The optimized design is fabricated using appropriate manufacturing methods.
- For dielectric absorbers, this involves the controlled mixing of polymer matrices with conductive or magnetic fillers.
- For meta-surface absorbers, precision printing or etching techniques are used to create patterned conductive structures.

(3) Performance testing:

- Prototypes are evaluated using measurement techniques to determine reflection loss, transmission, and bandwidth.
- Results are compared with simulations to verify accuracy and identify discrepancies.
- If performance deviates from design expectations, feedback is used to refine the simulation model, forming a closed-loop improvement cycle.

2.4 Performance measurement techniques

Validation of simulation results requires precise experimental measurements. These tests are divided into two main categories:

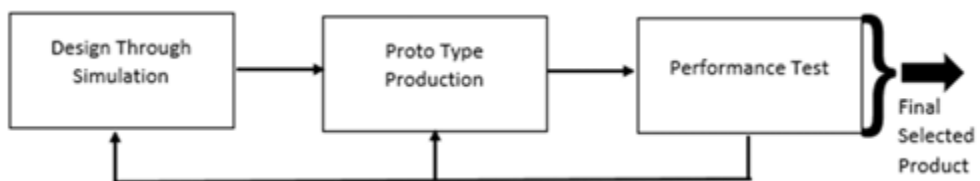


Figure 6.
General design-to-product workflow for RAM development.

2.4.1 Material electrical property measurements

- Determining permittivity (ϵ), permeability (μ), and loss tangents using:
 - Coaxial line or parallel plate fixtures.
 - Resonant cavity measurements for high accuracy.
 - Free-space measurements for broadband characterization [17].

2.4.2 Structural performance measurements

- **Reflection and transmission:** Measured using Vector Network Analyzers (VNAs).
- **Anechoic chamber testing:** For scattering and far-field performance.
- **RCS measurement:** Evaluates stealth characteristics of integrated antenna–RAM systems.

2.5 Fabrication techniques

2.5.1 Metamaterial fabrication

Metamaterials are fabricated using:

- PCB-based lithography for planar unit-cell patterns.
- 3D printing for advanced 3D absorbers.
- Laser etching for high-resolution structures [18].

2.5.2 Dielectric sheet fabrication

Dielectric RAMs are made by mixing a polymer base with:

- Carbon-based fillers to control permittivity.
- Magnetic powders (e.g., MnZn ferrite) are used to increase permeability.
- Conductive additives to adjust loss tangents.

2.6 Measurement techniques of permittivity and permeability

Modern measurement systems combine VNAs, precise fixtures, and automated data acquisition.

- **Permittivity:** Measured using impedance analysis, parallel plate methods, or free-space techniques.
- **Permeability:** Derived from reflection coefficients and cavity-based methods.

3. Results

This section presents the performance of various RAM structures designed using the methods described. Each design skill is illustrated with figures and supported by simulation and measurement results.

3.1 Scaling laws for frequency shifting

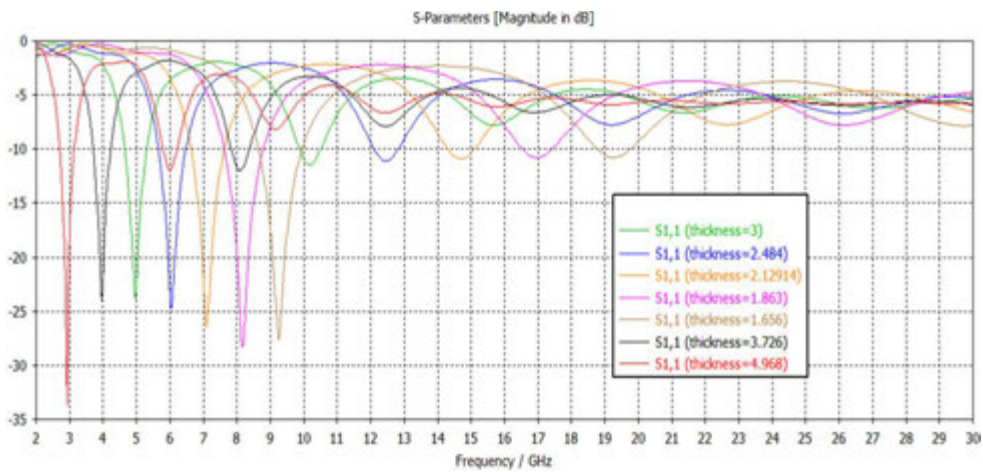
When a structure performs well at an undesired frequency band, scaling techniques can shift its resonance. The scaled model, designed at a frequency, equals the value of the operating frequency multiplied by the scaling factor (k). As an example, if the dimension of the model is $1/k$ of the scaled model, the new performance is shifted to kF_o (where F_o is the operating frequency of the original model). The scaling factor (k) relates to the old and new operating frequencies as given in Eq. (2):

$$k = \frac{\text{currentfrequency}(f_1)}{\text{newfrequency}(f_2)} \text{ or } k = \frac{\text{newdimensions}}{\text{olddiensions}} \quad (2)$$

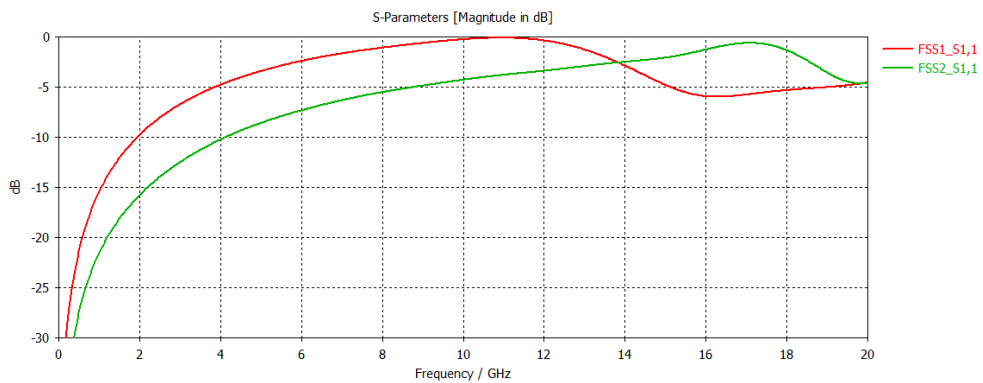
The scaling principles apply when the structure is geometrically scaled by a factor of k . As the physical length, wavelength, and area are reduced by factors of $1/k$ and $1/k^2$ respectively, the operating frequency increases proportionally by kF , maintaining EM similarity. Despite these geometric and frequency changes, the intrinsic material properties such as permittivity, permeability, and impedance remain unchanged (except for frequency-dependent materials), ensuring consistent EM behavior. This scaling law is particularly useful in predicting the performance of scaled absorbers and metamaterial unit cells, allowing designers to transition RAM concepts effectively between frequency regimes (e.g., from microwave to millimeter-wave bands) without altering the fundamental EM response.

Figure 7(a) illustrates the effect of absorber thickness on the simulated reflection coefficient (S_{11}) across a frequency range of 2–30 GHz. The results reveal a clear relationship between thickness and resonance behavior. Increasing the absorber thickness shifts the primary resonance toward lower frequencies, in accordance with the quarter-wave cancellation principle, while reducing thickness moves the resonance toward higher frequencies. The sample with the greatest thickness (4.968 mm, red curve) exhibits strong absorption at frequencies below 5 GHz, achieving a reflection null approaching -30 dB, indicating excellent impedance matching and high energy dissipation. Conversely, the thinnest sample (1.656 mm, brown curve) demonstrates peak absorption near 10 GHz, offering broader frequency coverage but with a shallower reflection null. These results highlight a fundamental design trade-off: (1) thicker absorbers deliver superior low-frequency performance and deeper nulls, making them suitable for narrowband stealth applications; (2) thinner absorbers provide broader operational bandwidths, ideal for wideband EMI shielding or multi-band communication systems. By selecting an optimal thickness, designers can tailor the absorber to balance compactness, bandwidth, and peak absorption levels to meet specific application requirements.

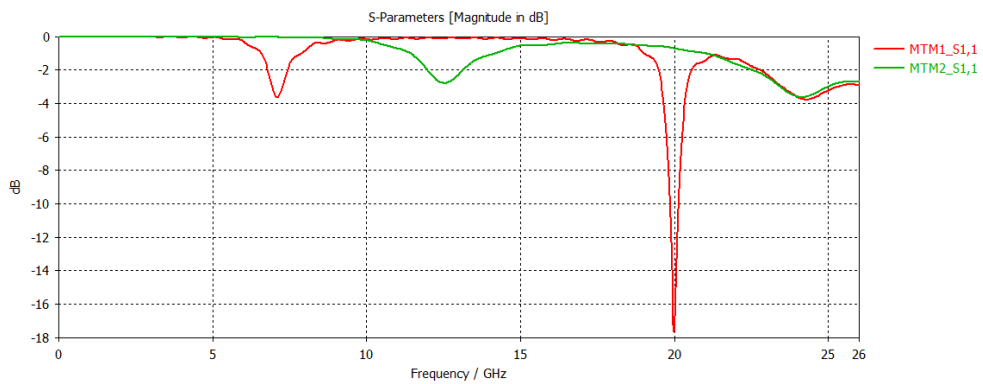
Figure 7(b) compares the simulated reflection coefficients (S_{11}) of two distinct FSS configurations across the frequency range of 0–20 GHz. The results indicate a significant difference in the performance of the two structures. The FSS1 design (red curve) exhibits strong low-frequency absorption, with a deep reflection null



(a) a dielectric sheet thickness scaling



(b) FSS ring 8 and 5 mm diameter



(c) MMT ring 8 and 5 mm diameter

Figure 7.
Thickness-based scaling of dielectric sheets and printed patches.

reaching approximately -30 dB near 1 GHz. However, its performance gradually deteriorates at higher frequencies, with reflection levels rising above -5 dB beyond 12 GHz. This behavior suggests that FSS1 is optimized for narrowband low-frequency applications, such as targeted stealth or specific EMI suppression bands.

In contrast, the FSS2 design (green curve) demonstrates broader frequency coverage, maintaining reflection levels below -10 dB over a wider range, particularly between 6 and 18 GHz. Although the peak absorption is slightly weaker than FSS1, the extended operational bandwidth makes FSS2 better suited for wideband applications, such as broadband RCS reduction or multi-band communication systems. These results highlight the design trade-off between peak absorption depth and bandwidth: (1) FSS1 prioritizes maximum absorption at a specific frequency band, and (2) FSS2 offers consistent performance across a wider frequency spectrum. Such comparative studies are critical in selecting an appropriate FSS design based on the target application requirements.

Figure 7(c) compares the simulated reflection coefficients (S_{11}) of two metamaterial-based absorber configurations, MTM₁ and MTM₂. The results highlight distinct differences in their resonance behavior and impedance matching. The MTM1 design (red curve) exhibits multiple resonances, with moderate absorption observed near 6 GHz and a pronounced deep reflection null at approximately 20 GHz, reaching below -17 dB. This indicates highly efficient impedance matching and strong absorption at the higher frequency, making MTM₁ suitable for narrowband high-frequency stealth applications. In contrast, the MTM₂ design (green curve) demonstrates more stable performance across the entire frequency band, maintaining reflection coefficients between -2 and -5 dB over a wide range. While it lacks a deep absorption peak, its smooth response suggests that MTM2 is optimized for broadband applications, such as EMI shielding or wideband RCS reduction. These results emphasize the trade-off between narrowband and broadband absorber designs: (1) MTM1 prioritizes peak absorption at a specific frequency, and (2) MTM2 provides moderate, consistent performance across a broader spectrum. Such comparisons are essential for tailoring MMAs to application-specific requirements, whether targeting stealth technology or broadband EMC.

3.2 Effect of material properties

In this section, we will discuss the effect of different parameters such as permittivity, permeability, and varying loss tangents on RAM performance.

3.2.1 Varying permittivity and permeability

Changing the dielectric constant ϵ and the permeability μ alters the electric thickness of the dielectric sheet (keeping the physical thickness constant). This may produce results similar to changing the thickness.

Figure 8(a) shows the effect of varying the relative magnetic permeability (μ_r) on the reflection coefficient (S_{11}) of a RAM structure across the 2–20 GHz frequency range, with relative permittivity (ϵ_r) fixed at 30 and a constant thickness of 3 mm.

The results reveal that increasing μ_r shifts the primary absorption resonances toward lower frequencies while simultaneously deepening the reflection nulls.

For $\mu_r = 4$ (red curve), the structure exhibits moderate absorption with a reflection minimum of approximately -20 dB near 11 GHz, but performance at lower frequencies remains limited.

Increasing μ_r to 10 (green curve) results in stronger low-frequency resonance and deeper absorption near 10 GHz, reaching approximately -30 dB.

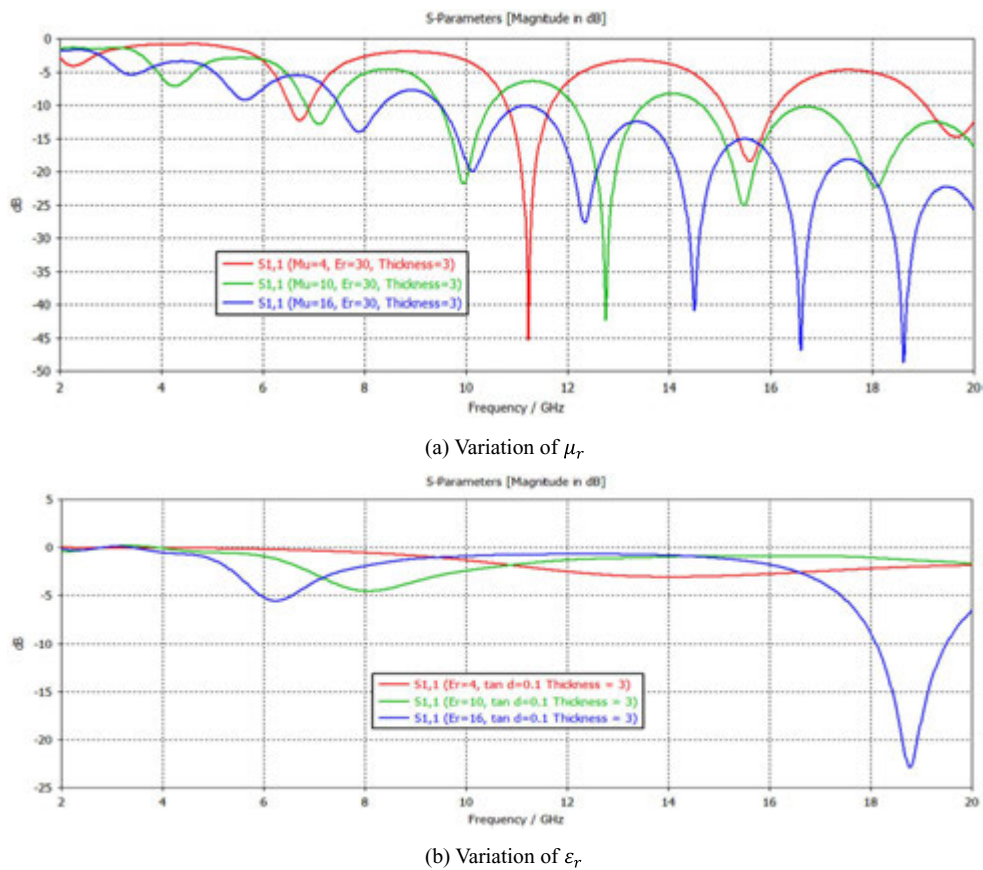


Figure 8.
 Variation of material properties ϵ_r and μ_r .

At $\mu_r = 16$ (blue curve), multiple deep nulls occur below 10 GHz, with the lowest reflection reaching nearly -45 dB, indicating excellent impedance matching and high energy dissipation.

This behavior aligns with the impedance matching principle, where varying permeability enables better coupling between the incident wave and the absorber. However, it also introduces increased resonance density, which can affect broadband performance. These findings highlight the critical role of μ_r in RAM design: (1) Low μ_r : Suitable for narrowband absorbers targeting higher frequencies. (2) High μ_r : Ideal for strong, low-frequency absorption but may require additional optimization to maintain wideband stability. By carefully selecting μ_r , designers can tailor the absorption characteristics to specific applications, such as low-frequency stealth or broadband EMI shielding.

Figure 8(b) illustrates the impact of varying relative permittivity (ϵ_r) on the reflection coefficient (S_{11}) for RAM structures across the 2–20 GHz frequency range. The simulations were performed while keeping the thickness constant at 3 mm and the loss tangent fixed at 0.1 to isolate the effect of permittivity. The results indicate that increasing ϵ_r causes the primary resonance frequency to shift toward higher frequencies, while also influencing the depth of absorption.

For $\epsilon_r = 4$ (red curve), the reflection response remains shallow across the band, with minimized absorption, indicating poor impedance matching with free space.

At $\epsilon_r = 10$ (green curve), a moderate absorption dip occurs near 6 GHz, demonstrating improved performance.

When $\epsilon_r = 16$ (blue curve), a pronounced reflection null appears near 18 GHz, reaching nearly -22 dB, signifying excellent impedance matching and strong energy dissipation.

This behavior aligns with EM wave-material interaction theory, where higher permittivity materials support stronger electric field coupling, leading to deeper and sharper resonances. However, extremely high permittivity may restrict bandwidth, as evident in the narrow absorption peak at higher frequencies for $\epsilon_r = 16$. This design has the following implications: (1) Low ϵ_r values result in broad but shallow absorption, ideal for EMI shielding where partial reflection suppression is sufficient. (2) High ϵ_r values produce strong, narrowband absorption suited for applications like radar stealth, where high attenuation at specific frequencies is required. By carefully selecting ϵ_r , designers can tune RAM performance to match operational frequency bands and application requirements.

3.2.2 Varying loss tangents

Changing $\tan\delta_e$ and $\tan\delta_m$ keeping ϵ_r , μ_r and sheet thickness unchanged.

Figure 9(a) presents the effect of varying the electric loss tangent ($\tan\delta_e$) on the reflection coefficient (S_{11}) of RAM over the 1–20 GHz frequency range. The simulations were performed with fixed material properties ($\epsilon_r = 1.2$, $\mu_r = 1.2$) and constant thickness (0.035 m).

The results clearly indicate that increasing $\tan\delta_e$ enhances both the absorption depth and bandwidth:

At a low loss tangent ($\tan\delta_e = 0.05$, red curve), the absorber exhibits relatively shallow reflection levels, with S_{11} remaining above -40 dB across most of the spectrum, indicating limited energy dissipation and poor overall performance.

- With a moderate loss tangent ($\tan\delta_e = 0.1$, green curve), the reflection decreases significantly, achieving -50 to -60 dB over a wide frequency range, demonstrating improved EM wave attenuation.

For a high loss tangent ($\tan\delta_e = 0.15$, blue curve), the absorber reaches deep reflection nulls of up to -85 dB, coupled with broadband absorption characteristics, indicating highly effective conversion of incident EM energy into heat.

This behavior is attributed to the increased internal dielectric losses that occur with higher $\tan\delta_e$, which enhance wave dissipation within the absorber material. However, excessively high loss tangents may introduce unwanted resonances, as seen in the oscillatory response of the blue curve beyond 10 GHz, potentially complicating absorber stability. This design has the following implications: (1) Low $\tan\delta_e$: Suitable for narrowband or low-loss applications where minimal energy dissipation is desired. (2) Moderate $\tan\delta_e$: Ideal for achieving a balance between absorption depth and bandwidth for multi-purpose designs. (3) High $\tan\delta_e$: Provides superior broadband absorption but may require advanced optimization to prevent instability and resonance clustering. These findings demonstrate that adjusting the

dielectric loss tangent is a powerful method for tuning RAM performance to meet the specific requirements of stealth, EMI shielding, or broadband EMC applications.

Figure 9(b) demonstrates the effect of varying the magnetic loss tangent ($\tan\delta_m$) on the reflection coefficient (S_{11}) of RAM. The simulations were conducted with fixed parameters: relative permittivity ($\epsilon_r = 1.2$), relative permeability ($\mu_r = 1.2$), and thickness (0.035 m), to isolate the role of magnetic losses.

The results show that increasing $\tan\delta_m$ significantly enhances both the absorption depth and operational bandwidth:

At a low magnetic loss tangent ($\tan\delta_m = 0.05$, red curve), the absorber demonstrates only shallow absorption, with S_{11} values between -5 dB and -15 dB, indicating limited energy dissipation.

A moderate magnetic loss tangent ($\tan\delta_m = 0.1$, green curve) improves performance substantially, achieving reflection nulls of -25 to -30 dB across multiple frequencies, indicating effective impedance matching and improved magnetic energy conversion.

With a high magnetic loss tangent ($\tan\delta_m = 0.15$, blue curve), the absorber achieves even deeper reflection nulls, reaching below -50 dB near 8 GHz, along with wideband absorption across the entire frequency spectrum.

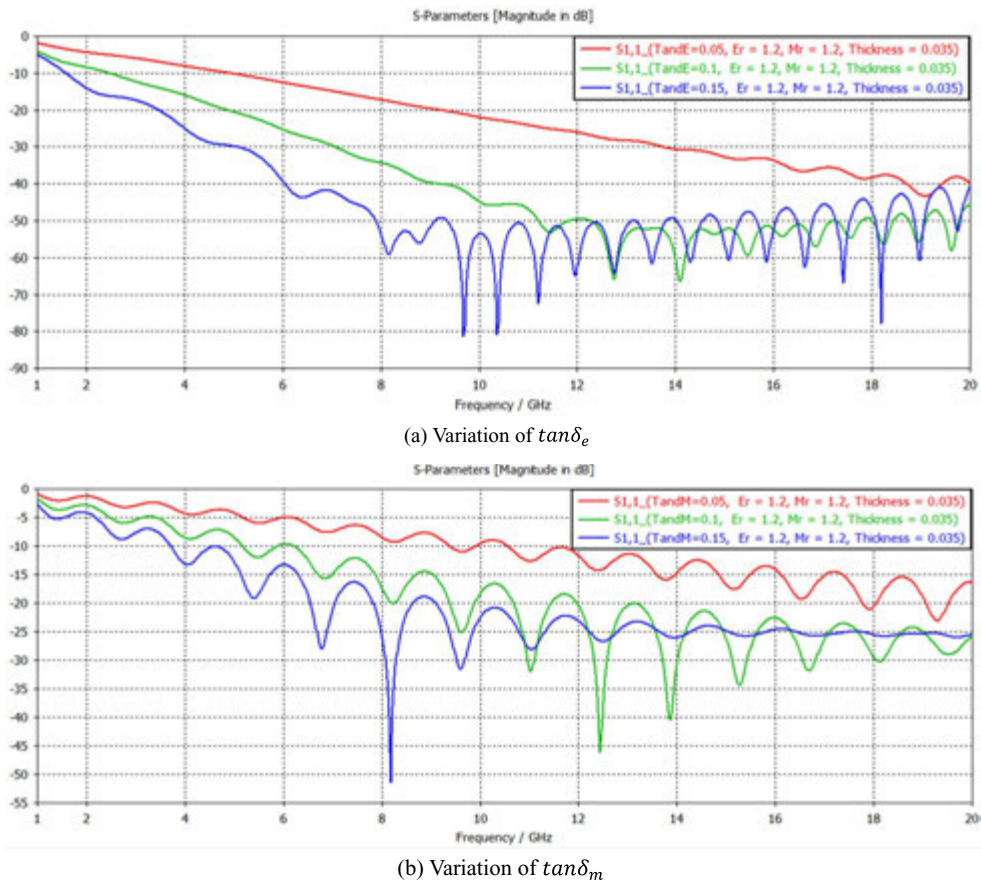


Figure 9.
Variation of material loss tangents $\tan\delta_e$ and $\tan\delta_m$

This behavior occurs because higher $\tan\delta_m$ enhances magnetic damping, increasing the conversion of EM wave energy into heat. However, as seen in the blue curve, excessive magnetic losses can also lead to resonance clustering, introducing oscillatory behavior and potentially reducing absorber stability at higher frequencies. This design has the following implications: (1) Low $\tan\delta_m$: Suitable for lightweight, low-loss designs targeting narrowband performance. (2) Moderate $\tan\delta_m$: Offers a balance between absorption depth and bandwidth, ideal for multi-purpose applications such as EMI shielding. (3) High $\tan\delta_m$: Delivers superior broadband absorption, critical for stealth technology and wideband EMC, but requires careful control to prevent instability. By carefully tuning $\tan\delta_m$, designers can optimize RAMs for specific performance requirements, tailoring them to operational frequency ranges and application constraints.

3.3 μ_r equal ϵ_r absorber sheets

When μ_r equals ϵ_r , the absorber matches free-space impedance, minimizing reflections. This type of absorber has a non-resonant type of performance and have promising performances. We run the CST program to study the general behavior of this type of absorber. For impedance, consider Eq. (3)

$$Z = Z_0 \sqrt{\frac{\mu_r}{\epsilon_r}} \quad (3)$$

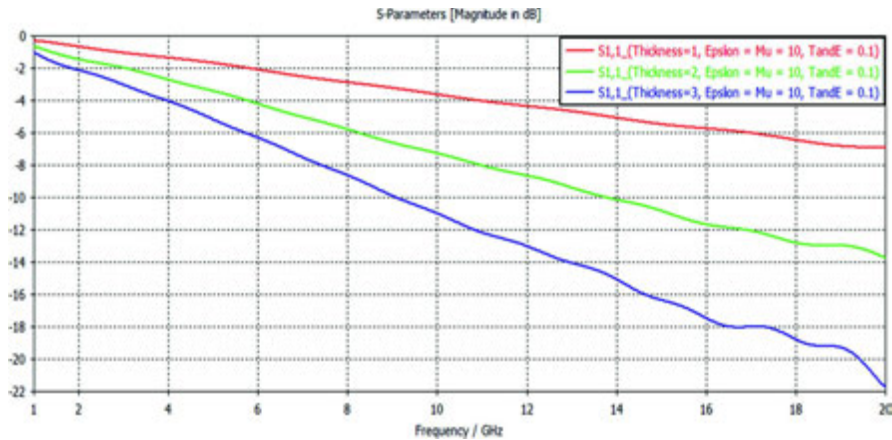
As Z and Z_o are the characteristic impedances of the absorber and free space, respectively, for μ_r equal ϵ_r , absorber Z matches free space impedance ($Z = Z_o$). It offers a high loss in a broad frequency band, over a specific low frequency limit for a selected minimum reflection threshold.

Figure 10(a) shows the influence of RAM thickness on the reflection coefficient (S_{11}) across the frequency range of 1–20 GHz. The material properties were fixed at $\epsilon_r = 10$, $\mu_r = 10$, and $\tan\delta_m = 0.1$, ensuring that thickness was the only variable in the study.

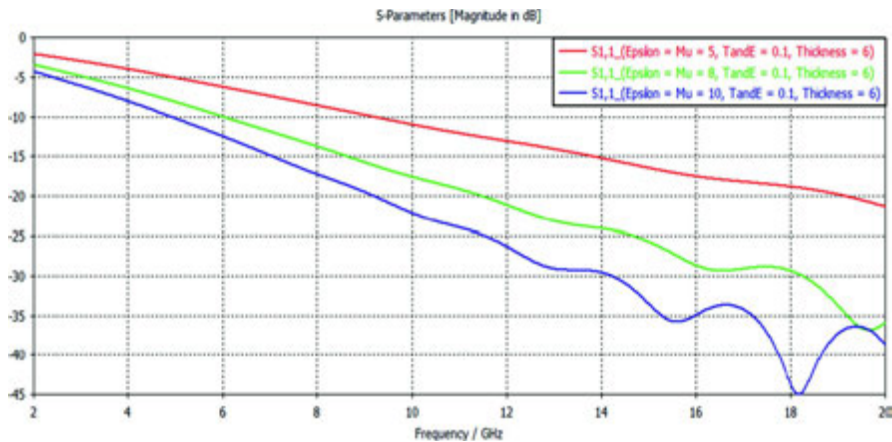
The results clearly demonstrate that increasing thickness significantly improves low-frequency absorption:

- At 1 mm thickness (red curve), the reflection coefficient remains relatively high, with S_{11} close to 0 dB at lower frequencies, indicating minimal absorption and poor impedance matching.
- At 2 mm thickness (green curve), the absorber shows improved performance, with S_{11} values decreasing steadily, reaching approximately -12 dB at 20 GHz, signifying moderate wave attenuation.
- At 3 mm thickness (blue curve), a dramatic improvement is observed, with S_{11} reaching -21 dB at 20 GHz and exhibiting strong low-frequency absorption. This demonstrates effective impedance matching and significant energy dissipation.

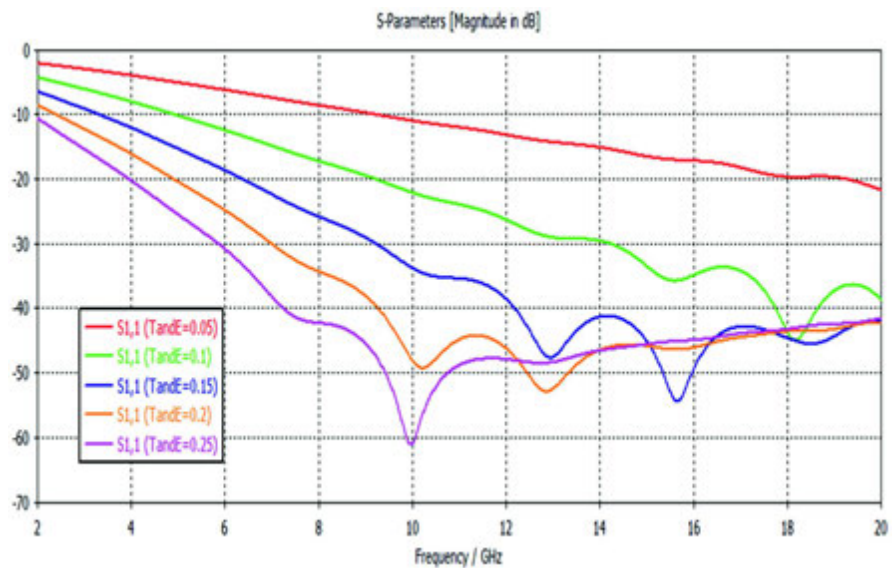
These results align with the quarter-wave cancellation principle, where thicker absorbers support stronger destructive interference of reflected waves, particularly at lower frequencies. As a result, thickness directly influences the operational frequency band and absorption depth. This design has the following implications:



(a) Varying the thickness



(b) Variable μ_r , equal ϵ_r values



(c) Varying loss tangent

Figure 10.
Impedance matching sheet.
105

(1) Thinner absorbers are lightweight and compact but have limited low-frequency performance, making them suitable for high-frequency or space-constrained applications. (2) Thicker absorbers achieve deeper nulls and broader low-frequency absorption, ideal for stealth and EMI shielding where weight and size are less critical. (3) Designers must balance performance, weight, and physical size when selecting the appropriate thickness for a given application.

Figure 10(b) shows the relationship between the material's relative permittivity (ϵ_r) and permeability (μ_r) and the reflection coefficient (S_{11}) across the frequency range of 2–20 GHz. The simulations were performed with a constant thickness of 6 mm and a fixed loss tangent of 0.1 to isolate the effects of these fundamental EM parameters.

The results reveal that increasing ϵ_r and μ_r significantly improves absorber performance by lowering the reflection coefficient and broadening the operational bandwidth:

For $\epsilon_r = \mu_r = 5$ (red curve), absorption performance is limited, with S_{11} remaining above -15 dB across most of the frequency range, indicating insufficient impedance matching.

Increasing to $\epsilon_r = \mu_r = 8$ (green curve) enhances absorption, with reflection nulls reaching -30 dB between 14 and 18 GHz, showing improved energy dissipation and better impedance matching with free space.

At $\epsilon_r = \mu_r = 10$ (blue curve), the absorber achieves broadband and deep absorption, with S_{11} falling below -40 dB near 18 GHz, making it highly effective for wideband stealth and EMI shielding applications.

This behavior is explained by the impedance-matching principle:

- When the intrinsic impedance of the absorber approaches that of free space (Z_0), reflections are minimized, and incident electromagnetic waves are efficiently absorbed and dissipated.

Higher ϵ_r and μ_r increase the stored electric and magnetic energy, enabling better coupling with incoming waves and more effective destructive interference within the absorber layers.

This design has the following implications: (1) Low ϵ_r, μ_r : Lightweight designs with limited absorption, suitable for high-frequency or partial shielding. (2) Moderate ϵ_r, μ_r : Balanced designs offering reasonable absorption and moderate bandwidth. (3) High ϵ_r, μ_r : Optimal for broadband absorption and deep nulls, ideal for advanced stealth systems and wideband EMI mitigation. These findings emphasize the importance of selecting materials with suitable EM properties to achieve desired performance targets while considering factors such as weight, cost, and fabrication complexity.

Figure 10(c) illustrates the influence of the electric loss tangent ($\tan\delta_e$) on the reflection coefficient (S_{11}) of RAM across the 2–20 GHz frequency range. As $\tan\delta_e$ increases from 0.05 to 0.25, a significant enhancement in both absorption depth and operational bandwidth is observed. At low loss tangent values ($\tan\delta_e = 0.05$), the material exhibits poor energy dissipation, with S_{11} remaining above -20 dB, indicating ineffective impedance matching and narrowband performance. Increasing moderate levels ($\tan\delta_e = 0.1$ and 0.15) results in the emergence of deep resonance dips, with a pronounced minimum near 10 GHz, demonstrating highly efficient absorption at targeted frequencies. At higher values ($\tan\delta_e = 0.2$ and 0.25),

the response transitions into a smoother and broader bandwidth absorption profile, suitable for wideband applications such as EMI shielding and RCS reduction. These results highlight the critical role of optimizing $\tan\delta_e$ to balance absorption depth and bandwidth, enabling the design of high-performance RAM structures tailored for specific stealth and EMC applications.

3.4 Cascaded dielectric laminates

This structure can be easily simulated by a multi-section transmission line, as shown in **Figure 11**.

Figure 11(a) shows the physical layout, where individual absorber sheets are arranged sequentially, each with distinct material properties and thicknesses ($d_1, d_2, \dots, d_n$). Meanwhile, **Figure 11(b)** presents the equivalent transmission line model, where each layer is represented by a segment with its characteristic impedance (Z_{o1} to Z_{on}). This model is essential for understanding the impedance-matching process across multiple interfaces. **Figure 11(c)** compares the absorptivity of single-layer designs versus the combined multilayer structure as shown: Layer 1 (green curve) provides limited performance, with absorptivity gradually increasing

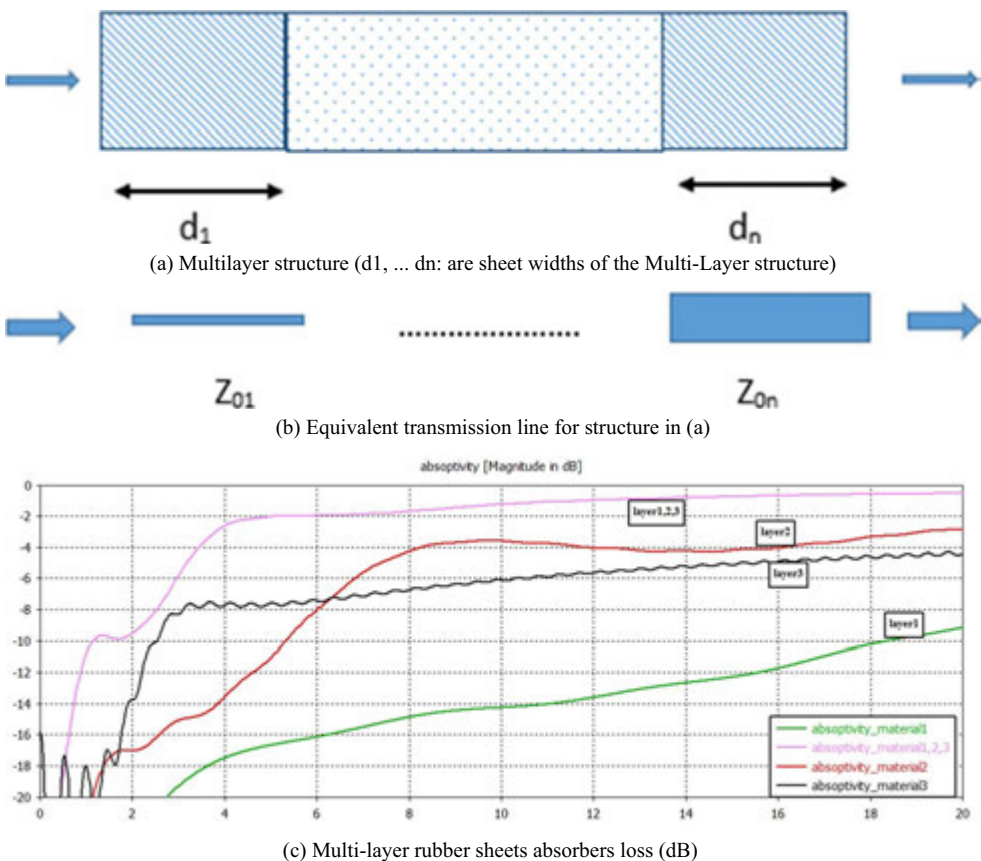


Figure 11.
Cascaded dielectric laminates.

from -18 dB at low frequencies to -10 dB at 20 GHz. Layer 2 (red curve) achieves better performance across mid frequencies, reaching around -5 dB above 10 GHz. Layer 3 (black curve) performs best at higher frequencies but exhibits a non-uniform response due to localized resonances. Combined multilayer (pink curve) maintains absorptivity close to -2 dB across the entire frequency range, demonstrating broadband and stable performance superior to any single layer.

The multilayer structure significantly enhances broadband absorption through:

- (1) Gradual impedance matching: Each successive layer transitions closer to the free-space impedance, reducing surface reflections and enabling efficient wave penetration.
- (2) Complementary frequency coverage: Different layers target different frequency regions, collectively achieving wideband absorption.
- (3) Multiple resonance mechanisms: Layer stacking introduces a series of resonances, enhancing total energy dissipation within the structure.

While the multilayer approach delivers outstanding performance, it introduces trade-offs, including increased thickness, weight, and fabrication complexity. These factors must be optimized based on specific applications. For instance, lightweight, compact systems such as UAVs may prioritize fewer layers with targeted frequency response, while high-performance stealth systems may use more layers to achieve maximum broadband coverage. This design has the following implications: the results confirm that cascaded multilayer structures are ideal for broadband stealth and EMI shielding applications, where wide frequency coverage and high absorptivity are critical. However, successful implementation requires precise control of material parameters and layer sequencing to balance performance, weight, and manufacturability.

3.5 Cascaded μ_r equal ϵ_r laminates

We investigated a stepped parameter multi-layer of μ_r equal ϵ_r forms, we got encouraging results as shown in **Figure 12**, compared with results in Section 3.3. The structure consists of three layers, each with a thickness of 1 mm, but different EM properties:

Layer 1: $\epsilon_r = 4$, $\mu_r = 4$, $\tan\delta_e = 0.1$, $\tan\delta_m = 0.1$

Layer 2: $\epsilon_r = 6$, $\mu_r = 6$, $\tan\delta_e = 0.12$, $\tan\delta_m = 0.12$

Layer 3: $\epsilon_r = 10$, $\mu_r = 10$, $\tan\delta_e = 0.15$, $\tan\delta_m = 0.15$

The plot compares the absorption performance of the structure with and without a ground metal plate, highlighting the enhanced absorption achieved through grounding.

Figure 12 illustrates the simulated absorption performance of a three-layer stacked stepped absorber with and without a conductive ground plane. each layer was designed with a uniform thickness of 1 mm, while the relative permittivity ()

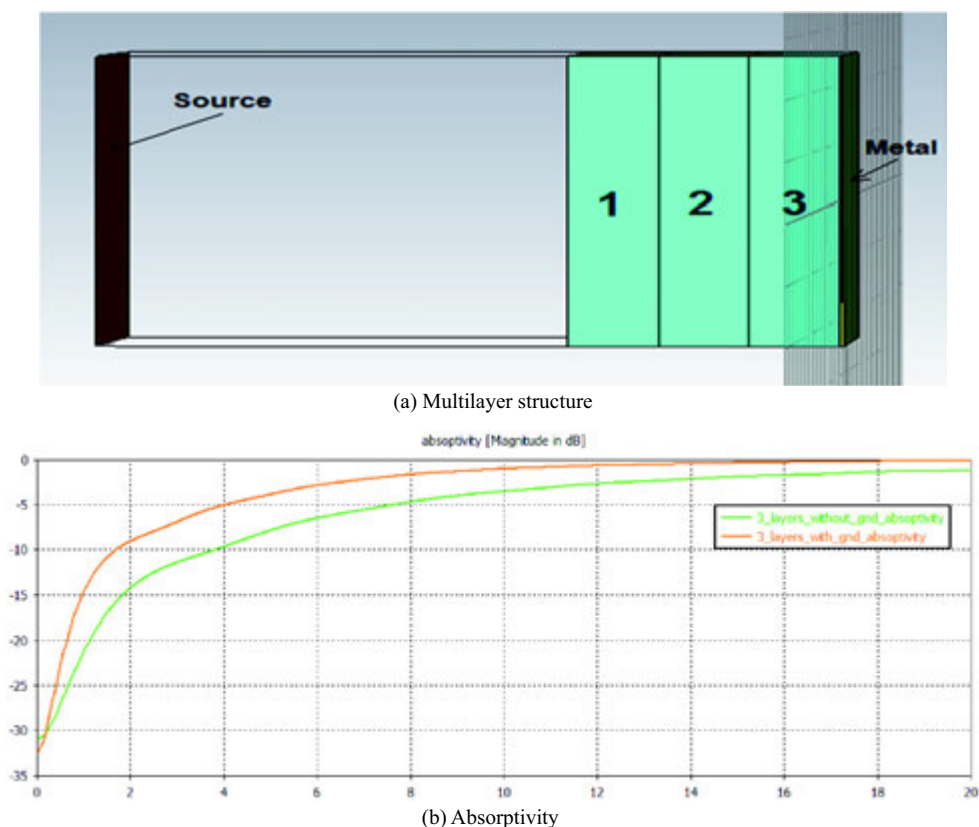


Figure 12.
 The three layers of stacked stepped μ_r equal ϵ_r sheets with/without a ground metal plate – absorption

and permeability (μ_r) were progressively increased across the stack to achieve a stepwise impedance matching profile. the EM loss factors were also varied, with the first, second, and third layers having ϵ_r , μ_r values of 4, 6, and 10, and loss tangents $\tan\delta_e$,) of 0.10, 0.12, and 0.15, respectively. The results clearly show that the addition of a ground metal plate significantly enhances absorber performance across the 2–20 GHz frequency range. without grounding (orange curve), the structure exhibits moderate absorption, with a maximum absorptivity of approximately -15 db occurring at mid-band frequencies. In contrast, the grounded configuration (green curve) demonstrates superior broadband performance, maintaining an absorptivity below -25 db across a wide spectrum and reaching peak values near -35 db. This improvement can be attributed to the destructive interference mechanism, whereby the ground plane reflects transmitted waves back into the absorber, promoting multiple internal reflections and increased energy dissipation.

Furthermore, the stepwise gradient of ϵ_r and μ_r across the layers provides a gradual impedance transition, reducing mismatch at the air–absorber interface and enhancing the coupling of incident EM energy into the structure. This combination of graded material properties and a reflective backing enables the generation of multiple overlapping resonances, which collectively broaden the absorption bandwidth.

These findings highlight the critical role of structural optimization in RAM design. The use of progressive material parameters in conjunction with a ground plane not only improves peak absorption levels but also delivers wideband stability, making the design highly suitable for advanced stealth technology and EMI shielding applications. Such multi-layered and grounded configurations are particularly advantageous in scenarios requiring low RCS signatures over a broad operational frequency range.

3.6 Combined FSS and rubber absorbers

This area has attracted a great number of researchers who have developed plenty of structures, each offering a distinguished performance. Anyone looking for a specific parameter form can find it. In our work, we investigated the mixed form of meta-surfaces and a lossy dielectric absorber, as shown in **Figure 13**.

Figure 13 compares the simulated reflection coefficients (S_{11}) of two FSS configurations, Case 5 (green curve) and Case 7 (orange curve), over the frequency range of 0–20 GHz. The performance differences between these designs highlight the significant impact of structural variations on EM behavior and absorption characteristics. The Case 5 design (green curve) demonstrates high selectivity and strong resonance, with a deep reflection null of approximately -10 dB occurring near 11 GHz. This indicates efficient impedance matching and localized absorption within a narrow frequency band, making it suitable for targeted narrowband applications, such as RCS reduction at specific operational frequencies. Outside this resonance region, the structure maintains very low reflection, signifying minimal interaction with EM waves across other parts of the spectrum. Conversely, the Case 7 design (orange curve) exhibits a broader but shallower response, with gradually improving reflection behavior from 2 GHz upwards and a smoother transition across the mid-frequency range. Although its peak absorption is weaker than Case 5, the gradual roll-off indicates that Case 7 is optimized for broader frequency coverage, such as EMI mitigation, where wideband but moderate absorption is preferable.

The performance differences observed can be attributed to geometric and material variations between the two cases:

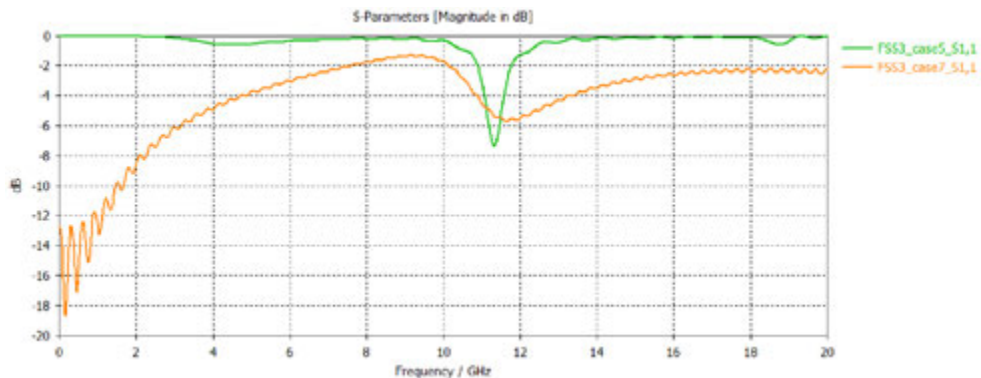


Figure 13.
The two different metamaterial patches stacked with a 5 mm air gap and lossy 5 mm dielectric sheets.

- (1) Case 5 likely employs resonant periodic elements, which create sharp, destructive interference at specific frequencies, leading to pronounced absorption peaks.
- (2) Case 7 likely incorporates a more resistive or gradient-based structure, dispersing energy dissipation across a wider range while limiting peak absorption depth.

These findings emphasize the design trade-off between narrowband selectivity and broadband performance:

- Case 5 is ideal for applications requiring high absorption at targeted frequencies, such as stealth operations and precision filtering.
- Case 7 is more suitable for general EMI shielding or systems that require moderate attenuation over a wide frequency range.

The comparative results in **Figure 10** underline the importance of multi-objective optimization in FSS design. By carefully balancing structural geometry, material properties, and periodicity, designers can tailor absorber performance to meet the specific needs of advanced radar, communication, and stealth systems. Future research should focus on hybrid designs, combining the sharp selectivity of Case 5 with the broadband behavior of Case 7, potentially through metamaterial integration or layered structures.

3.7 Planar and stacked patches

Figure 14 presents the simulated reflection coefficients (S_{11}) for three FSS configurations: FSS1 (blue curve), FSS2 (green curve), and FSS3 (red curve). The comparison highlights the complex interactions that occur when multiple FSS structures are integrated into a single composite design. The results clearly indicate that the response of FSS3 is not a linear summation of FSS1 and FSS2, as explicitly noted in the figure annotation ($FSS3 \neq FSS1 + FSS2$). While FSS1 and FSS2

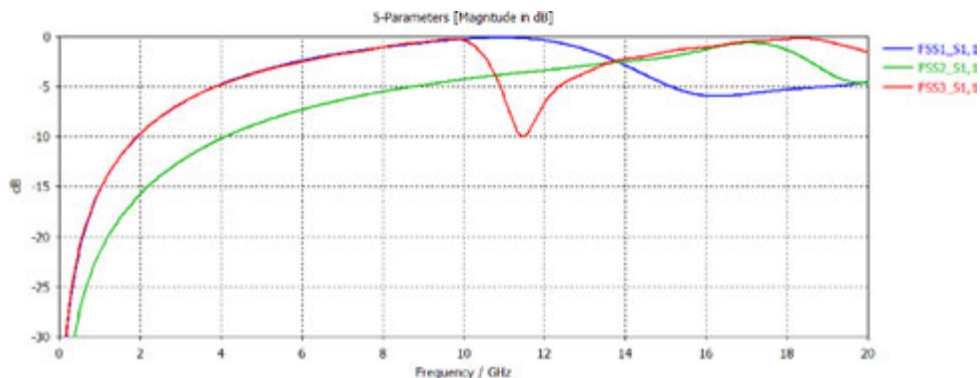


Figure 14.
 Patch 1 (curve FSS1) separate, patch 2 (curve FSS2) separate, and patch 1 and patch 2 both in one plane (curve FSS3)

individually exhibit distinct resonance behaviors – FSS1 with a strong, narrow resonance around 12 GHz and FSS2 with broadband but moderate absorption across the low- to mid-frequency range – the combined configuration, FSS3, produces a unique response profile. Specifically:

- At lower frequencies (<5 GHz), FSS3 follows a trend closer to FSS2, suggesting the dominance of the broadband absorbing layer.
- Near 12 GHz, instead of displaying two independent resonance peaks, FSS3 exhibits interference effects, resulting in a single, shifted resonance dip.
- At higher frequencies (>15 GHz), the interaction between the two structures creates a smoother roll-off rather than additive deep nulls.

This behavior is caused by EM coupling and mutual interaction between the layers in FSS3:

- (1) Near-field coupling: The fields generated by one FSS layer alter the effective boundary conditions of the other, modifying its resonance characteristics.
- (2) Multiple reflection paths: Internal reflections between layers create constructive and destructive interference, yielding a nonlinear combined response.
- (3) Impedance mismatch effects: The overall impedance of the composite system differs from that of the individual FSS layers, leading to emergent behavior that cannot be predicted by simple addition.

These findings demonstrate that when designing multilayer FSS absorbers, engineers must account for inter-layer EM interactions rather than relying on the superposition of individual responses. The performance of composite designs like FSS3 must be optimized holistically using full-wave simulations and advanced modeling techniques.

This principle is especially critical for applications such as:

- RCS reduction, where precise control over multi-band absorption is required.
- Broadband EMI shielding, which demands seamless integration of narrowband and wideband elements.

Reconfigurable meta-surfaces, where dynamic tuning relies on predictable layer interactions.

Figure 15 compares the simulated reflection coefficients (S_{11}) of three different metamaterial (MTM) configurations – MTM1 (red curve), MTM2 (green curve), and MTM3 (blue curve) – across the frequency range of 0–25 GHz. The figure illustrates the complex EM interactions inherent in metamaterial-based absorbers and demonstrates why the superposition of individual responses is invalid, as indicated

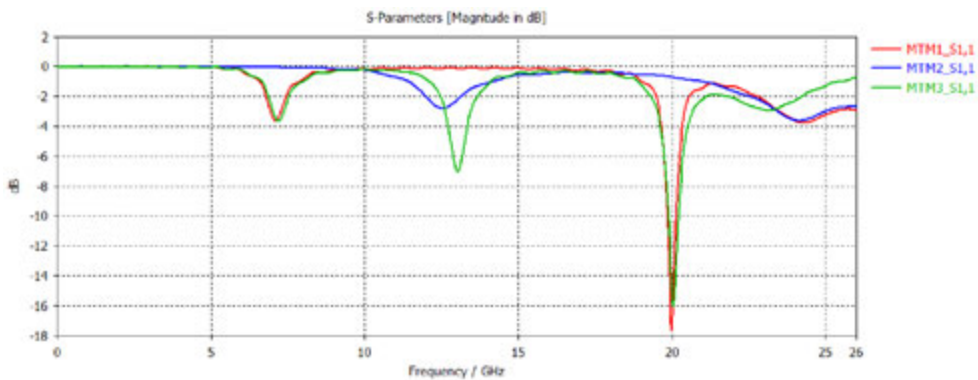


Figure 15.
 Metamaterial forms.

in the annotation. The results show that each metamaterial structure produces distinct and sharp resonance dips at different frequencies:

- MTM1 (red) exhibits a strong resonance at approximately 20 GHz, indicative of highly localized resonant behavior.
- MTM2 (green) has a pronounced resonance near 8 GHz, reflecting its tailored geometry and unit-cell properties.
- MTM3 (blue) demonstrates moderate broadband characteristics, with resonances spread across the mid- to high-frequency range.

When the three configurations are combined into a composite structure, the resulting response cannot be predicted by the simple summation of the individual curves. Instead of displaying three independent resonances, the combined response shows nonlinear coupling effects, including resonance shifting, amplitude suppression, and the emergence of new resonant features. This phenomenon occurs because metamaterials exhibit strong near-field coupling and EM interaction between neighboring unit cells and layers:

- (1) Field coupling: Localized fields in one metamaterial layer alter the boundary conditions of adjacent layers, modifying their natural resonance frequencies.
- (2) Multiple reflection paths: Constructive and destructive interference within the stacked structure lead to new emergent behaviors.

Effective medium variation: The combined structure behaves as a new homogenized medium with different effective permittivity (ϵ_r) and permeability (μ_r) compared to the sum of its parts.

The findings confirm that MMA design must be approached holistically, with the entire composite structure optimized using full-wave simulations rather than relying on the linear addition of individual responses. This is especially important for:

- Stealth technology, where precise control of RCS across multiple frequency bands is required.
- Broadband EMI shielding, where multi-resonant behavior must be carefully engineered for stability and wide coverage.
- Reconfigurable or tunable metamaterials, where dynamic control of resonance is achieved through complex layer interactions.

It can be concluded from the analysis in **Figure 15**, which demonstrates that superposition is invalid for metamaterial forms, emphasizing the need for advanced modeling and design techniques to fully exploit the unique capabilities of metamaterials. Future work should explore ML-based optimization and parametric studies to systematically account for these nonlinear interactions and accelerate the development of high-performance broadband absorbers.

3.8 Superposition

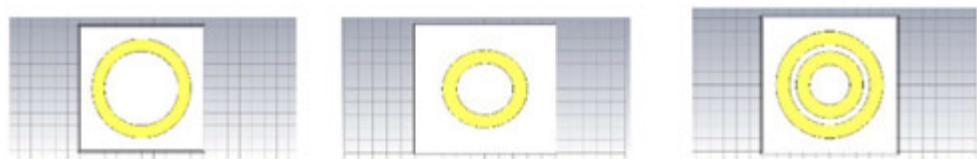
This section investigates the validity of the superposition principle. Two circular patches, separate and coplanar with a ground plane, were calculated for S_{11} .

Figure 16 compares the EM responses of two different configurations of planar patches: (1) the geometric structures of the patches, (2) their behavior as MMAs, and (3) their performance as FSS without a ground plane. This comparison highlights the fundamental differences in how patch interactions manifest in these two distinct configurations and provides critical insights into design considerations for advanced EM absorbers.

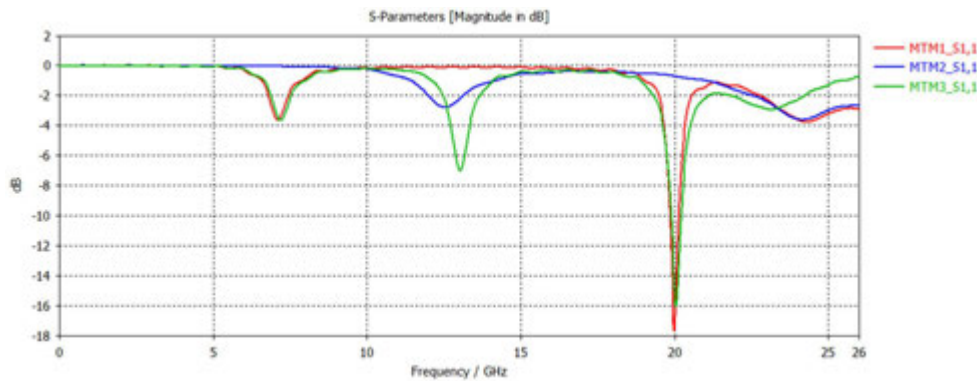
In **Figure 16(b)**, the three patches configured as metamaterials exhibit localized and highly resonant behaviors, with distinct, sharp dips in the reflection coefficient (S_{11}). The red, green, and blue curves correspond to three unique patch designs. The sharp resonances result from near-field coupling and localized current distributions, which enable efficient energy trapping and dissipation at specific frequencies. Importantly, the combined behavior of the patches cannot be predicted by linear superposition, as the close proximity of the metamaterial elements introduces strong EM interactions, shifting resonance frequencies and amplitudes. This validates the nonlinear nature of metamaterial-based designs, where the collective behavior differs fundamentally from the sum of individual responses.

Conversely, **Figure 16(c)** shows the same patches configured as FSS elements without a ground plane. In this case, the responses are much smoother and more broadband, with reflection coefficients demonstrating gradual attenuation rather than sharp resonance nulls. The absence of a reflective backing allows EM waves to partially transmit through the structure, resulting in weaker coupling and more predictable performance. Here, the behavior of the combined structure is closer to linear superposition, though still influenced by limited inter-element interactions. This makes FSS designs suitable for applications such as broadband EMI shielding or moderate absorption, where partial transmission is desirable.

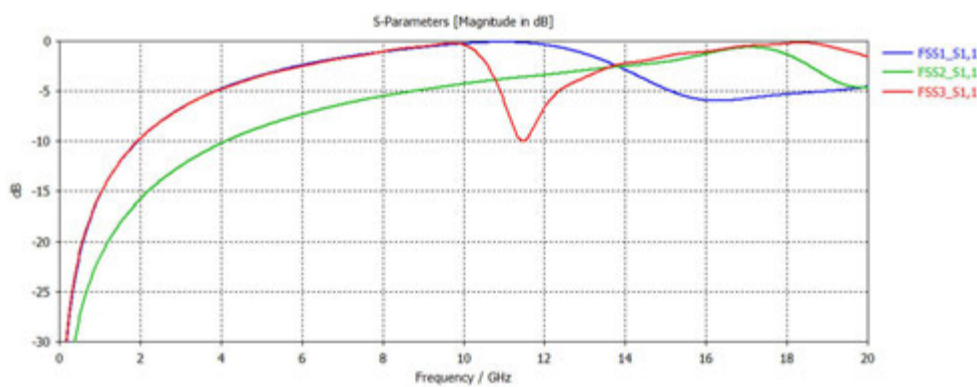
The structural configurations are: (1) Metamaterial configurations, as shown in **Figure 16(b)**: These exhibit nonlinear, strongly coupled responses due to near-field interactions. They are ideal for stealth and RCS reduction applications, where sharp



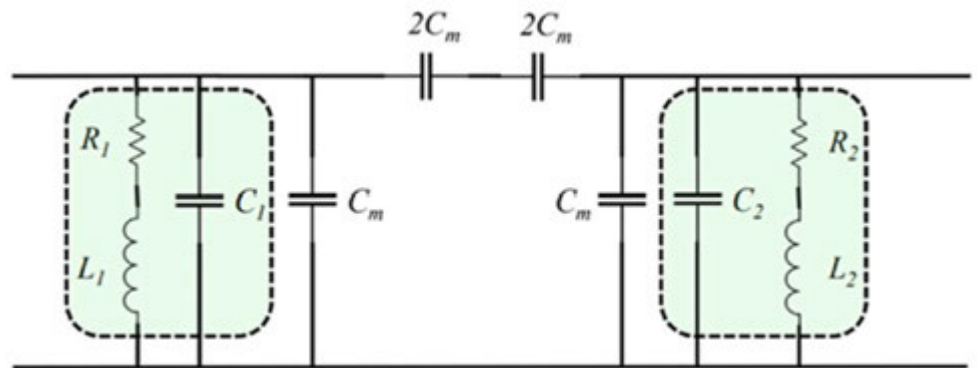
(a) Various structures



(b) Separate and planar patches of metamaterial



(c) Separate and planar patches without a ground plane (FSS)



(d) Equivalent circuit [19]

Figure 16.
Superposition of two different patches, in separate and planar forms.

and targeted absorption at specific frequencies is critical, and they require holistic optimization rather than a linear summation of individual patch behaviors. (2) FSS configurations, as shown in **Figure 16(c)**: These display weaker coupling with smoother, broadband responses. They are suitable for EMI shielding and applications requiring partial transparency, and their combined response approximates additive behavior but still requires careful modeling for accurate prediction.

It is concluded from **Figure 16** that it demonstrates the profound difference in interaction mechanisms between metamaterial and FSS structures. While FSS designs operate under relatively linear and predictable principles, metamaterials exhibit complex, nonlinear superposition effects, requiring advanced full-wave simulations and precise geometric tuning. These findings are crucial for designing next-generation EM absorbers that balance broadband performance, peak absorption, and structural compactness. Future work should focus on hybrid approaches that merge the sharp resonant characteristics of metamaterials with the broadband stability of FSS configurations, potentially leveraging tunable or reconfigurable geometries to dynamically adjust system performance.

The equivalent circuit, which helps clarify the operating principle of the design, is presented in **Figure 16(d)**. As illustrated, each ring can be represented by a parallel LC branch, with each branch corresponding to a separate resonant frequency. The resistive elements account for the losses in the structure, and their values can be adjusted to control the absorption level.

3.9 Reciprocity

In most passive EM devices, the principle of reciprocity applies. This principle states that the input and output ports of a system can be interchanged without affecting its performance, meaning the reflection, transmission, and loss characteristics remain identical. In other words, a reciprocal device exhibits symmetrical behavior regardless of the direction of EM wave propagation.

For radio frequency (RF) absorbers, this principle can be analyzed using a two-port network model, as shown in **Figure 17**:

- Port 1 acts as the input terminal, where the incident wave enters.
- Port 2 serves as the output terminal, where the transmitted wave exits.

The system's performance is characterized by its scattering parameters (S-parameters), which relate incident and reflected waves at both ports:

$$\begin{bmatrix} b1 \\ b2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a1 \\ a2 \end{bmatrix}$$



Figure 17.
Two-port absorber.

For a reciprocal system, the following conditions must be satisfied [19]:

- $S_{11} = S_{22}$ (equal reflection coefficients at both ports)
- $S_{12} = S_{21}$ (equal transmission coefficients in both directions)

To evaluate reciprocity, a two-layer absorber structure consisting of dielectric rubber sheets and FSS was simulated without a ground plane, as illustrated in **Figure 17**.

The simulation results of the two-port network are shown in **Figure 18**.

- The red curve represents the reflection coefficient (S_{11}), indicating the amount of energy reflected back toward the source.
- The green curve represents the transmission coefficient (S_{21}), corresponding to the wave transmitted through the structure.

The plots confirm that $S_{11} = S_{22}$ and $S_{12} = S_{21}$, validating that the cascaded multilayer absorber satisfies the reciprocity condition. This demonstrates that interchanging the source and load ports yields identical EM performance.

The reciprocity behavior arises because of the structure:

- (1) Lacks active components, such as amplifiers or isolators, which would otherwise break reciprocity.
- (2) Exhibits uniform and symmetric material properties, ensuring that EM wave propagation is consistent in both directions.
- (3) Has no ground plane, allowing bidirectional transmission without asymmetrical reflections.

In practical terms, this means the absorber behaves identically whether the EM wave impinges from the front or back, making it suitable for applications requiring bidirectional EMI shielding or symmetrical stealth performance.

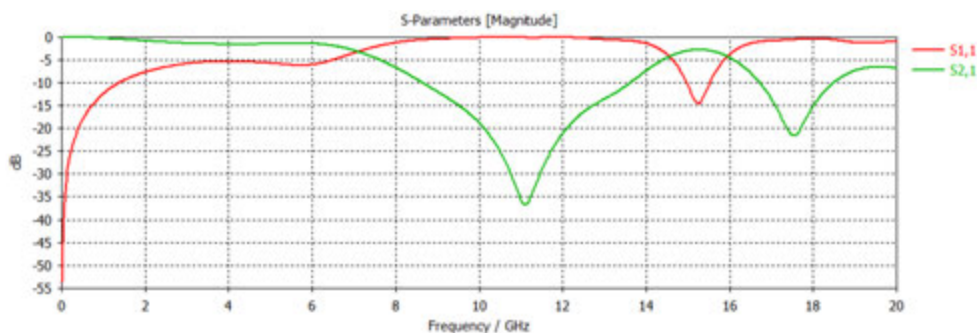


Figure 18.

The multilayer rubber sheets and multilayer printed patches without ground planes (FSS) satisfy the reciprocity principle.

It is concluded from the results that it is confirmed that multilayer rubber absorbers and multilayer printed patches without ground planes inherently satisfy the reciprocity principle. This behavior is essential for ensuring reliable and predictable performance in applications where signals may originate from multiple directions, such as RCS reduction, broadband EMI mitigation, and dual-function communication–absorption systems.

3.10 Complementary and polarization-sensitive designs

Complementary structures are a fundamental design approach in FSS and metamaterials, based on Babinet's principle. According to this principle, the EM response of a structure and its geometric complement are closely related, with the fields interchanged between the aperture and metallic regions. This concept enables engineers to tailor reflection, transmission, and absorption characteristics by pairing complementary patterns, providing enhanced control over bandwidth and polarization behavior.

Figure 19(a) illustrates the geometric configuration of a standard FSS patch and its complementary structure. The complementary design replaces metallic regions with apertures, resulting in inverted EM behavior.

- Red curve (FSS): Represents the standard FSS configuration, showing a strong resonance near 9–12 GHz, where transmission is suppressed and reflection peaks due to impedance mismatch.
- Black curve (Complementary FSS): Displays the complementary response, demonstrating minimal reflection in the same frequency region while offering broader passband characteristics.

It is concluded that, from **Figure 19(b)**, the results confirm that standard and complementary FSS elements exhibit inverse frequency responses, as expected from Babinet's principle.

In **Figure 19(c)**, MMAs and their complements are compared to evaluate how the complementarity principle applies to advanced resonant designs.

- Blue curve (metamaterial – MTM): Exhibits multiple deep and narrow resonances, with high selectivity across the 8–24 GHz range. These features are due to localized EM field interactions within the periodic structure.
- Red curve (complementary metamaterial – CMTM): Shows an inverted response compared to the original MTM, where transmission peaks and reflection nulls occur at frequencies complementary to those of the original design.

As a conclusion, FSS satisfies Babinet's principle, though only partially for metamaterials. By carefully selecting geometric configurations, designers can create hybrid structures with enhanced bandwidth and controllable anisotropic properties.

Polarization sensitivity is a critical factor in the design of FSS and MMAs. The response of a structure to incident EM waves depends strongly on its geometrical symmetry relative to the polarization orientation of the incident field.

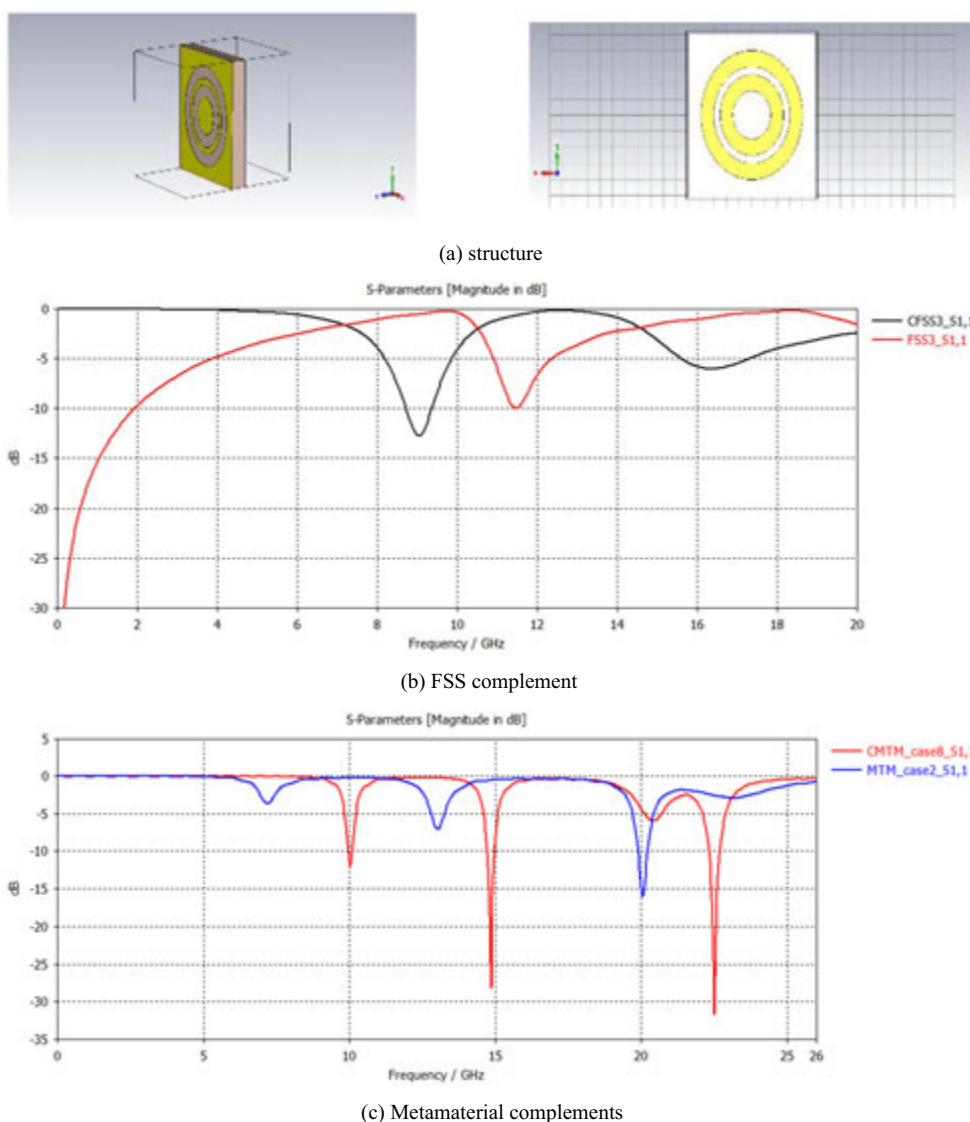


Figure 19.
 The complement patches comparison.

Figure 20 shows a cross-shaped patch that exhibits polarization sensitivity, as evidenced by the difference in its reflection response for different polarization orientations:

- When illuminated with vertical or horizontal polarization, the structure produces distinct resonance frequencies due to the directional alignment of the incident electric field with the patch geometry.
- When the incident field is rotated to a 45° inclined polarization, the coupling mechanism changes significantly, causing noticeable variations in the S-parameter response.

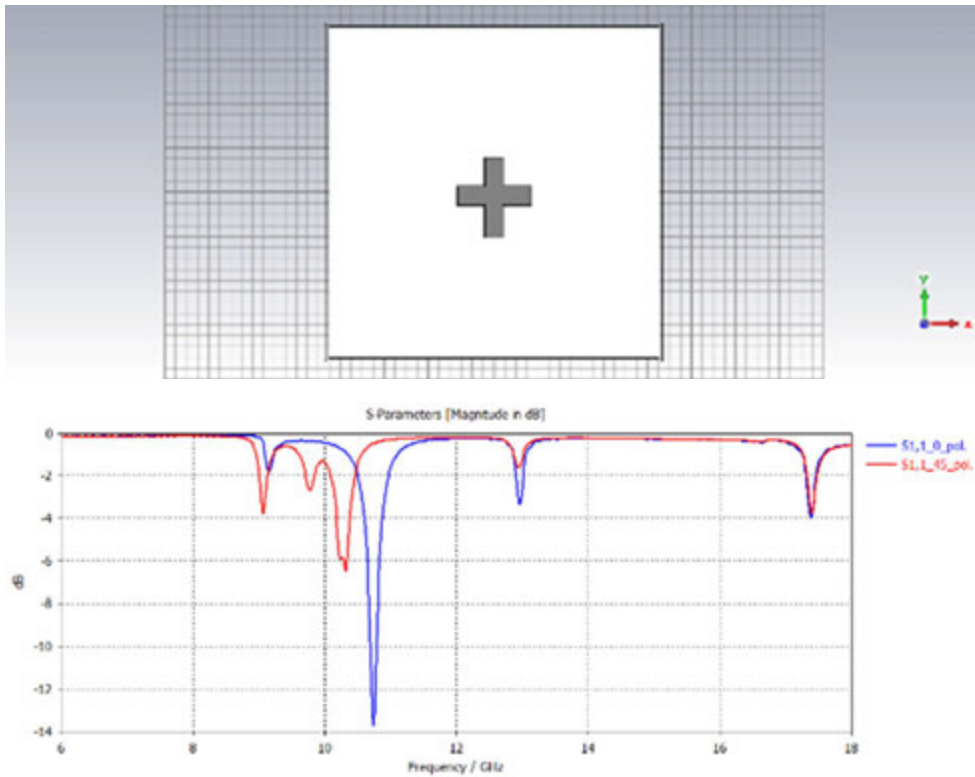


Figure 20.
Polarization sensitivity patch.

The plot clearly illustrates this behavior, where separate curves correspond to different polarization angles. This sensitivity arises because the cross-shaped patch lacks rotational symmetry, making it highly dependent on the incident field's orientation. Such designs are particularly useful in applications requiring polarization-selective filtering, but they may be undesirable in broadband absorbers where consistent performance is needed across all polarization states.

In contrast, **Figure 21** presents a circular patch with unlimited symmetry axes. The circular geometry ensures that the EM response remains invariant under any polarization rotation.

- Regardless of whether the incident field is vertically, horizontally, or diagonally polarized, the induced surface currents and resulting resonance patterns remain identical.
- This results in stable and predictable absorption or reflection characteristics across all polarization states.

Such polarization-insensitive designs are highly desirable for:

- Stealth and RCS reduction, where detection signals may arrive from arbitrary orientations.

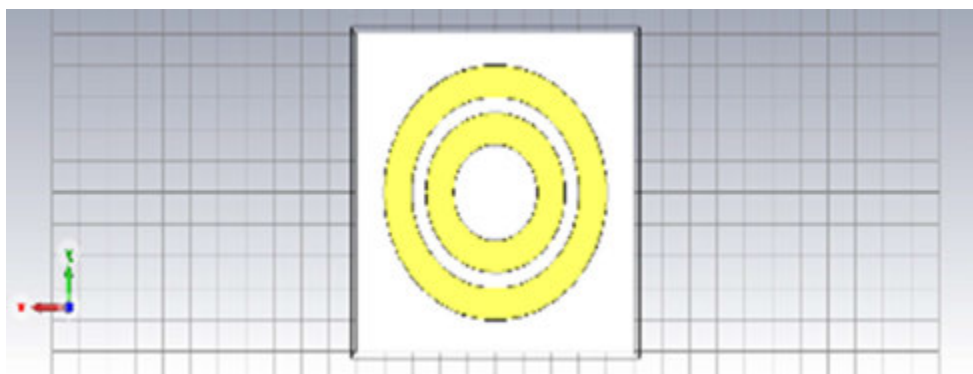


Figure 21.
Circles with unlimited symmetry axes.

- Broadband EMI shielding, ensuring consistent attenuation of interfering signals.
- Wireless communication systems, where unpredictable polarization changes occur due to multipath propagation.

The contrast between **Figures 20** and **21** highlights the importance of geometrical symmetry:

- Limited symmetry (cross-shaped patch): Enables polarization-selective behavior, offering tunability but limiting generality.
- High symmetry (circular patch): Provides robust, polarization-insensitive performance; ideal for wide-angle, multi-directional applications.

By carefully selecting the geometric design of patches, engineers can tailor the polarization sensitivity of metamaterial and FSS structures to meet specific application requirements. Polarization-sensitive patches are well-suited for specialized filtering and sensing, while polarization-insensitive patches, such as those shown in **Figure 21**, are optimal for broadband stealth and shielding applications, where uniform performance across all polarizations is essential.

3.11 Absorption bandwidth boarding techniques

In most practical applications of RF absorbers, the key performance target is to achieve a broad frequency band over which the absorption level exceeds a specified reference threshold (e.g., -3 dB, -5 dB, -10 dB). A common design objective is to maintain the reflection coefficient below -10 dB across a wide operational bandwidth. However, achieving this goal is inherently challenging, as it often leads to an unavoidable increase in the thickness of the dielectric composites forming the absorber structure. This trade-off between bandwidth and thickness remains one of the most significant challenges in the design of broadband, low-profile absorbers.

3.11.1 Metamaterial-based approaches

Recent research efforts in MMAs have focused on innovative strategies to overcome these limitations. Most approaches fall into one or more of the following design methodologies:

- **Root (1): Advanced geometrical design**

Developing complex unit-cell geometries that inherently support broadband characteristics through multi-resonance behavior or gradual impedance-matching techniques.

- **Root (2): Combination of multiple resonators**

Mixing several resonant shapes and carefully tuning their resonant frequencies to lie close to one another can result in broader overall bandwidth, similar to techniques employed in filter design theory. This method requires precise optimization to balance coupling effects while avoiding undesirable resonance suppression.

- **Root (3): Incorporation of resistive elements**

Introducing resistive elements in the gaps between conductors can significantly broaden the bandwidth by dissipating EM energy across a wider frequency range. This technique offers a practical way to improve absorber performance without excessively increasing the structural complexity.

3.11.2 Dielectric absorber strategies

For dielectric absorbers, we investigated bandwidth enhancement using multi-layer structures, as detailed in Sections 3.4 and 3.5. These designs successfully demonstrated moderate bandwidth broadening, though the absorption depth was somewhat limited. A key observation is that simultaneously achieving wide bandwidth and deep absorption nulls remains difficult, as improving one parameter often comes at the expense of the other.

Figure 22 illustrates the simulation trials conducted to broaden the absorption bandwidth of RAM. The three plots represent different design iterations aimed at achieving a reflection coefficient below -10 dB over a wider frequency range.

(1) Initial design:

The baseline RAM structure exhibits two distinct resonance dips at specific frequencies. While these resonances provide strong absorption at narrow bands, the separation between them results in limited overall bandwidth, leaving significant regions with poor performance.

(2) Modified multi-resonant design:

By introducing multiple resonant elements and optimizing their spacing, the absorption peaks move closer together, producing a broader composite response.

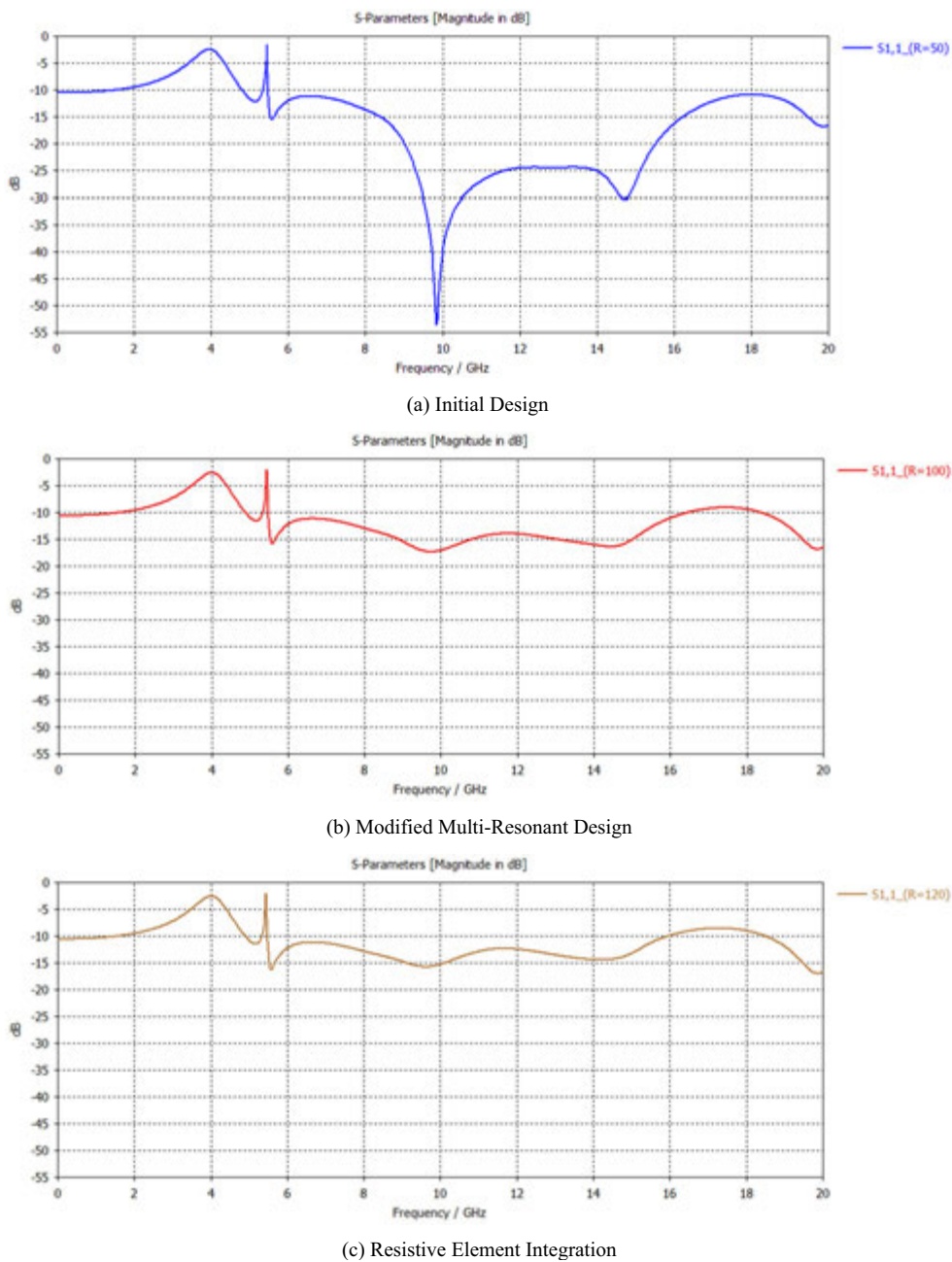


Figure 22.
RAM bandwidth broadening trials.

This approach is like multi-resonant filter design, where overlapping resonances create continuous wideband coverage. However, while the bandwidth improves, the peak absorption depth at individual resonances is slightly reduced.

(3) Resistive element integration:

The final trial incorporates resistive loading into the absorber structure. The resistive elements dissipate EM energy across a wider range of frequencies, further flattening the response and smoothing sharp resonance peaks. This results in a more uniform and broadband absorption profile, demonstrating the effectiveness of combining geometrical optimization with resistive techniques.

These results highlight the inherent trade-offs in broadband RAM design:

- Initial designs with isolated resonances provide high absorption depth but limited bandwidth.
- Multi-resonant approaches increase bandwidth but require precise tuning to avoid destructive interference.
- Resistive loading offers the most stable broadband performance, though at the cost of slightly lower peak absorption.

The progressive improvements shown in **Figure 22** confirm that a hybrid design strategy is essential for next-generation RAM, combining advanced geometry, multi-resonant structures, and resistive elements to meet the stringent requirements of stealth applications, EMI shielding, and wideband RCS reduction.

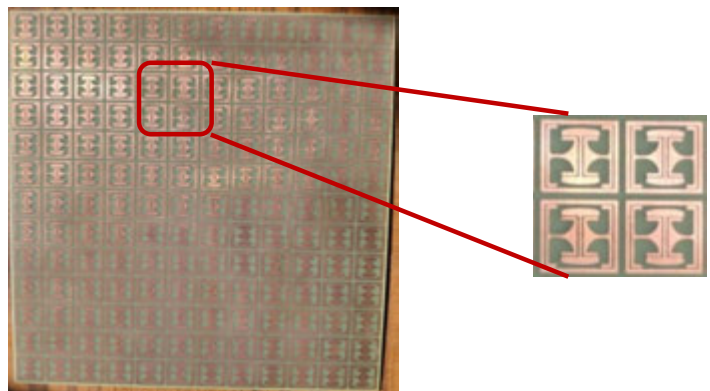
3.12 Absorption sensitivity to incident wave parameters

We mean by incident wave parameters: Normal and oblique, TE and TM incidences, in addition to polarization. The wave frequency is executed in this discussion. In the study by Edries et al. [11], the polarization insensitivity of the absorber design is discussed. The propagation direction of the EM wave remains fixed, while the orientations of the electric field (E) and magnetic field (H) rotate through various polarization angles (φ) under normal incidence, as illustrated in **Figure 23(b)**. The polarization angle φ is swept from 0° to 90° in increments of 15° . The results show that the proposed MMA maintains the same absorptivity for all polarization angles, demonstrating polarization-insensitive behavior. This invariance is attributed to the structural symmetry of the MMA design. Furthermore, the absorption performance of the proposed MMA is examined for different incidence angles θ . As shown in **Figure 23(c)**, the absorber's response is evaluated at incidence angles of 0° , 15° , 30° , and 45° . The results indicate that, at all resonant frequencies, the structure maintains strong absorption exceeding 80% for incidence angles up to 45° .

In the case of dielectric absorber sheets, they behave similarly to radomes; they are highly sensitive to both the angle of incidence and the type of incidence (TE or TM).

The case of metamaterials, as reported in the enormous number of publications, is sensitive to polarization at normal incidence for cases of non-or-limited shape symmetries.

Broadband metamaterial is highly dependent on the angle of incidence and type (TE or TM).



(a) Fabricated prototype of the proposed metamaterial absorber

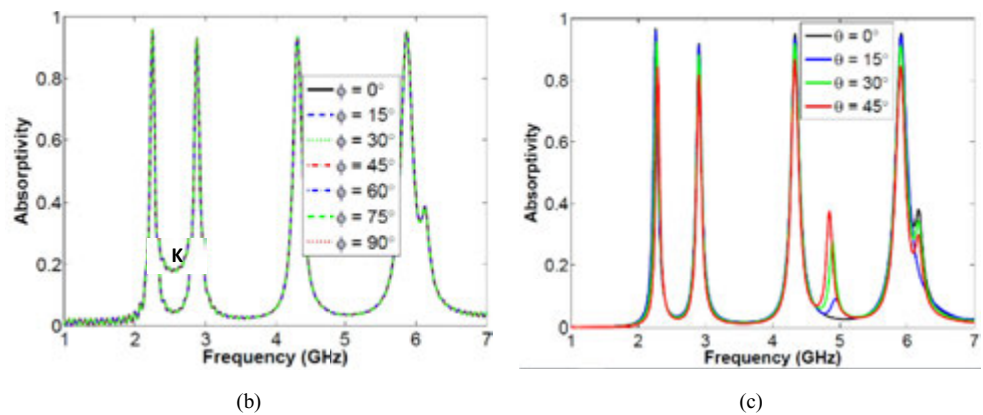


Figure 23.
Fabricated prototype and the absorptivity curves of the metamaterial [11].

3.13 Graphene-based absorbers

Reduced graphene oxide (RGO) is a new and promising EM microwave absorption material, owing to its unique two-dimensional structure and remarkable properties, such as low density, high specific surface area, tunable electrical conductivity, and abundant functional groups. The tunability and frequency-independent surface impedance of graphene at the microwave band have been theoretically investigated and proven feasible as an FSS. However, the fabrication process of graphene FSS-based composites is complex, time-consuming, and expensive. Additionally, FSS is usually on the surface of the composite and directly exposed to the air, which increases the possibility of oxidation and degradation of microwave absorption. An example of the measured performance of a graphene-based absorber, as reported in the study by Rahman et al. [12], is shown in **Figure 24**.

Table 1 presents a concise validation of the key design principles investigated in this chapter, linking each skill to its corresponding section and performance impact. It highlights how scaling, material parameters, and structural configurations influence frequency response, absorption level, and bandwidth. Multilayer and hybrid designs are shown to enhance loss and bandwidth, while the applicability of

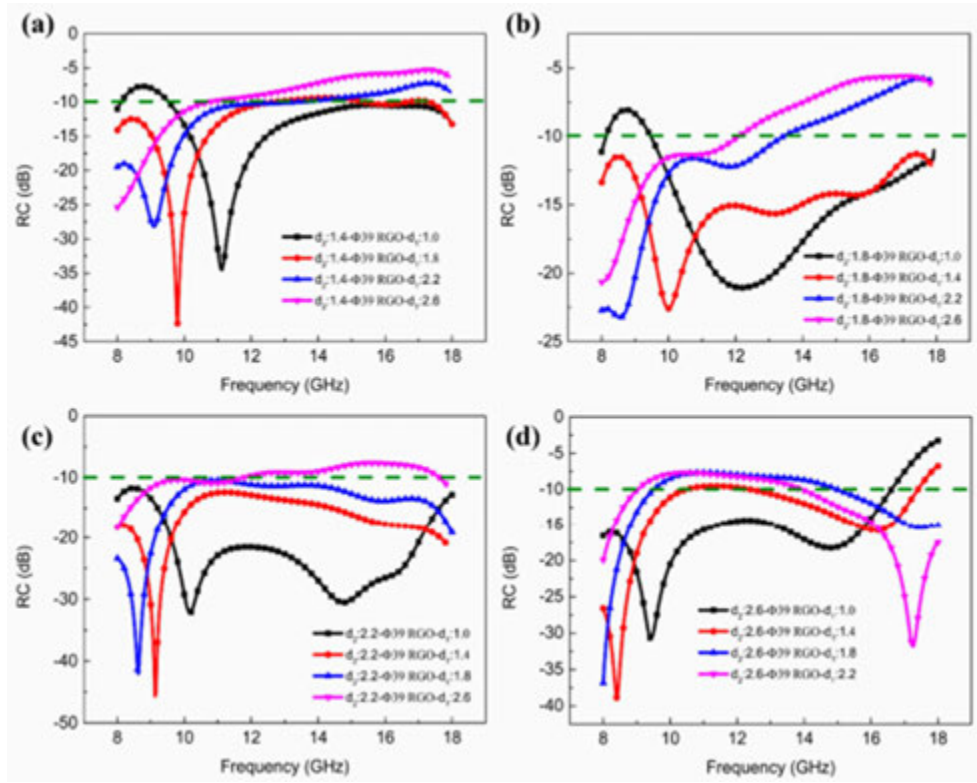


Figure 24.
A graphene absorber [20].

concepts such as superposition, reciprocity, and polarization sensitivity is clarified for different RAM architectures.

4. Conclusion

This chapter has presented a comprehensive framework for the design, analysis, and optimization of RAMs, addressing both theoretical foundations and practical implementation strategies. By exploring a wide range of techniques – including scaling methodologies, impedance matching theory, multi-layer dielectric laminates, and metamaterial-based absorbers – the study demonstrated how absorber performance can be tailored to meet specific operational requirements, such as stealth technology, EMI mitigation, and broadband RCS reduction. Key findings highlight that material properties, including permittivity (ϵ_r), permeability (μ_r), and loss tangents, play a pivotal role in determining absorption depth, bandwidth, and frequency response. Multilayer designs with stepwise impedance matching significantly enhance broadband performance, while metamaterial and FSS configurations enable precise control of resonance characteristics. Additionally, the integration of resistive elements and hybrid geometries was shown to be highly effective in overcoming the classical trade-off between absorber thickness and

Skill	Section	Validation evaluation
Scaling	3.1	For frequency performance shifting: • Thickness scaling is valid with dielectric sheets. • For FSS, patch dimensions are scalable. • For metamaterial patch scaling, it is mostly valid.
μ_r and ϵ_r variations	3.2.1	Causes frequency performance shifting of dielectric sheets, such as thickness scaling.
Loss tangents $\tan\delta_e - \tan\delta_m$	3.2.2	Causes frequency performance shifting of dielectric sheets, with some performance shape variations.
μ_r equal ϵ_r absorbers	3.3	These dielectric sheets have a decreasing slash-type response, with a slope dependent on permittivity, = permeability, and loss tangent values.
Cascaded dielectric laminates	3.4	Cascading different dielectric sheets with varying input parameters increases the structural loss and enhances bandwidth.
Cascaded μ_r equal ϵ_r sheets	3.5	Improves the structural loss and bandwidth enhancement.
Combined FSS and rubber absorbers	3.6	Stacking a meta material with a lossy dielectric sheet backed by a ground plane improves the loss.
Planner and stacked patches	3.7	Stacking FSS (ring/square shape or others) in multilayers or on one plane creates a variety of frequency responses that might meet the designer's objectives. As for metamaterial printing, all patches in one plane generate nearly the sum of the performance of each individual patch.
Superposition	3.8	It is valid for metamaterial as it produces an integrated performance, and invalid for FSS patches.
Reciprocity	3.9	The method is valid with dielectric sheets and multi-layer FSS (without a ground plane).
Complementary and polarization-sensitive designs	3.10	FSS compliments have inverted responses, while metamaterial is mostly invalid. 360-degree axial unsymmetrical patch cuts are polarization-dependent.
Absorption bandwidth boarding	3.11	See comments in this section.
Graphene-based absorbers	3.13	See comments in this section.

Table 1.
Skills validation summary.

operational bandwidth. The chapter also established that nonlinear coupling effects in metamaterial-based absorbers require holistic optimization, as the simple superposition of individual responses is invalid. Conversely, FSS structures exhibit more predictable, quasi-linear behavior, making them well-suited for broadband EMI shielding. Furthermore, concepts such as reciprocity, polarization sensitivity, and complementary design principles were addressed, providing deeper insights into advanced absorber behavior and design constraints. Emerging technologies, such as graphene-based absorbers and AI-driven optimization techniques, were identified as promising avenues for next-generation RAM development. These innovations offer the potential for lightweight, reconfigurable, and adaptive absorbers capable of

meeting the evolving demands of 5 G/6 G communications, autonomous systems, and multi-functional stealth platforms.

RAM types of rubber sheets, also known as Salisbury screens and metamaterials, each have a distinguished basic property that makes each type suit specific applications. Rubber sheets can handle high-power EM radiation while material mass and shape are tolerated. Synthesizing specifically designed material parameters – permittivity, permeability, and loss tangent – is a crucial job. For printed patch RAM, the EM power handling levels are bounded at low values. Manufacturing a specifically selected design can be easily achieved.

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Bandgap Structure in the Terahertz Region for Reflecting Electromagnetic Waves: A Theoretical Approach Considering Dispersion

Jesús Valenzuela-Sau, Raúl García-Llamas and Jorge Gaspar-Armenta

Abstract

In this chapter, a bandgap structure in the terahertz region is studied, considering the dependence of the dielectric constant value on the frequency of electromagnetic waves. First, the electromagnetic equation for an inhomogeneous medium is derived from Maxwell's equations. Then, the basic theory of photonic crystals (PCs) is presented, emphasizing their geometry and how it is defined. The relevance of the dielectric constant in a PC is addressed next, showing the dispersive nature of alumina with experimental data of the dielectric constant, which exhibited values ranging from very large to very small, even showing vacuum-related values. Later, the general eigenvalue equation for any PC is obtained through the application of the plane wave expansion method. As a study case, the in-plane and out-of-plane band structures of a dispersive alumina two-dimensional PC are presented in the results section, commenting on expected but also unusual behaviors in the bands due to the contrasting values of the dielectric constant. Two absolute band gaps appear between approximately 12.6 and 17 THz. Lastly, some conclusions are given.

Keywords: dielectric constant, frequency, electromagnetic equation, band structure, bandgap, plane wave expansion method, terahertz

1. Introduction

The dielectric constant plays a crucial role in the study and development of technologies in which electromagnetic waves interact with matter as part of their operation. Well-known examples of this include optical fibers, lenses, radars, Xray machines, just to name a few. More recently, metamaterials and metasurfaces have been demonstrated to possess innovative properties for possible applications across a wide range of the electromagnetic spectra: for instance, in RF communications

[1], in the development of 5G [2], sub-6G [3], and ultrawide-band monopole [4] antennas, in band gap filters for electromagnetic interference [5], and even in cloaking [6]. Metamaterials are not restricted to electromagnetics; possible applications have also been found in fields such as acoustics [7], plasmonics [8], and mechanics [9].

Photonic crystals (PCs) can be seen as predecessors of metamaterials. They both influence electromagnetic waves, although in different ways. PCs have structures of a size close to the wavelength, whereas in metamaterials, it is the opposite; the structures are much smaller than the wavelength. In both cases, the dielectric constant not only plays a part, but it is also a fundamental building block, not only because of its numerical value but also because this value will certainly change for different frequencies in the electromagnetic spectrum. Let's recall that this value tells us how much a medium can be polarized.

PCs emerged in the 1990s as a promising subject in optics, given their main property: forbidding the propagation of light through them. We can think of them simply as structures where the dielectric constant varies periodically in space. They have a pattern, called unit cell (UC), that is repeated, creating a periodicity which can be of three types: one-, two- and three-dimensional, where the UC is a length, an area, and a volume, respectively.

In principle, PCs can be made of any material, which we can separate into two main types: electrically conductive and non-conductive; the latter better known as dielectrics. The materials used in a PC, usually vacuum and a dielectric, are described electromagnetically by their dielectric constants. A contrast of a few units in the relative dielectric constant between the constituents is enough for the band gaps to appear. Regarding this, a major issue to deal with is the dispersion phenomenon, where the refractive index of materials is a function of the frequency of light. This dependency can lead to major changes in the dielectric constant. For one frequency, a band gap might exist, but for another frequency, the PC could be "seen" as vacuum.

What about the size of a PC? Well, it depends. But let's consider a better question: What is the size of the UC of a PC? Again, it depends. The physical size of the UC must be approximately of the order of magnitude of the light's wavelength we want to "forbid." A PC with a photonic band gap (PBG) within the visible would have a UC size of a few hundred nanometers, while a PBG in the microwave would require a UC of millimeters. To give an answer to the first question, in theory, a PC should be infinite in size, with its UC repeating itself, occupying the whole space. In practice, PC effects can be achieved with as few as 10 UCs.

The key phenomenon that allows the existence of a PBG is called destructive interference. This happens when the waves within a PC cancel themselves as a direct consequence of the periodicity and the contrast of the dielectric constants of the materials that form the PC. Destructive interference effects are more likely when the propagation of light is exclusively along the periodicity of a PC. For instance, in two-dimensional PCs, when the propagation is in the plane of periodicity, an absolute band gap [10] can appear, meaning that the electric and the magnetic fields are both forbidden. But when the propagation is out of the plane, the periodicity of the PC fades, and destructive interference is less likely to occur.

Some promising applications for PCs are in the transition from electronics to photonics, where the intention is to emulate the operation of electronic devices such as the diode [11] and the transistor [12] with all-optical equivalents. A current application that takes advantage of band gaps is the use of PC-based optic fibers, where a core surrounded by a PC transmits light for different purposes [13].

In this chapter, a forbidden bandgap structure in the terahertz region is studied, considering dispersion. Said structure is a PC in which the value of the dielectric constant undergoes drastic changes over a relatively large frequency spectrum. In the theory section, we will present a basic approach to the plane wave expansion method, first obtaining the electromagnetic wave equation, then defining mathematical functions for the dielectric constant and the electromagnetic field, and later deriving the general equation for one-, two-, and three-dimensional PCs. Additionally, a discussion will be provided about the relevance of dispersion in dielectrics, particularly in aluminum oxide (alumina). In the results section, the in-plane and out-of-plane band structures of an array of circular cross-sectional dispersive alumina rods ordered in a honeycomb lattice are presented, along with some discussion about the results. Lastly, some conclusions will be provided.

2. Theory

Starting from the macroscopic Maxwell equations and considering harmonic behavior in time for the field, we will obtain the electromagnetic wave equation for a PC. The basic theory for PCs will be presented, emphasizing the vectors that define their type of periodicity and lattice. Later, a discussion about the relevance of dispersion in dielectrics is given, presenting experimental data. Definitions for the dielectric function and the field will be proposed and substituted into the wave equation to derive the general eigenvalue equation to obtain the electromagnetic modes of a PC.

2.1 The wave equation

The Maxwell equations for a macroscopic medium are as follows:

$$\vec{\nabla} \cdot \vec{D} = \rho, \quad (1)$$

$$\vec{\nabla} \cdot \vec{B} = 0, \quad (2)$$

$$\vec{\nabla} \times \vec{E} = -\frac{\partial \vec{B}}{\partial t}, \quad (3)$$

and

$$\vec{\nabla} \times \vec{H} = \frac{\partial \vec{D}}{\partial t} + \vec{J}. \quad (4)$$

These equations describe the relation between the electric and magnetic fields, which are generated for charges densities ρ , and current densities $\vec{J}$. $\vec{D}$ is the electric displacement, $\vec{B}$ is the magnetic flux density, $\vec{E}$ is the electric field, and $\vec{H}$ is the magnetic field.

The presence of an electric field in an isotropic dielectric medium of linear response has the effect described by the equation:

$$\vec{D} = \epsilon \vec{E}, \quad (5)$$

where ϵ is the electric permittivity of the medium. In vacuum $\epsilon = \epsilon_0$. Similarly, the effect of a magnetic lux density in a medium with magnetic permeability μ is described by the equation:

$$\vec{B} = \mu \vec{H}. \quad (6)$$

Magnetic permeability is the grade of magnetization that a material has in the presence of a magnetic field. In this analysis, we will consider no magnetic media; $\mu = \mu_0$, no charges; $\rho = 0$, and no currents; $\vec{J} = 0$, no absorbing media, harmonic behavior in time for the fields, and an electric permittivity, that we will call dielectric constant, in function of space and frequency; $\epsilon = \epsilon(\vec{r}, \omega)$. Thus, Maxwell's equations became:

$$\vec{\nabla} \cdot (\vec{\epsilon}(\vec{r}, \omega) \vec{E}(\vec{r})) = 0, \quad (7)$$

$$\vec{\nabla} \cdot \vec{B}(\vec{r}) = 0, \quad (8)$$

$$\vec{\nabla} \times \vec{E}(\vec{r}) = -i\omega \vec{B}(\vec{r}), \quad (9)$$

and

$$\vec{\nabla} \times \vec{H}(\vec{r}) = -i\omega \epsilon(\vec{r}, \omega) \vec{E}(\vec{r}); \quad (10)$$

where the dependency of $\vec{E}$ and $\vec{B}$ on ω is implicit.

To obtain the wave equation, first we solve for $-i\omega \vec{E}(\vec{r})$ from Eq. (10), and apply the curl of the resulting equation to obtain:

$$\vec{\nabla} \times \left[\frac{1}{\epsilon(\vec{r}, \omega)} \vec{\nabla} \times \vec{H}(\vec{r}) \right] = \vec{\nabla} \times [-i\omega \vec{E}(\vec{r})]. \quad (11)$$

Equation (9) is substituted in Eq. (11) and $\vec{B}(\vec{r})$ is written in terms of $\vec{H}(\vec{r})$, getting:

$$\vec{\nabla} \times \left[\frac{1}{\epsilon(\vec{r}, \omega)} \vec{\nabla} \times \vec{H}(\vec{r}) \right] = \omega^2 \mu_0 \vec{H}(\vec{r}). \quad (12)$$

Finally, multiplying Eq. (12) by ϵ_0 and considering that $c = (\mu_0 \epsilon_0)^{-1/2}$, the electromagnetic wave equation is:

$$\vec{\nabla} \times \left[\frac{1}{\epsilon(\vec{r}, \omega)} \vec{\nabla} \times \vec{H}(\vec{r}) \right] = \frac{\omega^2}{c^2} \vec{H}(\vec{r}). \quad (13)$$

Now, $\epsilon(\vec{r}, \omega)$ will be the relative dielectric constant.

Equation (13) is the wave equation for the magnetic field for an inhomogeneous media. For the case of a homogeneous media with electric permittivity ϵ and magnetic permeability μ , the equation

$$\vec{\nabla}^2 \vec{H}(\vec{r}) + \omega^2 \mu \epsilon \vec{H}(\vec{r}) = 0 \quad (14)$$

is obtained, which is known as the Helmholtz equation.

2.2 The medium

A PC, which is an inhomogeneous medium, can have one-, two-, or three-dimensional periodicity. The UC is the basic unit that is repeated in the whole space

to form the PC, and it is defined with a length, an area, and a volume, for 1D, 2D, and 3D PCs, respectively. The UC is composed by media of different dielectric constants, usually two, one with a “high” value ϵ_c , and the other with a “low” value ϵ_e . The UC can form different lattices, called Bravais lattices. For the three-dimensional case, there are 14 lattices, for the two-dimensional five, and the one-dimensional just one. **Figure 1** shows cuts of three examples of PCs for the three types of periodicities.

The center of each UC in the PCs of **Figure 1** is given by:

$$\vec{R}_{m,n,o} = \vec{a}_1 m + \vec{a}_2 n + \vec{a}_3 o, \quad (15)$$

with m , n and o integers numbers, and where $\vec{a}_1$, $\vec{a}_2$, and $\vec{a}_3$ are no collinear vectors of the real space. For the 1D PC, n and o are zero. While for the 2D case, o is zero. Vectors $\vec{a}_j$ are expanded as follows:

$$\vec{a}_j = a_{jx}\hat{i} + a_{jy}\hat{j} + a_{jz}\hat{k}, \quad (16)$$

where the values of components a_{ji} depend on the type of lattice. $\hat{i}$, $\hat{j}$, and $\hat{k}$ are orthogonal unit vectors in the directions x , y and z , respectively, and fulfill the property $\hat{i} \times \hat{j} = \hat{k}$.

In the reciprocal space, the centers of the UCs are given by the following:

$$\vec{G}_{p,q,r} = \vec{g}_1 p + \vec{g}_2 q + \vec{g}_3 r, \quad (17)$$

with p , q , and r integer numbers, and where $\vec{g}_1$, $\vec{g}_2$, and $\vec{g}_3$ are the base vectors. As in the real space case, for a 1D PC, q and r are zero, and for a 2D PC, r is zero.

The vectors of the real space and the base vectors of the reciprocal space satisfy the relation:

$$\vec{g}_i \cdot \vec{a}_j = 2\pi\delta_{ij}. \quad (18)$$

For our case study, we will consider a 2D PC with circular cross-section rods ordered in a honeycomb lattice. **Figure 2** shows a PC in a hexagonal lattice that has the same base vectors. The corresponding real-space vectors are as follows:

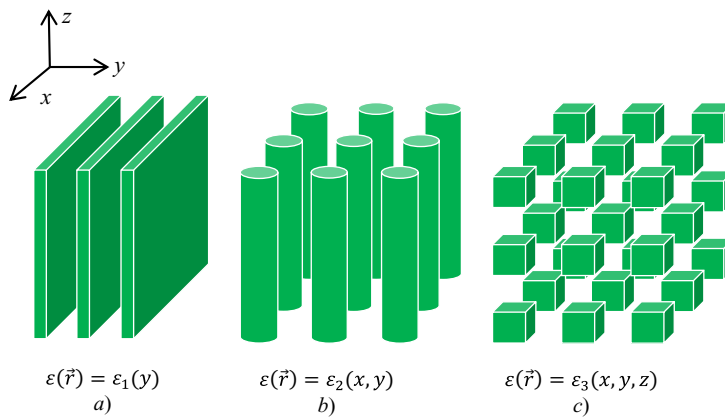


Figure 1. Cut of a (a) one [(b) two-] [(c) three-] dimensional PC. The coordinate axes are indicated in the upper left corner. In (a), the periodicity is along the y -axis, while it is homogeneous in the x and z axes. In (b), the periodicity is in the plane x - y , while it is homogeneous in the z -axis. In (c), the periodicity is in the whole space. In the three cases, the spatial dependence of the dielectric constant is indicated.

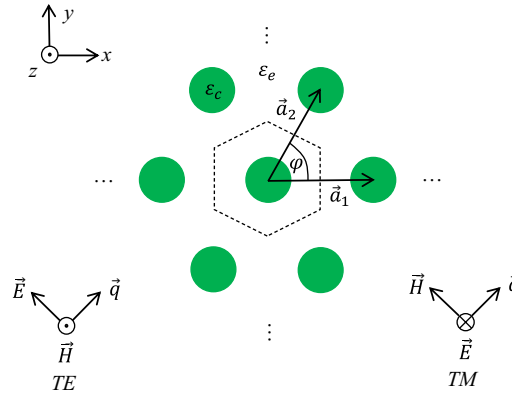


Figure 2.

Parallel cut to the x–y plane of a 2D PC of dielectric cylinders (ϵ_c) of circular cross section ordered in hexagonal lattice in vacuum (ϵ_e). The lattice constant is a . The UC is delimited in discontinuous line in hexagonal shape. Left (right) inset shows in-plane TE (TM) polarization.

$$\vec{a}_1 = a\hat{i} \quad (19)$$

and

$$\vec{a}_2 = a \cos \varphi \hat{i} + a \sin \varphi \hat{j}, \quad (20)$$

with $\varphi = 60^\circ$. Obtained from Eqs. (18–20), the reciprocal space vectors are as follows:

$$\vec{g}_1 = \frac{2\pi}{a \sin \varphi} (\sin \varphi \hat{i} - \cos \varphi \hat{j}) \quad (21)$$

and

$$\vec{g}_2 = \frac{2\pi}{a \sin \varphi} \hat{j}. \quad (22)$$

Until now, we have considered only the geometrical definition of our PC. Let's now talk about the electromagnetic parameter of our PC: the dielectric constant.

As can be seen in **Figure 2**, two dielectric constants are defined, ϵ_c for the cylinders, and ϵ_e for the vacuum matrix space. The contrast between these two parameters has a crucial part in the properties of a PC. For instance, it could be the case that a ratio $\epsilon_c/\epsilon_e = 10$ can cause the existence of a band gap, while a ratio $\epsilon_c/\epsilon_e = 8$ not. Equally important for the formation of a band gap, is the type of lattice and the filling fraction of the PC. The latter being the ratio between the cross-sectional area of the cylinders and the area of the UC. These three factors: the contrast between the dielectric constants, the lattice, and the filling fraction, are the parameters that, in synergy, made possible the formation of destructive interference, giving rise to band gaps. The only way of knowing if a PC has a band gap, is performing the calculations.

Alumina was one of the first materials in the study of PCs, probing the existence of TE and TM band gaps for square lattices [14]. Alumina, or aluminum oxide (Al_2O_3) has several industrial uses given its properties. It is a hard material and has more than three times the melting point of aluminum, so it is used in abrasive applications. As a dielectric, it has very high electric resistivity with low thermal conductivity,

which makes it ideal as an insulator in electric circuits with high voltages and high currents. It is also used in ceramics and refractory applications.

Regarding the use of alumina in PCs, a possibility is to consider its relative dielectric constant between 8 and 10, which can be an acceptable approximation if the frequency of light is low, for instance, in the order of MHz. But if the application is for wavelengths lower than $40\mu\text{m}$, it is necessary to address the dispersive nature of the material.

Dispersion is a universal phenomenon which basically tells us that the refractive index of materials is in function of the frequency of the light; $n = n(\omega)$. This is true for all materials, the which show unique responses. For example, the values of relative dielectric constant of magnesium oxide can be of a few units, or very high, very negative, zero, or even one; behaving like vacuum [15]. For our study case, we will consider alumina rods. **Figure 3** shows the values of dielectric constant calculated from experimental data of the refractive index of alumina [16] from 40 to $10\mu\text{m}$.

In **Figure 3**, instead of presenting the data in function of the wavelength, as is costume, the frequency in THz is used for a clearer reference to the band structures presented in the results section. The curve $\varepsilon(\omega)$ starts with $\varepsilon = 9.8$, and increases gradually to 11.71 with 7.49THz ($\lambda = 40\mu\text{m}$). Keeps growing in a logarithmic fashion to almost 30 with $\sim 11.39\text{THz}$ ($\lambda \cong 26\mu\text{m}$), has a small fluctuation and grows rapidly to ~ 44.5 with $\sim 12.88\text{THz}$ ($\lambda \cong 23\mu\text{m}$). A drastic fall occur from there to 13.49THz ($\lambda = 22.22\mu\text{m}$) dropping to $\varepsilon = 0.585$. Grows again rapidly to reach almost 55 with $\sim 16.78\text{THz}$ ($\lambda \cong 17.8\mu\text{m}$). Another drastic fall gives $\varepsilon = 0.266$ with 17.98THz ($\lambda = 16.67\mu\text{m}$). Finally, a pretty stable behavior with the lowest value of $\varepsilon = 0.014$ is

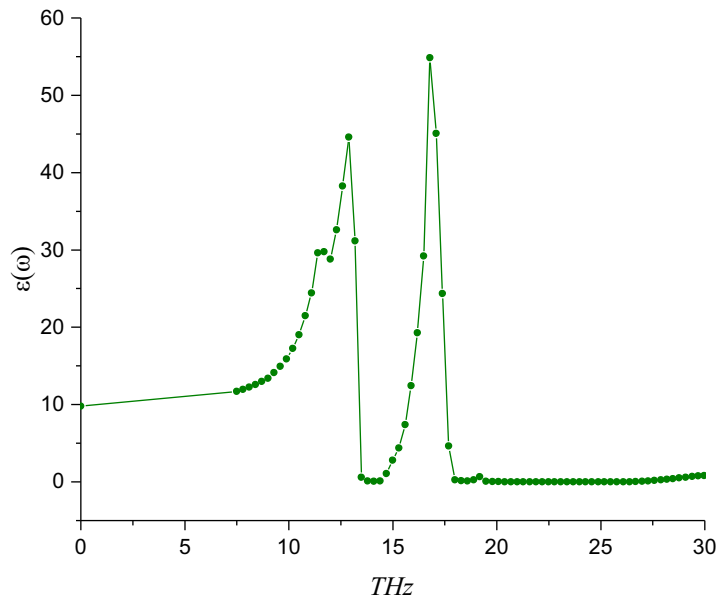


Figure 3. Dielectric constant of alumina (Al_2O_3) in function of the frequency in terahertz. $a = 10\mu\text{m}$ is the period of the PC. $c = 299792458\text{m/s}$ is the speed of light in vacuum. For the lowest value with 1MHz, the low frequency value $\varepsilon = 137$ is used for completeness.

consistent until $\sim 26.67\text{THz} (\lambda \cong 11.2\mu\text{m})$ where ϵ gradually increases to 0.823 with $29.97\text{THz} (\lambda = 10\mu\text{m})$.

The curve in **Figure 3** shows the relevance of considering the dispersive nature of materials, as a change of one unit in the relative dielectric constant can significantly alter a band structure, enabling or disabling the formation of a band gap.

2.3 The eigenvalue equation

An eigenvalue equation is an equation that equates an operator acting on a function (eigenfunction) to the same function multiplied by a constant (eigenvalue). In this case, the operator is a matrix, which has as many associated vectors (eigenvectors) as the order of the matrix. At the same time, each eigenvector has a corresponding eigenvalue.

In order to solve Eq. (13), the plane wave expansion method will be applied, proposing mathematical definitions for the inverse of the dielectric constant and the magnetic field. A general eigenvalue equation will be derived, which can be applied to any type of PC.

2.3.1 The dielectric function

Given the periodicity of the PC, the inverse of the dielectric constant can be expanded in a Fourier series as follows:

$$\frac{1}{\epsilon(\vec{r}, \omega)} = \sum_{\vec{G}} [\epsilon_{\vec{r}, \omega}^{-1}]_{\vec{G}} \exp(i\vec{G} \cdot \vec{r}), \quad (23)$$

with $\vec{r} = x\hat{i} + y\hat{j} + z\hat{k}$ and where $\vec{G}$ is defined by the reciprocal lattice with the vectors from Eqs. (21–22). Eq. 23 involves a sum of sines and cosines that converges to a periodic function. This function corresponds to the dielectric constant of the PC. **Figure 4** shows a representation of how a sum of sines can form a rectangular function. To form a perfect periodic function, the number of summed waves should be infinite. In practice, even 10 can form a decent representation. The coefficients $[\epsilon_{\vec{r}, \omega}^{-1}]_{\vec{G}}$ give the amplitude of the waves. The idea represented in **Figure 4** can be extrapolated to two- and three-dimensional functions.

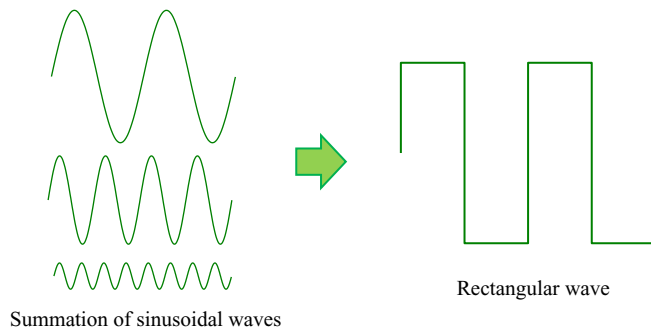


Figure 4.
Representation of the sum of sinusoidal waves that form a rectangular wave.

To obtain the coefficients $[\varepsilon_{\vec{r},\omega}^{-1}]_{\vec{G}}$ from Eq. (23), it is integrated in the limits of the UC, and the property or orthogonality of Fourier functions is used to obtain the following:

$$[\varepsilon_{\vec{r},\omega}^{-1}]_{\vec{G}} = \frac{1}{A} \int_{-UC}^{+UC} \frac{1}{\varepsilon(\vec{r},\omega)} \exp(i\vec{G} \cdot \vec{r}) d\Delta, \quad (24)$$

where A is a length, surface, or volume of the UC and $d\Delta$ is the linear, surface, or volume differential according to the type of PC. $+UC$ and $-UC$ are the limits of the UC. And $\varepsilon(\vec{r},\omega)$ is the relative value of the dielectric constant that takes the values of **Figure 3**. The coefficients $[\varepsilon_{\vec{r},\omega}^{-1}]_{\vec{G}}$ are obtained numerically dividing the UC in small fractions. The UC corresponding to the following calculations is shown in the inset of **Figure 8** delimited in segmented line.

2.3.2 The electromagnetic field

The fields inside periodic media can be expressed as Bloch waves. The Bloch-Floquet theorem states that a Bloch wave is the multiplication of a plane wave by a periodic function with the same periodicity as the medium in which it exists. From this, the magnetic field is defined as follows:

$$\vec{H}(\vec{r}) = \sum_{\vec{G}} \vec{H}_{\vec{G}} \exp(i\vec{q}_{\vec{G}} \cdot \vec{r}), \quad (25)$$

where $\vec{H}_{\vec{G}}$ is a perpendicular vector to $\vec{q}_{\vec{G}} = \vec{q} + \vec{G}$. The propagation wave vector $\vec{q}$ varies in the first Brillouin zone for the in-plane modes, out of it for the out-of-plane ones. The components of $\vec{H}_{\vec{G}}$ depend on the wave polarization and the type of PC. A plane wave is a wave that propagates in only one direction, meaning that its front waves are parallel to each other. **Figure 5** shows a representation of a plane wave.

2.3.3 The general equation

To obtain the eigenvalue equation, the definitions in Eqs. (24 and 25) are substituted into Eq. (13). The procedure is as follows: First, the curl of the field is as follows:

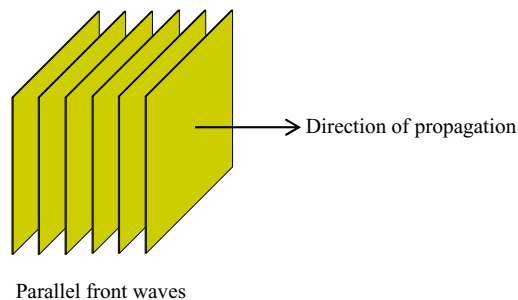


Figure 5.
Representation of a plane wave.

$$\vec{\nabla} \times \vec{H}(\vec{r}) = i \sum_{\vec{G}} (\vec{q}_{\vec{G}} \times \vec{H}_{\vec{G}}) \exp(i\vec{q}_{\vec{G}} \cdot \vec{r}). \quad (26)$$

Equation (23) is multiplied by Eq. (26), obtaining

$$\frac{1}{\varepsilon(\vec{r}, \omega)} \vec{\nabla} \cdot \vec{H}(\vec{r}) = i \sum_{\vec{G}''} \sum_{\vec{G}'} [\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G}''} (\vec{q}_{\vec{G}'} \times \vec{H}_{\vec{G}'}) \exp(i\vec{q}_{\vec{G}'' + \vec{G}} \cdot \vec{r}), \quad (27)$$

where $\vec{q}_{\vec{G}'' + \vec{G}} = \vec{q} + \vec{G}'' + \vec{G}'$.

The curl of Eq. (27) is

$$\vec{\nabla} \times \left[\frac{1}{\varepsilon(\vec{r}, \omega)} \vec{\nabla} \cdot \vec{H}(\vec{r}) \right] = i^2 \sum_{\vec{G}''} \sum_{\vec{G}'} [\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G}''} \left[\vec{q}_{\vec{G}'' + \vec{G}} \times (\vec{q}_{\vec{G}'} \times \vec{H}_{\vec{G}'}) \right] \exp(i\vec{q}_{\vec{G}'' + \vec{G}} \cdot \vec{r}). \quad (28)$$

Equations (28 and 25) are substituted into Eq. (13) to obtain the following:

$$i^2 \sum_{\vec{G}''} \sum_{\vec{G}'} [\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G}''} \left[\vec{q}_{\vec{G}'' + \vec{G}} \times (\vec{q}_{\vec{G}'} \times \vec{H}_{\vec{G}'}) \right] \exp(i\vec{q}_{\vec{G}'' + \vec{G}} \cdot \vec{r}) = \frac{\omega^2}{c^2} \sum_{\vec{G}'} \vec{H}_{\vec{G}'} \exp(i\vec{q}_{\vec{G}'} \cdot \vec{r}). \quad (29)$$

Equation (29) is multiplied by $\exp \left[-i(\vec{q} + \vec{G}) \cdot \vec{r} \right] d\Delta$ and integration on the complete UC is indicated, obtaining

$$\begin{aligned} & i^2 \int_{-UC}^{+UC} \sum_{\vec{G}''} \sum_{\vec{G}'} [\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G}''} \left[\vec{q}_{\vec{G}'' + \vec{G}} \times (\vec{q}_{\vec{G}'} \times \vec{H}_{\vec{G}'}) \right] \exp(i\vec{q}_{(\vec{G}'' + \vec{G}) - \vec{G}'} \cdot \vec{r}) d\Delta \\ & = \frac{\omega^2}{c^2} \int_{-UC}^{+UC} \sum_{\vec{G}'} \vec{H}_{\vec{G}'} \exp(i\vec{q}_{\vec{G} - \vec{G}} \cdot \vec{r}) d\Delta. \end{aligned} \quad (30)$$

Equation (30) is integrated using the orthogonality properties of Fourier functions, resulting

$$- \sum_{\vec{G}'} [\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G} - \vec{G}'} \left[\vec{q}_{\vec{G}} \times (\vec{q}_{\vec{G}'} \times \vec{H}_{\vec{G}'}) \right] = \frac{\omega^2}{c^2} \vec{H}_{\vec{G}}. \quad (31)$$

Finally, applying the rule of the double curl to Eq. (31) gives

$$\sum_{\vec{G}'} [\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G} - \vec{G}'} \left[\vec{H}_{\vec{G}'} (\vec{q}_{\vec{G}} \cdot \vec{q}_{\vec{G}'}) - \vec{q}_{\vec{G}'} (\vec{q}_{\vec{G}} \cdot \vec{H}_{\vec{G}'}) \right] = \frac{\omega^2}{c^2} \vec{H}_{\vec{G}}. \quad (32)$$

Equation (32) is the general equation to calculate the electromagnetic modes of 1D, 2D and 3D PCs. **Figure 6** shows a representation of the physical implication of this equation for a 2D PC. The red oscillation indicates the field $\vec{H}_{\vec{G}}$ with propagating wave vector $\vec{q}_{\vec{G}}$. The structure is given by the coefficients $[\varepsilon_{\vec{r}, \omega}^{-1}]_{\vec{G} - \vec{G}'}$, and the frequency f of the wave is in the term ω^2/c^2 , recalling that $\omega = 2\pi f$. If no frequency is found for a certain range of propagation vector, said frequency is in a forbidden bandgap.

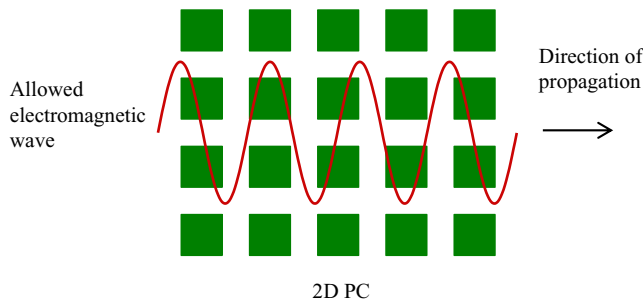


Figure 6. Representation of an allowed frequency in a two-dimensional PC of square cross-sectional cylinders ordered in a square lattice. The red oscillation indicates an electromagnetic wave propagating in the PC.

To check the explicit form of Eq. (32) for a 2D PC with out-of-plane propagation, see Ref. [17]. Equation (32) is an eigenvalue equation and has the following form:

$$A\vec{v} = \lambda\vec{v}, \quad (33)$$

where A is a square matrix, $\vec{v}$ is a column vector (eigenvector) and λ is a diagonal matrix (eigenvalue).

3. Numerical results

Once Eq. (32) is fed with the appropriate $\vec{q}_{\vec{G}}$ and $\vec{H}_{\vec{G}}$ for our 2D PC with out-of-plane propagation, the eigenvalue equation is solved with a standard matrix diagonalization method, obtaining the eigenvectors and the corresponding eigenvalues ω^2/c^2 which are used to implement the dispersive PC iterative method [17] to obtain the band structures. Error in the next calculations is estimated to be at most 1.5%.

The PC is an array of alumina rods of circular cross section ordered in a honeycomb lattice embedded in a vacuum matrix. The period of the PC is $a = 10m$. The rods radii are $r = 0.2a$. The alumina material is considered dispersive and its values for the dielectric constant $\epsilon(\omega)$, that can be seen in **Figure 3**, were calculated from experimental values of the real part of the refractive index n_r ; $\epsilon(\omega) = (n_r + ik)^2$. Although the imaginary part of the refractive index k is high for some frequencies within the studied spectrum, it was considered zero in the calculation of the dielectric constant for numerical convenience. Even though this approach is theoretical, it allows us to understand the behavior of the bands for unusual values and drastic changes in the dielectric constant.

In the following band structures, the electromagnetic modes are represented in red empty circles and are displayed in frequency against the propagation wave vector. The three grey segmented lines indicate the transparency windows with $\epsilon(\omega) = 1$ in with $13.49THz$, $14.67THz$ and $17.93THz$. In these three frequencies, the PC behaves as vacuum. It is important to recall that for each value of frequency there is a specific value for $\epsilon(\omega)$ (see **Figure 3**).

In **Figure 7**, the dispersive in-plane photonic band structure is shown. Two clear absolute band gaps appear approximately for $12.59 \leq THz \leq 13.19$ and $14.19 \leq THz \leq 17.09$. Another appears close around $17.39THz$. These gaps are indicated as grey rectangles. The regions where $\epsilon(\omega)$ is higher, the bands tend to be

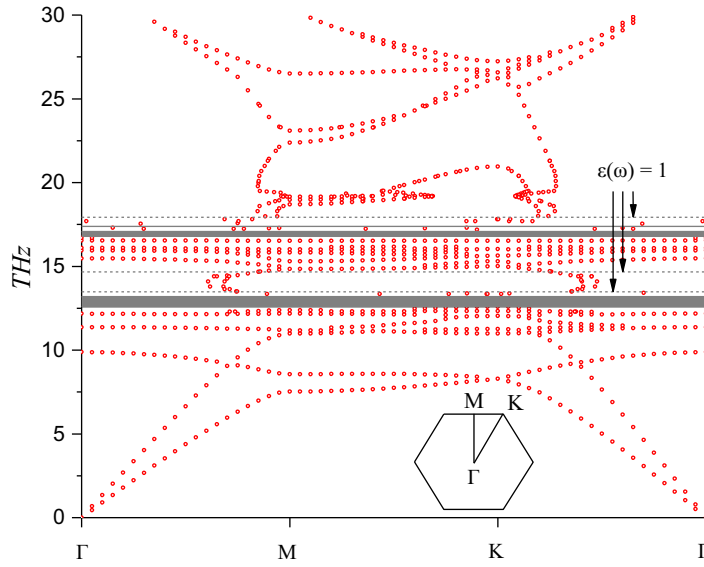


Figure 7.

Dispersive in-plane photonic band structure of alumina circular cross section rods ordered in honeycomb lattice embedded in vacuum. The vertical axis is the frequency in terahertz, and the horizontal one is the propagation wave vector. The period of the PC is $a = 10\text{m}$. The rods radii are $r = 0.2a$. Grey areas represent absolute band gaps. The inset shows the first Brillouin zone.

flat, indicating that the direction of propagation is less relevant. Between the first two transparency windows, and just above the third one, the band behavior making horizontal “U”s is considered unusual and is due to the “U” values of $\epsilon(\omega)$, where the bottom of “U” is almost zero. The band “cutting” seen in $\sim 19.18\text{THz}$ around the high symmetry point K is caused by the small peak in $\epsilon(\omega)$. Extra points were calculated there for a better appreciation. For the upper bands, the contrast between vacuum and the alumina decreases slowly.

In **Figure 8**, the dispersive out-of-plane photonic band structure starting at the high symmetry point Γ is shown. Flat bands appear for high values of $\epsilon(\omega)$ as in the in-plane case and tend to disappear going up in the normalized frequency as q_z is increased. The absolute band gaps also disappear, although there is not enough information to know at what point exactly. To know this, it is necessary to calculate the band structure in the irreducible Brillouin zone (Γ -M-K- Γ) plus the gradual increase of q_z until the band vanishes. A band touches the line of light in the three transparency windows. The areas named α and β show the horizontal “U” behavior due to the “U” values of $\epsilon(\omega)$. In α , $\epsilon(\omega)$ goes from 1 to close to zero then back to 1, so the band moves away from the line of light and returns to it. In β , $\epsilon(\omega)$ goes from almost zero to a small peak then back to almost zero, so the band gets close to the line of light and moves away from it rapidly. For upper frequencies, the band is in the left of the line of light since $\epsilon(\omega) < 1$ (not negative) and gets close when $\epsilon(\omega)$ tends to 1.

Our methodology presented erratic behavior due to some values of $\epsilon(\omega)$: in the two falls and where the value was very close to zero. Special care and some handwork were needed there.

The contributions of this chapter are listed below:

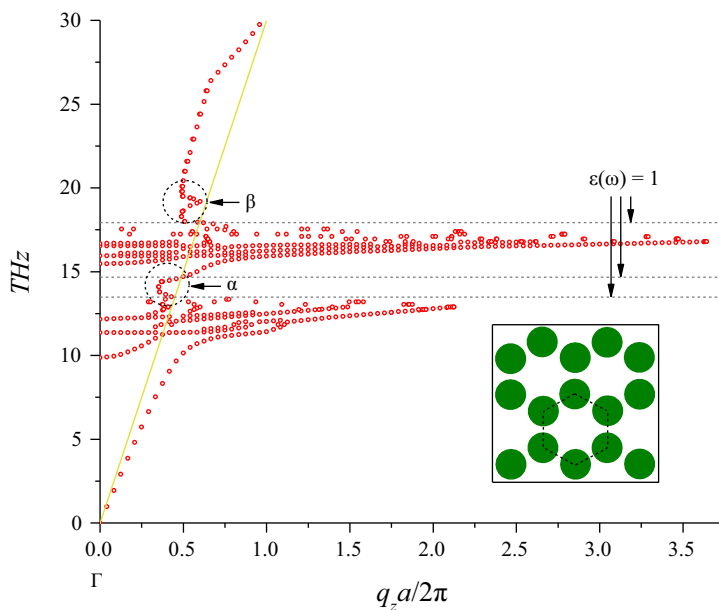


Figure 8.

Dispersive out-of-plane photonic band structure starting at Γ of alumina circular cross section rods ordered in honeycomb lattice embedded in vacuum. The vertical axis is the frequency in terahertz, and the horizontal one is the propagation wave vector with q_z its oblique component. The period of the PC is $a = 10\text{m}$. The rods radii are $r = 0.2a$. The yellow line represents the line of light in vacuum. The inset shows a not to scale transversal cut of the PC with the UC delimited in segmented line.

- I. Obtaining the electromagnetic wave equation from Maxwell's equations;
- II. Physical and mathematical definitions of a PC;
- III. Discussion of the dispersive nature of the dielectric constant of alumina;
- IV. Definition of the dielectric constant of a PC as a Fourier series;
- V. Definition of the magnetic field in a PC as a Bloch wave;
- VI. Obtaining the eigenvalues equation of a PC by substituting the definitions of the dielectric constant and the magnetic field into the electromagnetic wave equation;
- VII. Obtaining the dispersive in-plane photonic band structure of alumina; and
- VIII. Obtaining the dispersive out-of-plane photonic band structure of alumina.

4. Conclusions

In this chapter, a bandgap structure in the terahertz region is studied, considering the universal dispersive nature of the dielectric constant. In this study, the bands of a PC were calculated using the plane wave expansion method in conjunction with the dispersive PC iterative method. The dispersive in-plane and out-of-plane band structures of an alumina two-dimensional PC were obtained. First, the electromagnetic wave equation was derived from Maxwell's equations. The basic theory of PCs was presented later, discussing their definition, periodicities, geometries, and the dispersive nature of materials, while showing experimental data of the dielectric constant of alumina. These data showed that for some frequency values between 12 and 18 THz, the value of the dielectric constant can vary drastically. For a specific frequency, the value is more than nine times greater than that of diamond, and for some frequencies, the value is almost zero. Notably, the data showed that at three frequencies, alumina behaves like a vacuum, indicating that electromagnetic waves pass through the material freely. Later, the general eigenvalue equation was derived, proposing the dielectric function as a Fourier series and the magnetic field as a Bloch wave. In the results section, the band structures from 0 to 30 THz of a honeycomb lattice of circular cross-section alumina rods in a vacuum, considering experimental values for its dielectric constant, were presented, indicating notable behaviors due to unusual values of the dielectric constant. The in-plane band structure showed at least two absolute band gaps between 12.59 and 13.19 THz, and 16.79 and 17.09 THz. The out-of-plane band structure showed how the bands tend to disappear when the oblique component of the propagation wave vector increases. The behavior in both band structures was discussed as a function of the dispersive dielectric constant.

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Chapter 6

Suppress Mutual Coupling in MIMO Antennas Based on Artificial Electromagnetic Structures

Ijaz Khan, Yizhi Shao, Mian Muhammad Kamal, Chaoyun Song, Tianjun Ma and Mohammad Alibakhshikenari

Abstract

Modern wireless communication relies heavily on fast, reliable data transmission. Wireless technology faces stricter demands because spectrum resources are limited. As mobile devices incorporate multiple wireless systems, designing and situating numerous internal antennas within a confined space remains a formidable challenge. The demand for higher data rates and increased capacity without requiring additional radio spectrum or power in wireless communication systems has driven the development of multiple-input multiple-output (MIMO) antenna systems. MIMO antennas are used at both the transmitter and receiver ends of communication systems. This utilization aims to improve communication quality and increase channel capacity. One of the main challenges encountered when using this technology in a compact wireless terminal is reducing significant mutual coupling (MC) among antenna elements positioned close to one another. This phenomenon occurs due to the proximity of multiple antennas, leading to multi-antenna interference. The high MC between antenna elements leads to strong correlation, degrading MIMO system performance. Furthermore, it reduces the desired gain and efficiency because some of the power that could have been emitted is absorbed by the elements of other nearby antennas. The main aims of this chapter are to develop practical, compact artificial electromagnetic structures (AEMs) that effectively mitigate high MC in small wireless communication systems. A detailed analysis of the factors influencing MC, including antenna spacing, gain, axial ratio (AR), and the operating environment, is presented. Based on the above discussions, this chapter addresses the problems of MC between multiple antennas and the challenges of compact size.

Keywords: wireless communication, MIMO antenna, mutual coupling, gain, artificial electromagnetic structures

1. Introduction

Antennas are essential components for the efficient operation of wireless communication systems. The performance of antennas substantially enhances the features of wireless communication. Meeting the requirements for wireless communication may require using a single antenna in some circumstances. Wireless communication systems have increased data rates and channel capacity in applications such as Bluetooth, Wi-Fi, LTE, WLAN, and WiMAX. To fulfill this stipulation, employing a plurality of antennas is necessary. MIMO is a system that enhances system performance [1, 2]. The required data rate is increasing steadily as the user base for wireless communications grows. For instance, WLAN can support data rates up to 54.0 megabits per second. This capacity is expected to increase to 600 megabits per second soon. The successful transmission of audio and video streams in high-definition television format necessitates a transmission rate of 1 gigabit per second. Consequently, using a MIMO antenna is a commendable option for increasing channel capacity and enabling high data rates [3].

Microstrip antenna designers have recently incorporated MIMO technology into their designs, using two or more radiating patches. MIMO technology employs multiple antennas at the transmitter and receiver to improve the capacity of the communication channel while avoiding the need for additional power or frequency bandwidth. Wireless communication systems have significantly evolved across multiple generations, transitioning from single-input single-output (SISO) to MIMO configurations [4]. Providing telecommunications services requires high data transfer speeds. The quality of communication networks at the user end depends on their capacity. The microstrip antenna offers numerous advantages, making it the preferred choice for multiple elements in a MIMO antenna system. These advantages include a straightforward structure, a low profile, a simple fabrication process, and a conformal shape. MIMO antennas have garnered significant attention from communication professionals due to their ability to boost channel capacity without necessitating more frequency spectrum or power. The rapid expansion of wireless communication networks and the growing number of users have led to the widespread deployment of multiple-antenna systems to meet mandated requirements and achieve desired performance [5]. A MIMO system improves the reliability of radio communication links while enhancing spectral efficiency. The proliferation of wireless communication systems and the growing user base have necessitated the adoption of multiple-antenna systems to meet prescribed specifications and achieve desired performance levels. Inevitably, implementing a MIMO system enhances the spectrum efficiency of the radio communication link while simultaneously ensuring greater reliability. MIMO does not require more transmitted power or bandwidth than a SISO solution [6], despite utilizing numerous antennas on both the transmitter and receiver sides.

MIMO technology has been standardized and has become an essential component of consumer electronic devices, particularly for applications such as Wi-Fi, WiMAX, 4G, 5G, and LTE [7]. MIMO technology exhibits promising potential in enhancing the quality of service in communication scenarios where there is no direct line of sight (NLOS). The development of MIMO technology has remained influenced by various factors, including the capacity theorem and its applications in wireless technology involving pulse amplitude modulation signals [8]. Additionally, the study of multivariate analysis over the Gaussian channel with memory has advanced

MIMO [9]. Furthermore, multi-channel digital transmission systems have contributed to the development of MIMO [10]. Lastly, the use of beam-forming signal-processing applications for directional digital transmission and reception has also advanced the development of MIMO technology. The antennas within a wireless device are densely arranged due to limited space. The compact arrangement of an antenna array results in mutual coupling (MC), which adversely affects the radiation pattern, antenna gain, operational bandwidth, and radiation efficiency.

As satellite communications have advanced, circularly polarized (CP) antennas have been used to reduce physical losses, including fading and multi-pass losses. Incorporating a CP antenna is essential for accurately determining the position within the global positioning system (GPS). The use of the MIMO system for CP applications is attracting significant attention from industrial and academic researchers due to its multiple benefits in line-of-sight and multipath propagation environments. In line-of-sight and multipath situations, CP radiators exhibit lower polarization loss and less multi-path interference than linear polarization (LP) radiators [11]. In wireless communication, CP antennas are preferred over LP antennas because they maintain a stable connection between the sending and receiving devices regardless of orientation. Polarization mismatch is eliminated by equitably distributing power across the receiving radiators. This ensures that the signal is received with the same strength regardless of the CP radiator's orientation. This characteristic is significant in scenarios where the receiving antenna undergoes continuous changes in position and orientation relative to the transmitter. CP antennas are often recommended for practical wireless applications due to their advantageous characteristics. Specifically, a CP wave's electric field (E-field) component exhibits circular motion, reducing susceptibility to multipath fading. CP waves exhibit better resistance to propagation losses than linearly or elliptically polarized waves. CP is achieved by the truncated corner, resulting in a 90° phase difference. The commonly employed methodologies, namely, truncated corners, stubs, and slots, have been the subject of discourse. The CP-MIMO antenna has recently garnered significant attention in wireless applications, including satellite, WiMAX, WLAN, and 5G. Therefore, in these applications, using a CP-MIMO antenna becomes crucial. Consequently, it is imperative to thoroughly explore innovative designs that ensure CP in the broadside direction, minimize MC, and maintain a small form factor, all while offering sufficient bandwidth coverage [12]. In contemporary times, the main challenge in designing a CP-MIMO antenna is maintaining adequate sound isolation ($|S_{21}| > 15$ dB) and a consistent CP axial ratio (AR) of 3 dB across the intended operational bandwidth (BW).

Based on the preceding discourse, it is evident that the MIMO system represents the optimal solution for enhancing data rates and augmenting channel capacity. However, the performance of the MIMO system is further adversely affected by an additional detrimental factor: MC among the antennas. MC refers to the phenomenon wherein electromagnetic waves (EM) interact among the elements of an antenna system. This phenomenon is considered unfavorable because it causes energy to be absorbed by the adjacent antenna, preventing proper radiation. Several factors influence antenna array performance, including antenna radiation patterns and changes in input impedance. The extent of MC depends on the relative direction and spacing of the antenna parts. To mitigate MC effects, our study will involve novel artificial electromagnetic structures to enhance the performance of MIMO systems.

1.1 Performance evaluation of MIMO antenna system

1.1.1 MIMO system

Wireless communication technology has experienced significant growth, as evidenced by applications such as voice, video, and online gaming. In recent decades, there has been a notable shift in the field of antenna design for wireless technology toward the use of multiple radiators for various applications. This trend has gained prominence, particularly in response to customers' growing demand for higher transmission rates. The wireless communication system incorporates a substantial human factor. Many electronic and electrical devices widely use wireless systems in proximity. Modern applications require wireless communication systems to increase data speeds while maximizing channel capacity continually. Due to the high cost and limitations of higher-order modulation, increasing the bandwidth is not feasible. A novel strategy that harnesses the advantages of the transmission route must be embraced. MIMO technology offers a substantial enhancement to both data rate and channel capacity. MIMO methodology has garnered interest in contemporary wireless communication systems. Deploying multiple channels can provide a significant boost in capacity without requiring additional bandwidth or transmit power. Antennas are used for transmission to achieve both array gain and diversity gain, which ultimately improve spectral efficiency and reliability. MIMO antenna systems require substantial decoupling between antenna ports and a small physical footprint to be suitable for portable devices. Deploying multiple antennas on both links enables benefits beyond MIMO/diversity gains. MIMO systems can be described [13, 14] as using multiple antennas at both the transmitter and receiver. The fundamental idea behind using the MIMO system is the integration of sampled signals in the spatial domain at both transmitting and receiving ends. This integration aims to enhance data rates, improve communication quality, and enable the transmission of multiple parallel streams of spatial information. One benefit of using multiple antennas is that spatial geometry is added to time. The method of improving wireless technology capacity was first explored in the 1990s, and since then, it has been the focus of research worldwide [15, 16]. The utilization of multiple antennas at one end of the terminal until the 1990s was mainly focused on achieving diversity and estimating arrival directions. **Figure 1** demonstrates the MIMO antenna system block diagram.

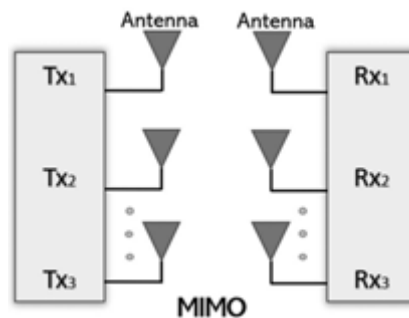


Figure 1.
Basic building diagram of the MIMO system [16].

This development subsequently paved the way for implementing spatial MIMO techniques, diversity, and beamforming strategies. The core idea behind space diversity is that the SNR can be enhanced by including a non-correlated output antenna element when random fading is caused by multipath propagation [17–20]. The ability of MIMO technology to improve channel capacity is based on the underlying concepts described by Shannon's channel capacity equation [1].

$$C = W \log_2 \left(1 + \frac{P}{N_0 W} \right) \quad (1)$$

where, C = channel capacity, W = bandwidth, P = transmitted power, and N_0 = power density noise.

The SNR is a crucial factor in the equation for channel capacity, expressed as (P/WN_0) . To enhance a system's channel capacity, you can either increase the SNR or expand the channel bandwidth. It is worth noting that the relationship between SNR and channel capacity is logarithmic: increasing SNR has diminishing returns beyond a certain threshold, and channel capacity saturates. Consequently, excessively boosting transmitted energy is inefficient and can lead to conflicts with other communication channels. Another way to improve channel capacity is to increase bandwidth, but this approach has limitations. Increasing bandwidth linearly is challenging due to the high cost of acquiring additional spectrum, multipath fading, and restrictions on transmitted energy to prevent interference with other channels. To overcome these constraints, MIMO technology has been employed to enhance channel capacity within specified energy and bandwidth constraints. Ultimately, MIMO antenna systems are pivotal components of many modern wireless technologies.

In the beginning, diversity was merely the initial step in mitigating the impacts of multipath propagation. Subsequently, when MIMO systems were introduced, diversity became a benefit for wireless communication systems. At present, diversity remains advantageous for wireless communication networks. The MIMO antenna system exploits latency spread and channel fading to achieve exponential increases in data transmission speeds. In contrast to the more common practice of using a single antenna, the MIMO method uses multiple sending antennas (denoted by the letter M) and multiple receiving antennas (denoted by the letter N). Each receiving antenna must be able to receive signals from all emitting antennas so that $M \times N$ signal transmission pathways can be created. The reliability of the multiple antenna system stems from its ability to provide supplementary spatial multiplexing and multiple independent signal propagation channels connecting the various transmitting and receiving antennas. Compared to more conventional single-antenna systems, we have a significant advantage in spectrum efficiency, enabling us to transmit several independent data flows across all those different pathways. MIMO is now an essential part of a wide variety of wireless standards, including GPS, WLAN, PCS, LTE, 4G, and 5G.

There are several types of wireless communication, as illustrated in **Figure 2** [21]. The first type of communication is single-input, single-output (SISO), which is the most fundamental. In a standard wireless communication system, a single antenna element is typically used at both the transmitter and receiver. This configuration is called a SISO system. Although this system is essential (usually does not require extensive signal processing), a severe radio multipath channel typically degrades the received signal quality (SNR), resulting in poor system reliability.

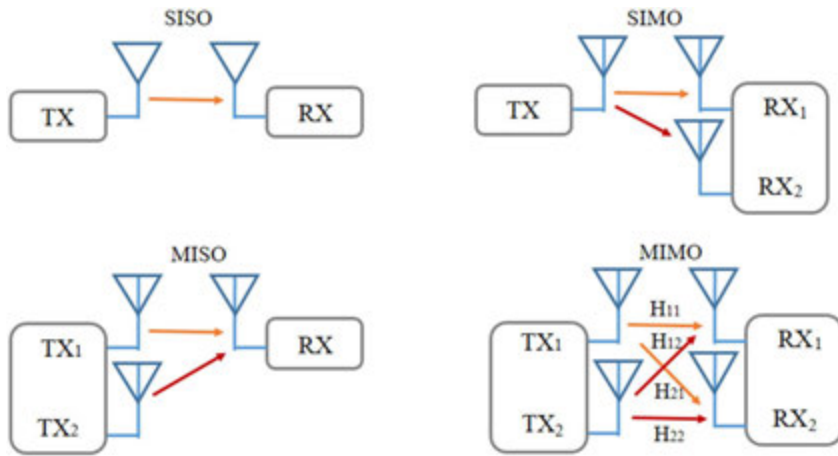


Figure 2.
Wireless communication types [21].

Moreover, according to the Shannon-Nyquist criterion [22], the data transmission rate of a SISO system depends on the SNR and the channel bandwidth, often referred to as the narrowband condition. The signal can be conveyed using the transmitter (T_x) and the receiver (R_x) in conjunction. One of the main benefits of this system is its high level of convenience. No additional post-processing is needed. Efficiency is reduced because a single antenna can only transmit to a single receiver. It is essential to have a vast number of costly broadcast and receive antennas if there are numerous users of the system. This complicates the receiver's ability to increase data throughput and reliability. Antenna diversity has been extensively used as an effective technique to improve the dependability of wireless systems and reduce the impact of fading.

The SISO system was transformed into a SIMO system with the inclusion of receive diversity in 1931. Since then, it has found extensive application in many wireless fields for 70 years. In the early 1990s, a novel form of multi-antenna technique was introduced. This method, known as transmit diversity, was employed to establish a multiple-input single-output (MISO) system. Its implementation was observed in the 2G mobile system to improve the base station's radio coverage in the downlink channel.

Moreover, the single-input multiple-output (SIMO) concept uses multiple receiving antennas and a single transmit antenna, as depicted in **Figure 2**. One advantage of this technology is its ability to automatically receive signals. The demand for a post-processing component has been identified. When the receiver is integrated into a mobile phone, its size is limited, and the battery capacity becomes an additional factor to consider. There are two potential architectures for single-input multiple-output: the first involves the path of the most potent signal and adjusting that antenna. Merging two signals into a unified signal is an alternative to enhancing the total signal strength. The fading effect must be accounted for to prevent the signal from becoming distorted. The signal-to-noise ratio may improve if the fading effect can be removed. **Figure 2** shows the concept of MISO communication. In this structure, identical data is transmitted across many cables, but a single antenna receives it. This method has the advantage that the transmitter

handles the signal processing. The transmitter performs the processing because it has a larger structure that can handle more complex and sophisticated tasks. The power consumption of the mobile phone's reception can be reduced.

In conclusion, the utilization of multiple reception devices at the transmit and receive ports, as depicted in **Figure 2**, enhances the throughput capacity in MIMO systems. In this context, the symbols T and R represent the signals used for transmission and reception, respectively. The variables m and n denote the number of transmitting and receiving apparatuses. Moreover, the symbol H represents the channel matrix, denoted as H_{nm} . This notation can be further refined as [23–25], resulting in a more precise signal.

$$R_1 = H_{11}T_1 + H_{12}T_2 + \cdots + H_{1m}T_m \quad (2)$$

$$R_2 = H_{21}T_1 + H_{22}T_2 + \cdots + H_{2m}T_m \quad (3)$$

$$R_n = H_{n1}T_1 + H_{n2}T_2 + \cdots + H_{nm}T_m \quad (4)$$

To combine the equations from above into a matrix, we have:

$$\begin{bmatrix} R_1 \\ R_2 \\ R_n \end{bmatrix} = \begin{bmatrix} H_{11} & H_{12} & \cdots & H_{1m} \\ H_{21} & H_{22} & \cdots & H_{2m} \\ H_{n1} & H_{n2} & \cdots & H_{nm} \end{bmatrix} \begin{bmatrix} T_1 \\ T_2 \\ T_n \end{bmatrix} \quad (5)$$

Another straightforward expression can be written as:

$$R = HT \quad (6)$$

When considering the presence of Gaussian noise,

where, $T \in C^m$, $R \in C^n$, $T \in C^n$, $R \in C^n$, $\omega \approx CN(0, N_0 1_n)$ and channel matrix $H \in C^{n \times m}$ thought to be consistent and acknowledged transmitter and receiver. The H_{nm} is the gain from the m transmitter to the n receiver. Consider how noise W causes the receiver symbol R to degrade. The linear equation appears as follows:

The detected transmitted signal can be calculated as follows, given an approximate estimate of the channel transmission matrix H :

$$T = H^{-1}\bar{R} \quad (7)$$

Communication systems that contain an array of transmitters and receivers larger than a 2×2 MIMO configuration are referred to as having a massive MIMO structure. The diversity pattern is a method that can be applied in communication to improve the accuracy of the communication signal.

1.1.2 MIMO antenna challenges

MIMO antennas can increase channel capacity without additional frequency spectrum or transmission power. However, several factors can limit system performance, particularly in small devices. Spatial limitations and mutual coupling constrain the design of MIMO antennas due to electromagnetic interactions between the components of the MIMO system. The limited physical space available in base stations and mobile terminals is a significant obstacle to increasing the number of antenna elements [26]. Radiating elements should be spaced more than half a wavelength apart to fully utilize the system's space and minimize correlation and total active reflection coefficients between channels. However, dense placement of

antenna elements can increase dimensional correlation and degrade system performance, thereby affecting spatial diversity within a fixed physical space. As the number of signal sources and antenna elements increases, spatial correlation grows, potentially decreasing the system's channel capacity. These considerations influence the feasibility of MIMO large arrays with confined spatial dimensions. Designing compact antennas without compromising performance is challenging due to the physical constraints on antenna bandwidth and efficiency [27]. The trend toward miniaturization in technology, driven by the prevalence of mobile terminals and compact wearable devices, necessitates optimizing antenna deployment within existing spatial constraints.

Mutual coupling, the electromagnetic interaction between MIMO radiating elements, presents an additional challenge. The current distribution on the radiating elements of one antenna is influenced by the fields generated by other elements, thereby affecting the radiation pattern and impedance matching of the MIMO element. Electromagnetic interactions among antenna elements can adversely affect the performance of a MIMO system [28].

A number of interrelated physical processes, including capacitive (electric-field) contact, inductive (magnetic-field) coupling, space-wave overlap, substrate-guided surface waves, and shared ground or feedline effects, can result in mutual coupling. In compact systems, many mechanisms coexist, creating a complex coupling environment that is difficult to classify or forecast accurately. When two radiators are positioned adjacently, one may conceptualize the overlap of their electric and magnetic fields, while surface waves traverse beneath or within the substrate [29]. **Figure 3** conceptually illustrates this multi-path interaction, summarizing the simultaneous electric, magnetic, surface-wave, space-wave, and ground-current coupling routes present in a compact MIMO environment.

To address this, it is necessary to consolidate all radiating elements into a single aperture, thereby transforming the antenna design into a multiport problem. Traditional array design principles may not apply to small devices. Researchers employ various techniques to mitigate electromagnetic interactions among array members, including multiport configurations, inserts, parasitic elements, DGS, FSS (frequency selective surfaces), and metamaterials. These methods will be elaborated upon in the subsequent discussion.

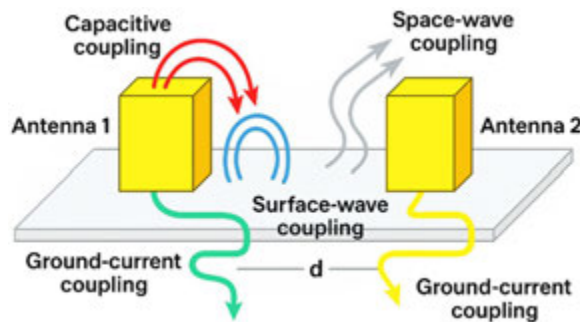


Figure 3. Conceptual depiction of mutual coupling processes between two compact MIMO antennas, illustrating the simultaneous presence of capacitive, inductive, surface-wave, space-wave, and ground-current coupling pathways in electrically tiny platforms [29].

1.1.3 MIMO system impedance matching

When multiple antennas are positioned close together, they interact with one another, a phenomenon known as mutual coupling. This interaction can be understood as one antenna element inducing a current in another due to their proximity. Consequently, this induced current affects the input impedance of both antenna elements. The input impedance is determined by two components: self-impedance Z_{ii} , which pertains to each antenna element's impedance in isolation, and mutual impedance Z_{ij} , which is the impedance resulting from their interaction. These changes collectively form an impedance matrix, as illustrated in Eq. (8).

$$Z = \begin{pmatrix} z_{11} & \cdots & z_{1j} \\ \vdots & \ddots & \vdots \\ Z_{i1} & \cdots & z_{ij} \end{pmatrix} \quad (8)$$

The creation of the impedance matrix has fundamentally changed how we understand and manage antenna bandwidth. Previously, antenna bandwidth was primarily determined by the reflection coefficient alone. However, with the introduction of the S-parameter matrix in Eq. (9), we now see that the antenna bandwidth depends on both the reflection and transmission coefficients. In this context, S_{ij} denotes the ratio of the reflected wave voltage to the incident voltage at the i^{th} antenna element port. This measurement is taken when all other elements are properly terminated with matched loads. It indicates how much of the incident voltage is reflected from the antenna element.

On the other hand, S_{ij} represents the ratio of the voltage transferred from the j^{th} element's port to the i^{th} element's port in relation to the incident voltage at the j^{th} element's port. This parameter, S_{ij} , is also commonly referred to as the MC coefficient. It quantifies the extent to which the voltage at one antenna element's port is coupled or transferred to another antenna element's port. So, to sum it up, the impedance matrix and the S-parameter matrix have expanded our understanding of antenna bandwidth by considering both reflection coefficients S_{ij} and the MC coefficients S_{ij} among elements, leading to a more comprehensive view of how antennas operate and affect one another.

$$S = \begin{pmatrix} S_{11} & \cdots & S_{1j} \\ \vdots & \ddots & \vdots \\ S_{i1} & \cdots & S_{ij} \end{pmatrix} \quad (9)$$

1.2 MIMO antenna key parameters

MIMO antenna systems require the calculation of additional vital metrics and characteristics beyond standard antenna parameters, such as resonance frequency, bandwidth, radiation pattern, efficiency, and gain. This is necessary to thoroughly analyze and characterize the MIMO system's performance. The following subsections will examine the performance metrics and parameters.

1.2.1 Envelope correlation coefficient of MIMO antenna

The envelope correlation coefficient (ECC) is an important parameter often discussed in MIMO system discussions. The ECC parameter is widely adopted as the standard for MIMO systems. The signals transmitted to other receivers in the preceding section were considered noise for particular antennae. The correlation between two antennas is known as ECC [30]. Diversity systems have employed two practical approaches to calculate the ECC between diversity antennas. The first method relies on analyzing the antenna's radiation patterns. However, this approach is known to be quite time-consuming because it requires intricate numerical or experimental analysis, which can be cumbersome. Examining the radiation patterns generated during the MIMO antenna design can determine the ECC. The performance of these antennas can be evaluated through measurements conducted in a reverberation chamber. This chamber can provide relative correlation coefficient measurements in a specific environment, offering a more practical and efficient assessment.

$$\rho_{ECC} = \frac{\iint_{4\pi} E_1(\theta, \phi) \cdot E_2(\theta, \phi) d\Omega}{\sqrt{\iint_{4\pi} E_1(\theta, \phi) \cdot E_1(\theta, \phi) d\Omega} \sqrt{\iint_{4\pi} E_2(\theta, \phi) \cdot E_2(\theta, \phi) d\Omega}} \quad (10)$$

E_1 and E_2 represent the far-field characteristics shown by antenna-1 and antenna-2, respectively. The objective is to reduce the ECC to achieve sufficient isolation and improve the SINR. The lower the ECC, the more data can be transmitted across the MIMO structure. There is one more calculation that is related to ECC. The calculation process can be simplified by obtaining the S-parameter values for $|S_{11}|$, $|S_{12}|$, $|S_{21}|$, and $|S_{22}|$, as well as the efficiencies η_1 and η_2 .

$$\rho_{ECC} = \frac{S_{11}S_{12}^* + S_{11}S_{22}^*}{\sqrt{(1-|S_{11}|^2 - |S_{21}|^2)(1-|S_{22}|^2 - |S_{12}|^2)}\eta_1\eta_2} \quad (11)$$

The claim is well-founded if there is no discernible association or if the losses among the two receiving lines are entirely independent. Eq. (10) exhibits lower computational complexity than Eq. (11) and can be conveniently measured using a vector network analyzer. However, the drawback of Eq. (10) is its reduced precision, which makes it difficult to determine antenna efficiency.

1.2.2 Diversity gain of MIMO antenna

Diversity gain is another criterion that may be used to evaluate the implementation of the MIMO system. The antenna DG is used as a metric to assess the performance enhancement achieved by implementing diversity. It quantifies the increase in SNR when using multiple antennas compared to a single antenna. The following equation [31] illustrates the relationship between ECC and the diversity gain.

$$G_{app} = 10 * \sqrt{1 - |ECC|} \quad (12)$$

In practical applications, it is observed that the diversity gain typically amounts to 10 dB.

1.2.3 Total active reflection coefficient of MIMO antenna

S-parameters are used to calculate power ratios between ports, with one port supplied and the others terminated with a matched load. However, in MIMO systems, several ports are used simultaneously. As a result, the S-parameters are insufficient to describe the MIMO system's performance accurately. Therefore, the total antenna reflection coefficient (TARC) parameter is effectively employed to enhance the characterization of MIMO systems [32]. The TARC in a multi-element antenna system is mathematically expressed as the ratio of the square root of the cumulative reflected power, obtained by summing the incident power across all ports and subtracting the total radiated power, divided by the square root of the overall incident power. The TARC can be determined for an N-element MIMO system using the formula presented in equation (13) [33].

$$\Gamma_a^t = \frac{\sqrt{\sum_{i=1}^N |b_i|^2}}{\sqrt{\sum_{i=1}^N |a_i|^2}} \quad (13)$$

The variables a_i and b_i represent the reflected and incident signals, respectively. In a multi-port network, with all ports having equal characteristic impedance, the relationship between reflected and incident signals can be expressed using the scattering matrix, as stated in Eq. (14).

$$[b] = [s][a] \quad (14)$$

The excitation vector is denoted as $[a]$, the scattered vector as $[b]$, and the scattering matrix of the multi-port antenna as $[s]$. In a MIMO system with N elements, the dimension of the matrix $[s]$ is $N \times N$. The TARC affects MIMO systems by enabling the estimation of the appropriate bandwidth of MIMO antennas, similar to how S-parameters are used to determine the bandwidth of single antennas. This technique considers coupling, return loss, and random signal mixtures across ports. A single TARC curve can be used to characterize the MIMO system for a given phase stimulation across the elements. The TARC for a 2×2 MIMO system can be calculated using the formula presented in [34].

$$\Gamma_a^t = \frac{1}{\sqrt{2}} \sqrt{|S_{11} + S_{12}e^{j\theta}|^2 + |S_{21} + S_{22}e^{j\theta}|^2} \quad (15)$$

where θ represents the input variable phase difference from the channel-induced excitation on the two-port, the phase difference can have substantial effects on the TARC magnitude [35]. The MIMO channel's complex structure makes the value of θ unpredictable. As a result, TARC needs to be analyzed for phase variations that range from 0 to 180 degrees. The TARC formula was derived from the S-parameters of a two-element MIMO system, as depicted in Eq. (16). Yet, we aim to extend this formula so that it can be used for any number of elements, as:

$$\Gamma_a^t = \frac{1}{\sqrt{N}} = \sqrt{\sum_{i=1}^N \left| \sum_{j=1}^N S_{ij} e^{j\theta} \right|^2} \quad (16)$$

The variable j_θ represents the phase difference between the excitation of the first and j^{th} elements. The TARC is a real number in the interval $[0, 1]$. When the TARC value is estimated to be zero, it indicates that all transmitted power is emitted and not reflected to a separate feeding port.

1.2.4 Channel capacity loss of MIMO antenna

Channel capacity loss (CCL) is the maximum data rate that can be managed in a given channel within a fading environment [36]. The following equation can be used to calculate CCL from either measured or simulated S-parameters while considering a high SNR value:

$$CCL = -\log_2 \det(\varphi^R) \quad (17)$$

Channel capacity [37] also affects MIMO performance and depends on antenna scattering characteristics. Additionally, the performance of the MIMO antenna is characterized by the number of transmitted and received streams. In this analysis, we will examine a transmission channel that exhibits multipath propagation. This channel is characterized by multiple transmitting antennas, N_t , on one end and receiving antennas, N_r , on the opposing end. The following equation can represent the mathematical expression for channel capacity.

$$C = \log_2 \left(1N_r + \frac{P}{NN_t} HH^* \right) \quad (18)$$

In the given context, P represents the signal power, and N represents the noise power. The symbol $*$ indicates the complex conjugate transpose operation. H denotes the channel matrix, whereas H^* represents the transpose of the channel matrix. The behavior of H depends on the correlation or MC between antenna elements, which is determined by many channel parameters, including delay, noise, gain, interference, and so forth. The channel parameters must be appropriately understood to determine the significance of H . Equation (19) depicts a two-port MIMO channel matrix:

$$H = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \quad (19)$$

1.2.5 Multiplexing efficiency of MIMO antenna

The criterion of multiplexing efficiency (ME) determines the power-related performance of a MIMO array antenna in a high-SNR environment at the receiver. It demonstrates the loss in power efficiency that occurs when multiple antennas are used instead of an ideal array arrangement in the same wireless channel to achieve the same channel capacity. This is used for spatial multiplexing, as described in the coming chapter, where we will understand how the decoupling structure enhances the ME of the suggested antenna. ME for two-port prototype:

$$\eta_{mux} = \sqrt{(\eta_1 \eta_2 (1 - |\Gamma|^2))} \quad (20)$$

where, η_i = Totalefficiencyofantenna $|\Gamma|$ = Complexcorrelation.

The variables η_1 and η_2 represent the total efficiency of the MIMO antenna elements. The ME, or η_{mux} , is introduced as a straightforward means to assess the performance of MIMO systems [38]. ME is the power penalty incurred by an efficient multiple-antenna system when aiming to achieve a specific capacity, unlike an ideal system with 100% total efficiency and zero correlation. In MIMO antennas, the embedded element efficiency is a crucial performance metric. This metric gauges the effectiveness of a MIMO antenna when only one of its ports is activated, with the remaining ports terminated adequately with matched loads. This parameter significantly influences the fundamental advantages that MIMO technology can deliver, such as diversity gain and channel capacity.

1.3 Multiple antennas mutual coupling

Mutual coupling is the energy a nearby antenna absorbs when another antenna emits electromagnetic radiation. MC significantly alters the MIMO antenna's input impedance, reflection coefficient, and radiation pattern. The theoretical framework for simplified MC, MC_{ij} , can be represented as [39].

$$MC_{ij} = \exp\left(\frac{2d_{ij}}{\lambda} (a + j\pi), i = j\right) \quad (21)$$

$$MC_{ij} = 1 - \frac{1}{N} \sum_i \sum_{i \neq j} MC_{ij} \quad (22)$$

The distance between the i^{th} and j^{th} antenna elements is denoted by d_{ij} . Moreover, the value α serves as the controlling variable for the coupling level, while N represents the total number of elements within the array. Array configuration and other elements' excitations influence MC. The dB-valued S-parameter between the i^{th} and j^{th} antenna elements is often computed, along with the isolation between them, which is represented as $-20 \log_{10} |S_{ij}|$. Nevertheless, it is essential to note that the precise mechanism of MC is greatly influenced by the transmitting and receiving modes [40, 41]. The transmitting and receiving modes determine the particular mechanisms of MC. The MC methods in the transmitting and receiving modes are discussed separately [40]. The arrays contain two antenna elements in the transmitting mode [42], as illustrated in **Figure 4(a)**. To facilitate the analysis, we will assume a two-element antenna, m and n , within an array. A source is assumed to be connected to element n , and the released source A energy radiates into space B and toward the m^{th} element C . A portion of the received energy by the m^{th} antenna is redirected into the surrounding space, indicated as D , and the remaining energy propagates toward the generator, denoted as E . The n^{th} element, F , will absorb a portion of the energy D that has been rescattered. This process of mutual interaction will continue. However, as the scattered energy from each iteration decreases, it is typically sufficient to consider only the first few iterations. The overall far-field is the result of combining the radiated and scattered fields. Thus, the MC tends to change the antenna array. The wave denoted as E combines vectorially with the reflected and incident waves originating from the m^{th} element.

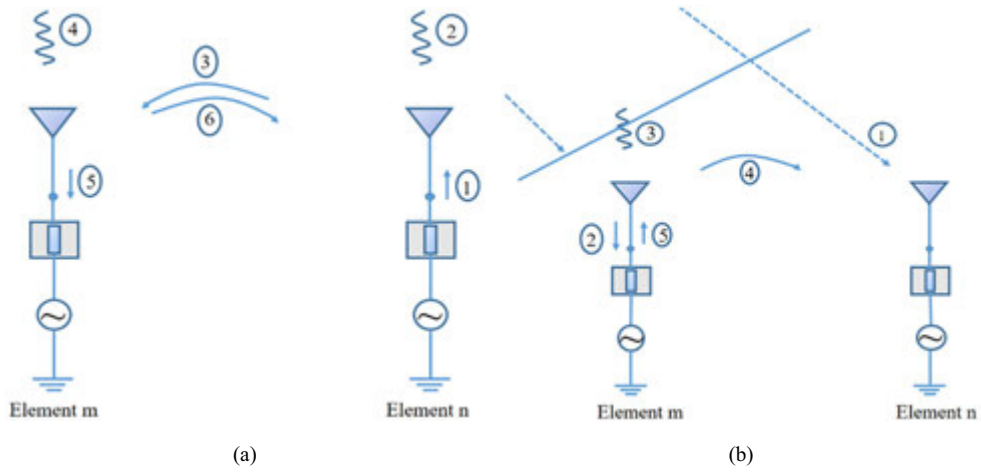


Figure 4. Diagram of mutual coupling mechanism, (a) Transmitting mode and (b) Receiving mode [42].

This interaction amplifies the standing wave, thereby modifying the input impedance of the m^{th} element. The presence of MC affects the antenna impedance and self-impedance.

In the transmitting mode, a large number of antenna system ports can produce random-phase excitations. TARC is commonly used to assess the reflection coefficient of a MIMO array. This evaluation is conducted by applying random-phase excitations to the array's element ports. TARC is calculated by dividing the total generated power by the square root of the total power created by all excitations, less the total power radiated [43]. When analyzing systems subjected to random-phase excitations at their terminals, the TARC considers several factors, including impedance matching, MC, and radiation efficiency. In the receiving mode, a wave enters the i^{th} element, producing an incident current, as shown in **Figure 4(b)**. Let us consider a scenario in which a plane wave, denoted as A, is incident upon an array, with the wave front initially reaching the m^{th} element of the array. The m^{th} element is initially introduced to the induction of an electric current. Some incident waves are transmitted toward the receiver as part B, while another portion is scattered back into the surrounding environment as component C. Some of the waves that have been scattered are directed toward the n^{th} element, denoted as D, where they combine vectorially with the incident plane wave, denoted as E. The wave an element receives is the vector sum of the waves coupled from other elements and those received directly. To optimize the amount of energy absorbed and minimize the energy scattered back, it is necessary to carefully choose the termination impedance of the m^{th} element. This selection should ensure that the reflected wave, denoted as E, effectively cancels the rescattered wave, denoted as C. The performance of the antenna element under investigation can be assessed in the receiving mode by stimulating it with another element that is terminated with a 50-ohm load.

In **Figure 5**, the illustration depicts the MC phenomena observed among elements within a close array. Specifically, it demonstrates the transfer of a fraction of the energy in an incoming field from a receiving element to an adjacent antenna element.

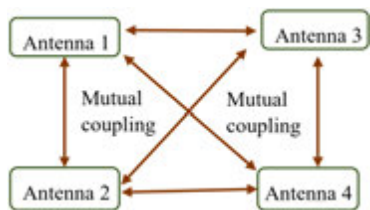


Figure 5.
A schematic representation of compact array elements mutual coupling connection [29].

Consequently, the current supplied to each transmitting antenna in an array is the sum of the currents generated during each transmission. It encompasses the current induced by neighboring antenna elements. The same principle applies to the array's reception components. This indicates that energy exchanges between a specific element within an array and a distant location occur via a direct path and, indirectly, via scattering from the other antennas within the array.

1.4 MIMO antenna artificial electromagnetic structures

1.4.1 FSS-based four-port MIMO antenna

This work presents a circular-slit four-port MIMO antenna that achieves effective mutual coupling suppression [44]. The circular-sloped radiating patch provides improved radiation performance, while the common ground plane simplifies fabrication and reduces inter-element interaction. The key feature is a band-stop FSS printed on the same layer as the radiator, acting as a spatial filter that blocks coupling paths and maintains stable behavior up to 40° for both TE and TM incidence. A circular-ring FSS unit cell positioned between the elements further boosts isolation to better than -20 dB and enhances gain. With its simple geometry, compact size, and ease of integration, the design offers an efficient and practical solution for improving isolation, gain, and overall MIMO antenna performance.

Figure 6 depicts the proposed band-stop FSS unit cell. The band-stop FSS is specifically engineered to reject a specific frequency range. This implies that it either

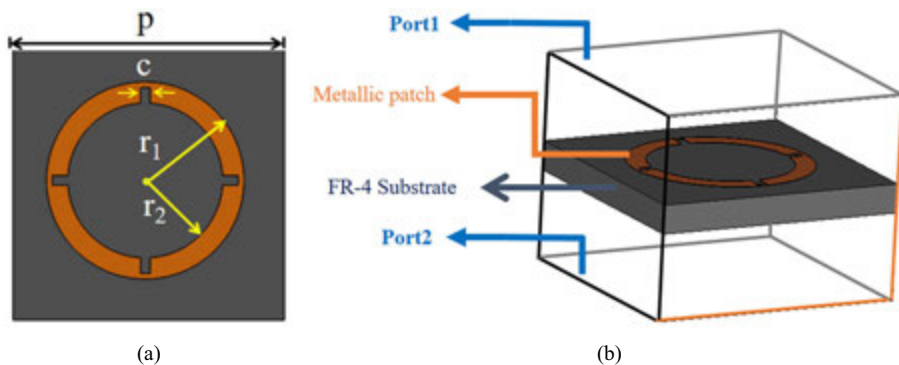


Figure 6.
(a) The proposed FSS unit cell front view, and (b) 3-D view of the unit cell [44].

reflects or absorbs electromagnetic waves in this range, thereby obstructing their transmission. The metallic slotted circular ring FSS is mounted on a 1.6 mm FR-4 substrate with a dielectric constant of 4.3. Full-wave electromagnetic simulations of the designed FSS element are carried out in CST Microwave Studio, along with the excitation of Floquet ports. **Figure 7(a)** presents the S-parameters of the proposed FSS unit cell. The slotted circular-ring FSS resonates at 8.3 GHz and provides a wide band-stop response from 7.2 to 9.1 GHz. The slot within the ring improves impedance matching across the desired band. The S-parameters confirm strong stopband behavior, with a 1.9 GHz bandwidth and a minimum $|S_{21}|$ below -30 dB. As shown in **Figure 7(b)**, the TE and TM responses are nearly identical, demonstrating polarization-independent performance. **Figure 7(c, d)** confirms that the proposed FSS unit cell remains highly stable under varying incident angles. The TE response is consistent up to 40°, with only minor resonant shifts, while the TM response shows minimal change. Because $|S_{21}|$ remains nearly constant in both cases, the FSS maintains strong band-stop behavior and effectively suppresses coupling, improving isolation between MIMO elements.

The four-element geometry proposed here aims to maximize input while minimizing interference. The suggested antenna was simulated on a 1.6 mm-thick FR4 substrate, as shown in **Figure 8(a)**. The overall dimensions of the four-port

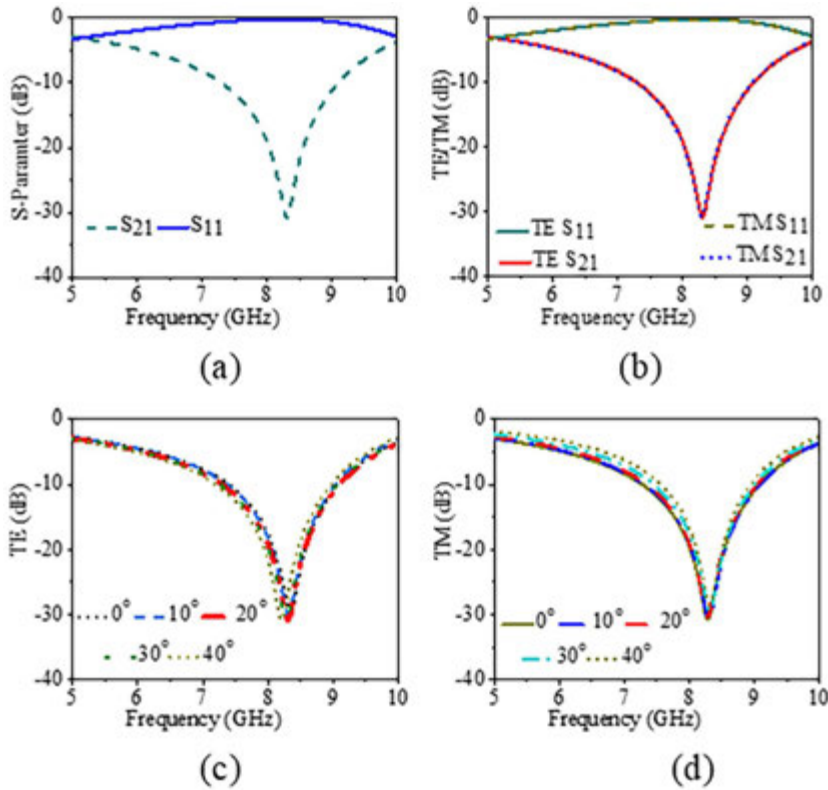


Figure 7.
 (a) The proposed FSS unit cell S-parameters, (b) TE and TM S-parameters, (c) TE incidence angles, and (d) TM incidence angles [44].

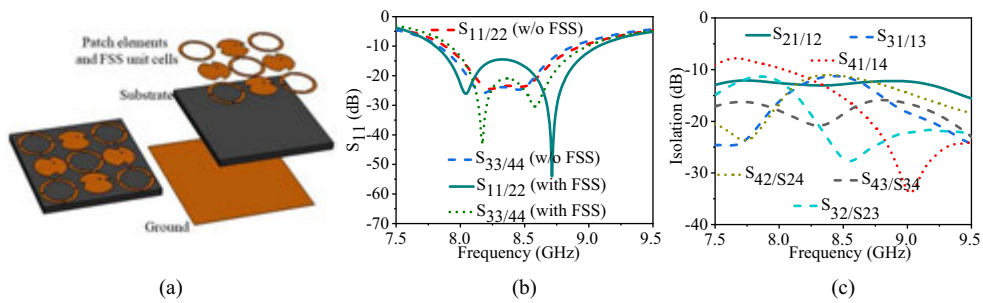


Figure 8. (a) Structure of the proposed four-port antenna with decoupling FSS, (b) Reflection coefficient with and without FSS, and (c) Isolation without FSS [44].

antenna are $33 \times 33 \times 1.6 \text{ mm}^3$. A 50Ω coaxial feed supplies power to the antenna. Following the research on a single-element antenna, the same radiator is uniquely placed on the common ground for a four-port antenna. For impedance matching, the patches at ports 1 and 4 are rotated anticlockwise. In contrast, the port-2 patch and port-3 patch are oriented in the same direction as the single-element antenna.

Figure 8(c) shows that the antenna without FSS offers weak isolation of about -11 to -12 dB . To mitigate this, a slotted circular patch and slotted-ring FSS cells are added without increasing the antenna size. Five FSS units placed between the closely spaced radiators (0.29λ apart) effectively block surface-wave propagation, substantially reducing mutual coupling and improving the overall MIMO performance.

As seen in **Figure 10(a)**, the FSS slightly broadens the $|S_{11}|$ bandwidth to $7.8\text{--}9.08 \text{ GHz}$. Isolation improves significantly: $|S_{21}|$ drops from about -11 dB to -25 dB , $|S_{13}|$ improves by $\sim 14 \text{ dB}$, and $|S_{14}|$ reaches -39 dB at 8.5 GHz . With the FSS, all port isolations exceed 20 dB , giving more than a 10 dB reduction in mutual coupling. The measurement setup is shown in **Figures 9**, and **10(b, c)**, confirming good agreement between simulated and measured S-parameters.

The gain performance of the proposed antenna was evaluated for Port 1, with similar behavior observed across all ports. Without the FSS, the MIMO antenna achieves a simulated peak gain of 3.7 dBi . With the FSS added, the peak gain

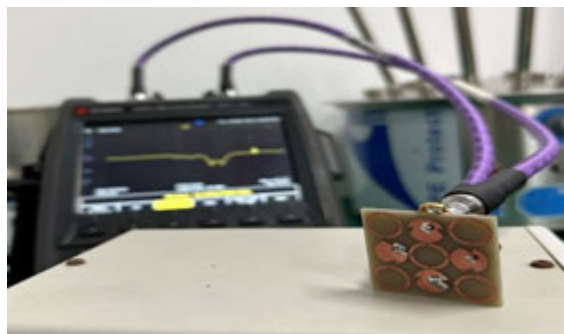


Figure 9. Measurement setup of the proposed antenna with FSS [44].

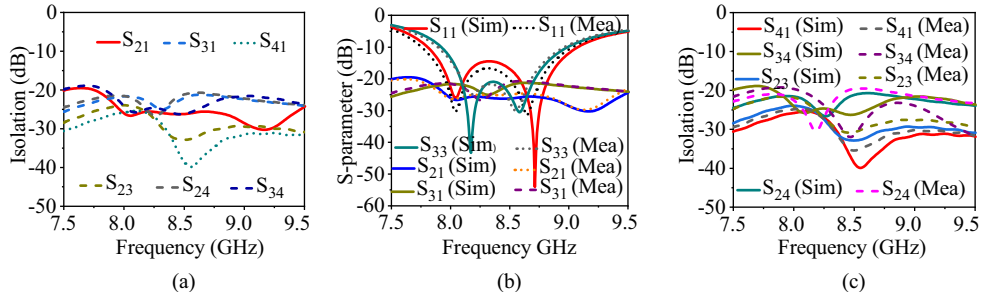


Figure 10.

(a) Isolation with FSS, (b) Simulation and measured S-parameters with FSS, and (c) Simulation and measured isolation with FSS [44].

increases to 5.2 dBi at 8.7 GHz, representing a 1.5 dBi improvement, as shown in **Figure 11(a)**. The close agreement between measured and simulated results further confirms the effectiveness of the FSS in enhancing antenna gain. **Figure 11(b)** presents the radiation efficiency with and without the FSS. The efficiency with the FSS exceeds 75% across operating frequencies. **Figure 12** shows that the proposed antenna maintains stable co- and cross-polarization patterns in both E- and H-planes at 8.3 GHz, with the H-plane remaining omnidirectional. Adding the FSS introduces only slight broadside changes while improving pattern stability by reducing mutual coupling. The measured and simulated patterns match closely, confirming the design's accuracy.

Figure 13 shows the surface current distribution at 8.3 GHz. Without the FSS (**Figure 13a**), exciting Port 1 causes a strong current to spill onto adjacent elements, indicating high coupling. With the FSS (**Figure 13b**), the current remains confined to the excited radiator and is absorbed by the FSS, preventing it from reaching the other ports. Similar behavior is observed for Ports 2, 3, and 4, confirming effective suppression of coupling. **Figure 14(a)** also represents the DG of the suggested MIMO antenna, where the DG is near 10 dB across the entire band with FSS. The CCL is shown in **Figure 14(b)** and is 0.1 bits/s/Hz across the whole band.

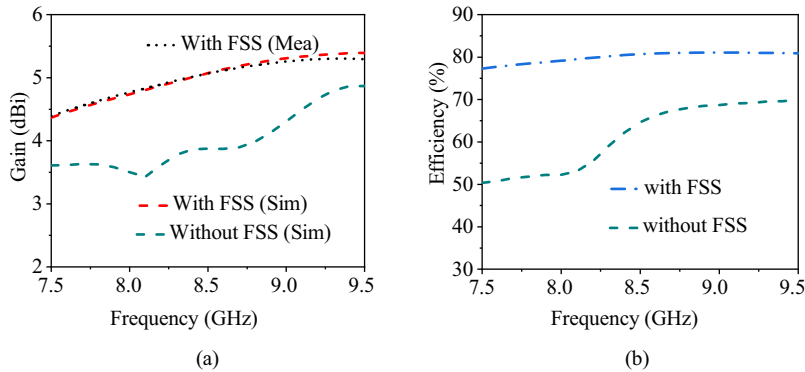


Figure 11.

(a) Simulated and measured the gain of the four-port antenna, and (b) Simulated efficiency with and without FSS [44].

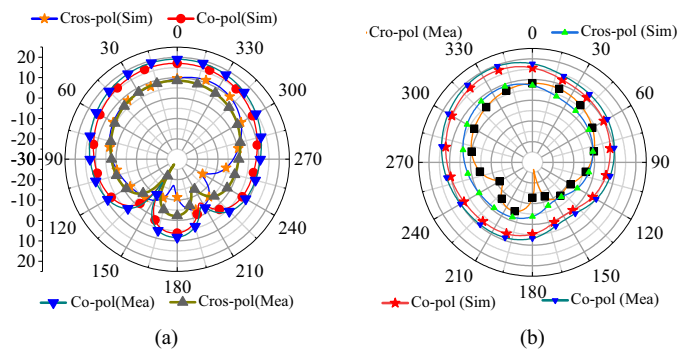


Figure 12.
Simulated and measured co and cross-radiation pattern at 8.3 GHz, (a) E-plane, and (b) H-plane [44].

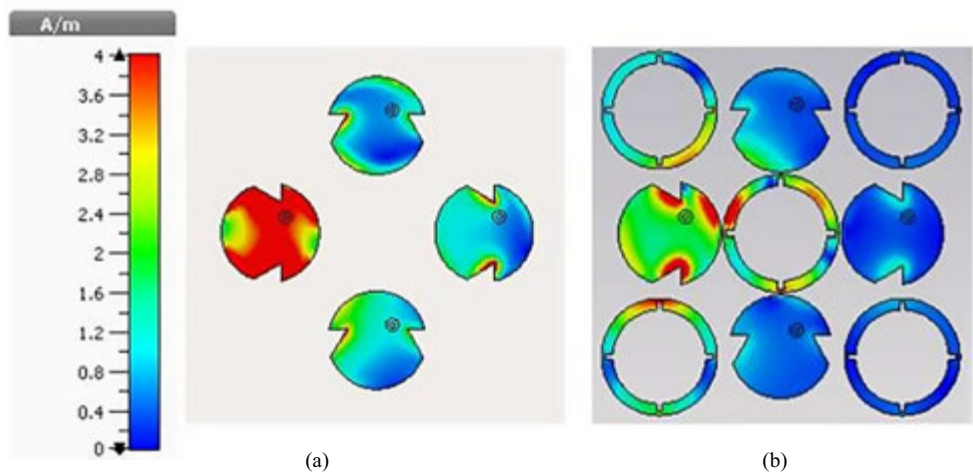


Figure 13.
Surface currents distribution at 8.3 GHz when port-1 is excited, (a) Without FSS, and (b) With FSS [44].

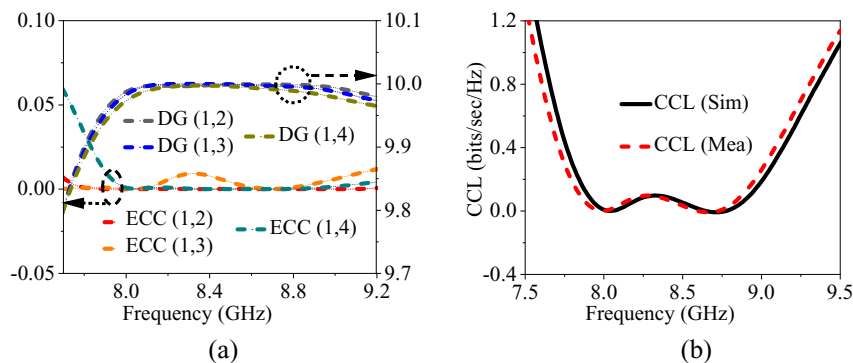


Figure 14.
(a) Simulated ECC and DG of the suggested antenna with FSS, and (b) Simulated and measured CCL [44].

1.4.2 Quad-Port MIMO antenna with superstrate

In this work, a wideband rectangular microstrip antenna is developed and enhanced through strategically introduced slots, which improve impedance matching and enable efficient operation from 3.4 to 5.4 GHz [45]. This single-element design is then extended into a four-element MIMO configuration while essentially preserving its original matching characteristics, demonstrating the robustness and adaptability of the proposed structure. The slot modifications play a central role in shaping the antenna's response, widening the usable bandwidth, and supporting its suitability for modern broadband MIMO applications.

To further suppress mutual coupling, a systematic metamaterial-based decoupling technique is employed. Using effective-parameter-derived metamaterial unit cells arranged in a 3×3 superstrate, the design achieves more than -23 dB isolation across all ports. In addition to significantly reducing coupling, the metamaterial layer improves the antenna's gain from 2.8 dBi to a peak of 5.3 dBi. The resulting system provides compact size, wideband operation, and a low-profile form factor, making it highly practical for Wi-Fi, WiMAX, and other contemporary wireless communication platforms that require enhanced isolation and diversity performance. The following section evaluates this effect by comparing mutual coupling levels and surface current distributions with and without the superstrate. The results confirm that the metamaterial loading significantly suppresses unwanted coupling and enhances overall MIMO antenna performance.

The metamaterial unit-cell method effectively enhances MIMO isolation by exploiting engineered electromagnetic properties to suppress dominant coupling paths. The optimized structure in **Figure 15(a)** is explicitly designed to minimize inter-element interaction. **Figure 15(b)** shows the compact 3×3 superstrate, built on a 1.6 mm FR-4 substrate using truncated square and slotted circular rings for proper impedance matching. CST simulations with periodic boundaries and Floquet excitation validate its behavior. When placed above the MIMO array, the superstrate forms a high-impedance surface due to its negative permittivity and permeability, thereby reducing mutual coupling while maintaining high efficiency. As seen in **Figure 16(a)**, the unit cell produces a 4.0–4.4 GHz bandstop with a 4.1 GHz resonance, $|S_{21}| < -10$ dB, $|S_{11}| \approx 0$ dB, and the constitutive parameters in **Figure 16(b)** confirm its metamaterial characteristics. The theoretical analysis retrieves the

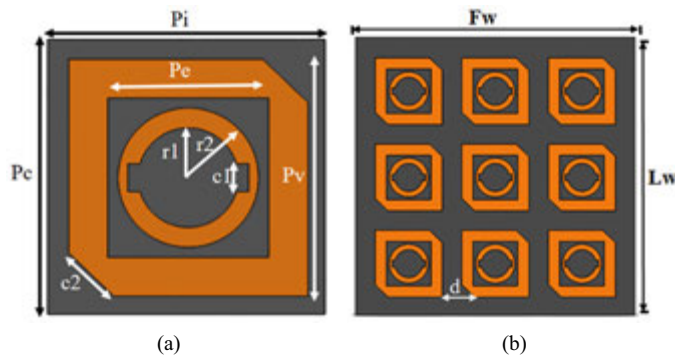


Figure 15.
(a) Metamaterial unit cell, and (b) Metamaterial superstrate [45].

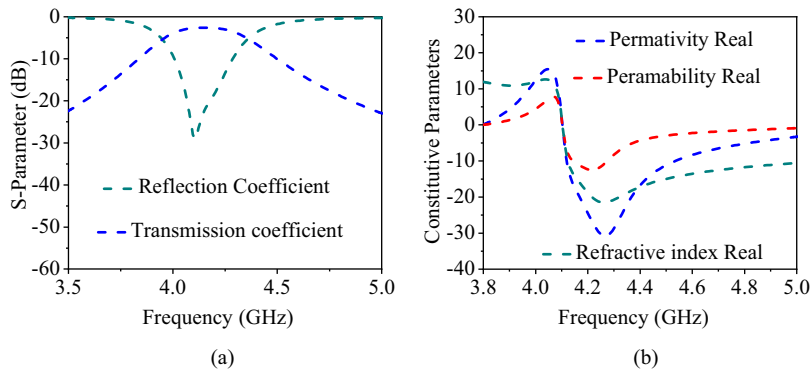


Figure 16.
 (a) Metamaterial unit cell S-parameters, and (b) constitutive parameters [45].

effective permittivity (ϵ), permeability (μ), and refractive index (n) from the unit cell's scattering parameters. The extracted values show negative ϵ and μ within the operating band, confirming that the proposed unit cell exhibits metamaterial behavior and functions as a potential double-negative (DNG) medium. These negative constitutive parameters reinforce the structure's ability to suppress electromagnetic coupling, thereby improving MIMO antenna isolation while enabling a more compact overall design.

As shown in **Figure 19(a)**, the antenna without decoupling operates from 3.2 to 5 GHz but exhibits poor mutual coupling, with isolation dropping below -20 dB (**Figure 19b**). To address this, a 3×3 metamaterial superstrate is placed $0.2\lambda_0$ above the compact $0.65\lambda_0 \times 0.65\lambda_0 \times 0.002\lambda_0$ four-port array (**Figure 17a, b**). The superstrate modifies the electromagnetic environment, suppressing both surface- and space-wave coupling, thereby enhancing isolation. Although the reflection bandwidth remains essentially unchanged, the $|S_{11}|$ resonance shifts toward 4 GHz. Without the superstrate, the worst coupling values are $|S_{12}| \approx -17$ dB and $|S_{14}| \approx -16$ dB, whereas with the superstrate, all ports achieve a minimum isolation of more than 23 dB and maintain at least 20 dB across the band. The measured S-parameters in

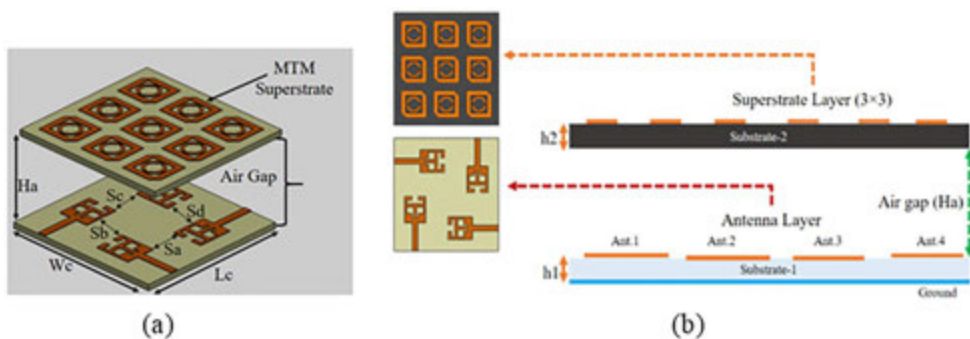


Figure 17.
 Prospective view of the suggested MIMO antenna with superstrate, and (b) Side view of the proposed antenna with superstrate [45].

Figure 19(c) agree well with simulations, with minor deviations due to fabrication tolerances. **Figure 18** shows the measurement setup.

Figure 20(a) shows that the metamaterial superstrate significantly enhances antenna gain. Without the superstrate, the MIMO antenna achieves an average gain of 2.5 dB and a peak of 3.1 dB, whereas adding the superstrate raises the average gain above 3.5 dB and increases the peak to 5.3 dBi at 4.8 GHz, an improvement of roughly 2.5 dB due to the resonant cavity formed between the antenna and the superstrate. The measured gain follows the same trend, only about 0.5 dB lower than the simulations. As illustrated in **Figure 20(b)**, the superstrate also boosts radiation efficiency by approximately 5%.

Figure 21 presents the co- and cross-polarization radiation patterns of the antenna in the H- and E-planes at 4 GHz, with Port 1 excited and the remaining ports terminated with 50 Ω loads. The antenna maintains an almost omnidirectional pattern with slightly reduced side and back lobes, both with and without the superstrate. The cross-polarization levels remain low and change only marginally after superstrate insertion, while the co-polarized main lobe shows a slight improvement. Because poor isolation often distorts radiation patterns in closely

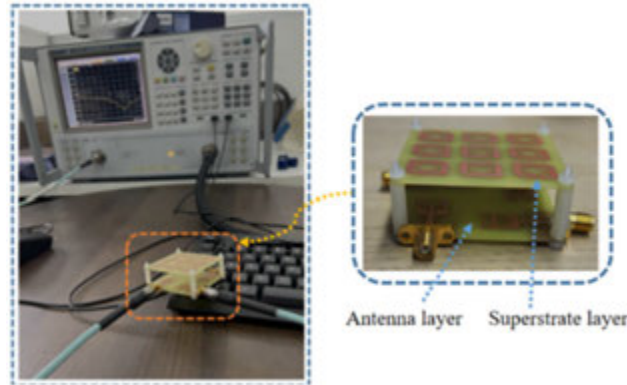


Figure 18.
The proposed antenna measurement setup with superstrate [45].

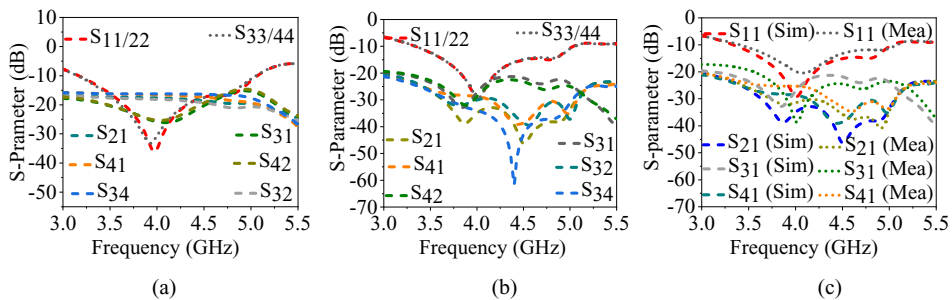


Figure 19.
(a) *S*-parameter without superstrate, (b) *S*-parameter with superstrate, and (c) Simulated and measured *S*-parameter with superstrate [45].

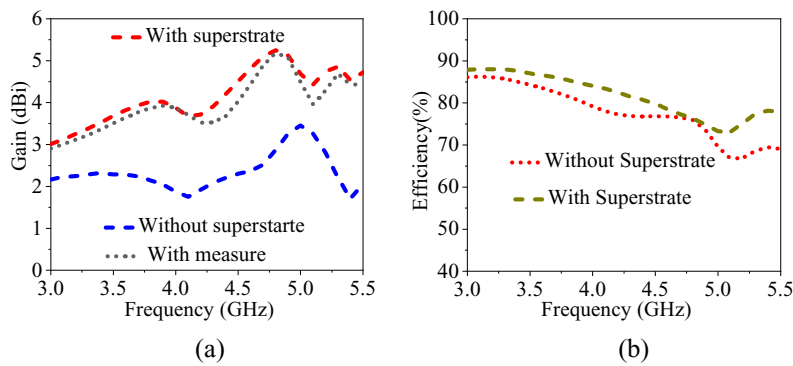


Figure 20. Gain and efficiency of antenna with and without superstrate, (a) Gain, and (b) Efficiency [45].

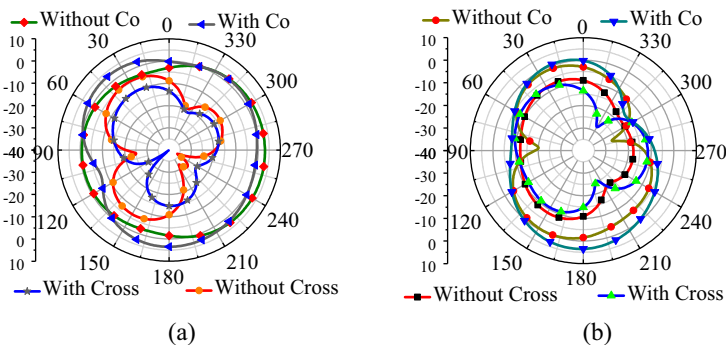


Figure 21. Simulated radiation pattern with and without superstrate at 4 GHz, (a) E-plane, and (b) H-plane [45].

spaced arrays, the stability of these patterns confirms the effectiveness of the decoupling design.

Figure 22 compares the 4 GHz surface current distribution with and without the superstrate. Without the superstrate (**Figure 22a**), the excitation of Port 1 produces substantial current leakage onto adjacent elements, indicating high mutual coupling. With the superstrate added (**Figure 22b**), these unwanted currents are significantly reduced, and **Figure 22(c)** shows that the driven-element current becomes confined within the metamaterial unit cells. The superstrate converts part of the surface waves into evanescent waves and absorbs them through its optimized unit-cell geometry, preventing their propagation to nearby elements. Combined with the orthogonal radiator arrangement, this mechanism effectively blocks current flow between antennas, thereby significantly enhancing isolation.

The superstrate ECC is less than 0.001, which is less than the 0.5 wireless system minimum, as shown in **Figure 23(a)**. **Figure 23(b)** indicates that the CLL is less than 0.32 bits/sec/Hz across the entire operating band with the superstrate, which is acceptable in a MIMO system. **Figure 23(b)** shows that the TARC is less than -10 dB within the working frequency range with the superstrate.

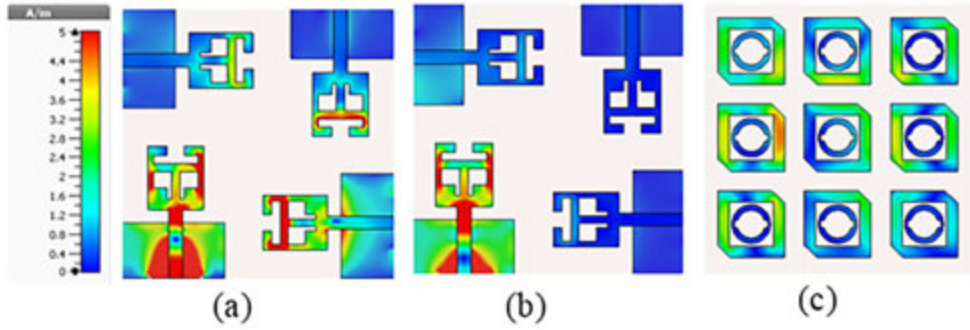


Figure 22. Current distribution of the planned MIMO antenna at 4 GHz, (a) Without superstrate, (b) With a superstrate antenna layer, and (c) Superstrate layer [45].

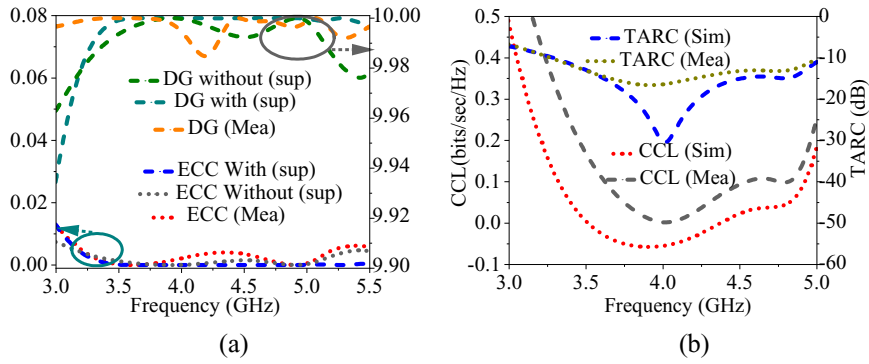


Figure 23. Simulated and measured diversity parameters of the proposed antenna, (a) ECC and DG, and (b) CCL and TARC [45].

1.4.3 Parasitic and DGS CP MIMO antenna

This work presents a compact hybrid CP MIMO antenna for mm-wave bands, operating from 23.34 to 28.77 GHz [46]. Rectangular and G-shaped slots in the radiators enable a wide impedance bandwidth and generate circular polarization. Four orthogonal elements are integrated with a Jerusalem Cross Defected Ground (JCDG) and a Hexagonal Ring Hybrid-Coupler Parasitic (HRHCP) structure, forming a $35 \times 35 \times 0.508 \text{ mm}^3$ array ($2.91\lambda_0 \times 2.91\lambda_0 \times 0.0423\lambda_0$). The combined JCDG-HRHCP technique reduces mutual coupling below -29 dB while enhancing gain, isolation, and axial-ratio bandwidth. The antenna achieves $|S_{11}| < -10 \text{ dB}$ over a 20.4% bandwidth, an 8.5% AR bandwidth, and peak gains of 6.4 and 7.2 dBi. Simulated, measured, and circuit-model results closely agree, validating the design. Despite its compact size, the array delivers intense CP and diversity performance for mm-wave applications. Parasitic resonators further improve decoupling by creating a tunable secondary coupling path that cancels the original coupling through destructive interference. This is more challenging in CP antennas because both co- and cross-polarized coupling components must be controlled.

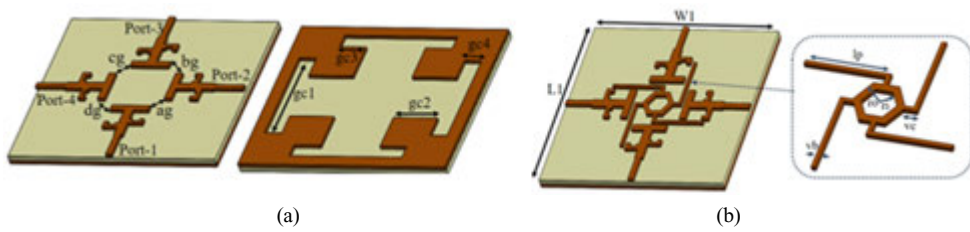


Figure 24.

(a) The proposed 3-D view of the MIMO CP antenna with decoupling ground, (b) the MIMO antenna with parasitic element [46].

Figure 24 shows the proposed four-port MIMO antenna, where four identical radiators are printed orthogonally on a Rogers RT/duroid 5880 substrate ($\epsilon_r = 2.2$, $\tan\delta = 0.0009$, $h = 0.508$ mm). The compact array measures $35 \times 35 \times 0.058$ mm³ and is excited using a 50 Ω microstrip feed. Although orthogonal placement helps reduce coupling, the minimal edge-to-edge spacing still results in strong mutual coupling and inadequate isolation. To overcome this limitation, a hybrid geometric and excitation-based decoupling strategy is introduced. **Figure 24** shows the antenna integrated with the proposed decoupler. The four radiators maintain an edge-to-edge spacing of $0.198\lambda_0$ at 25.2 GHz, making effective decoupling essential for improving overall MIMO performance.

The proposed MIMO antenna employs slotted radiators, a backside decoupling ground, and front-side parasitic elements positioned between closely spaced radiators. Without these features, the minimal inter-element spacing leads to strong mutual coupling. Isolation is first improved through a JCDG, created by etching vertical and horizontal slots in the ground plane to suppress surface-wave propagation and expand the impedance bandwidth. In the second stage, an HRHCP element is added between the radiators to enhance isolation further and generate circular polarization. The combined action of the patch slots, JCDG, and hybrid parasitic structure yields a compact CP MIMO antenna with a wider operating bandwidth and significantly reduced mutual coupling. Comparisons with configurations lacking these decoupling mechanisms confirm the effectiveness of both ground-based and parasitic approaches. **Figure 25** shows the fabricated prototype and measurement setup of the decoupled CP MIMO antenna.

Figure 26(a) shows the simulated reflection coefficients of the CP MIMO antenna. Because the elements are orthogonally arranged, $|S_{11}| \approx |S_{33}|$ and $|S_{22}| \approx |S_{44}|$. Without a decoupler, the antenna covers only 23.4–26.7 GHz, missing the full 23.34–28.67 GHz range of the single radiator. Adding the JCDG improves the bandwidth to 23.5–28 GHz, but full coverage is achieved only when the HRHCP element is included, indicating the combined effectiveness of both structures. **Figure 26(b)** displays the corresponding mutual coupling behavior with and without the JCDG. Due to the orthogonal element placement, the coupling terms exhibit symmetry: $|S_{21}| \approx |S_{23}| \approx |S_{32}|$ and $|S_{41}| \approx |S_{43}| \approx |S_{34}|$. Without a decoupler, the worst isolation values, $|S_{12}| \approx -16.5$ dB and $|S_{41}| \approx -17.3$ dB, remain below 20 dB over 24.5–29 GHz, while $|S_{31}|$ is about -20.3 dB. With the JCDG applied, the coupling improves substantially, with $|S_{21}|$, $|S_{31}|$, and $|S_{171}|$ all better than -23 dB across the band. **Figure 27** demonstrates the significant improvement in isolation achieved by adding the HRHCP parasitic element. With this element in place, $|S_{12}|$, $|S_{13}|$, and $|S_{14}|$ all improve to better than -29 dB, as shown in

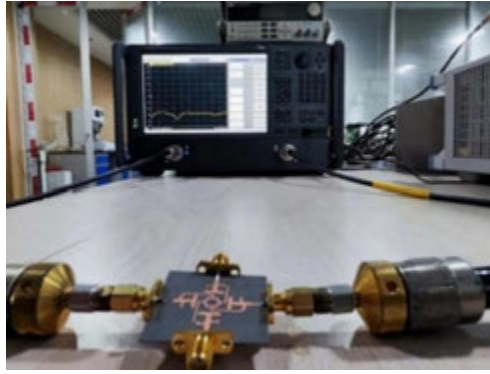


Figure 25.
Fabricated photo of the proposed antenna, measurements setup [46].

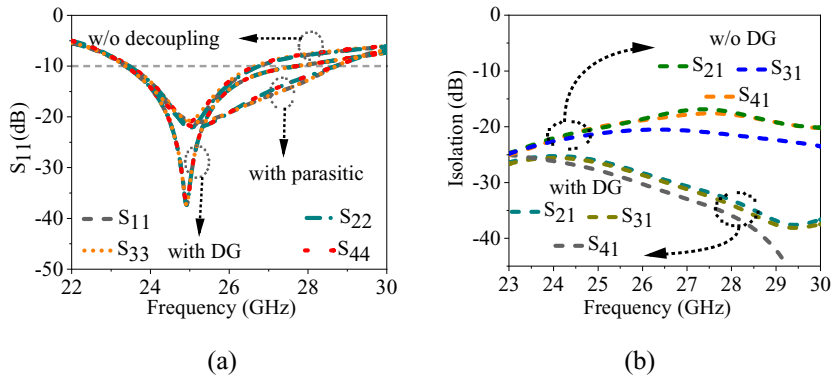


Figure 26.
Simulated S -parameters of the MIMO antenna with and without decoupling, (a) Reflection coefficient and (b) Isolation [46].

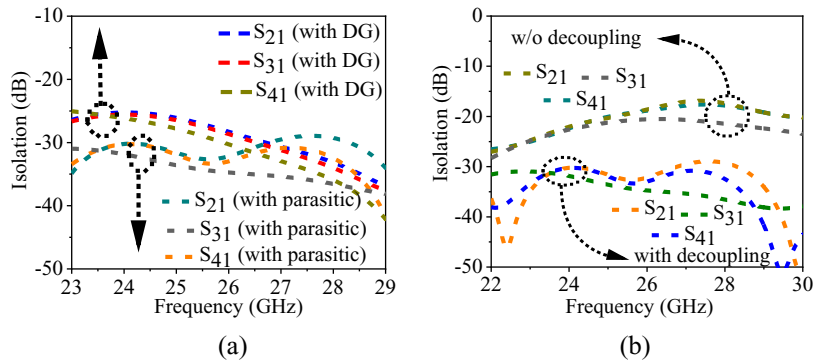


Figure 27.
(a) Simulated isolation with DG and parasitic element, and (b) Isolation without and with decoupler [46].

Figure 27(a). While the JCDG alone provides about a 4 dB reduction in coupling, combining it with the HRHCP yields nearly a 12 dB improvement across the band, as illustrated in **Figure 27(b).** **Figure 28** shows strong agreement between the simulated and measured S-parameters, with $|S_{11}|$ below -10 dB from 23.4 to 28.6 GHz and measured coupling remaining below -25 dB. Minor discrepancies are attributed to fabrication tolerances and measurement losses.

The simulated AR for the suggested MIMO antennas, with and without a decoupler, is shown in **Figure 29(a).** The AR of the recently proposed antenna presents a comparative analysis of its performance with and without the hybrid approach. As elucidated in the earlier section, the MIMO antenna under consideration exhibits circular polarization (CP) properties, achieved by incorporating multiple slits into the single proposed radiator. When coupled with the decoupling ground, there is a variation in the AR within specific frequency ranges. The MIMO antenna without a decoupler has a simulated axial-ratio bandwidth (ARBW) that remains above 3 dB over the operating frequency, indicating the antenna is linearly polarized. Nevertheless, after inserting the parasitic decoupler, the alteration is within 3 dB. This observation suggests that the MIMO antenna retains its CP properties. The proposed MIMO antenna achieves a simulated 3-dB axial-ratio bandwidth of 2.4 GHz (26.1–28.5 GHz). As shown in **Figure 29(a)**, the measured ARBW of 26.1–28.1 GHz is only about 0.4 GHz lower than the simulation,

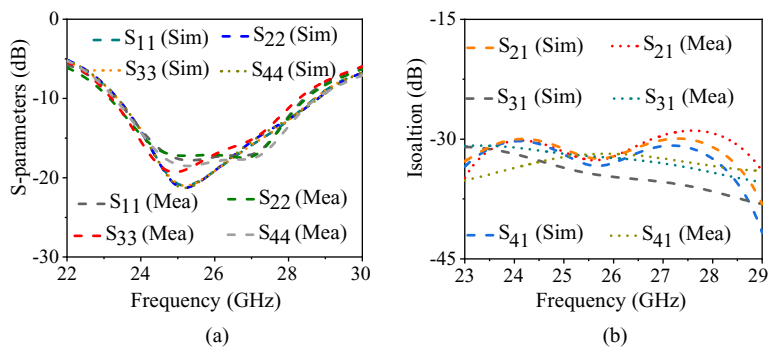


Figure 28. Simulated and measured S-parameters of the MIMO antenna, (a) Reflection coefficient, and (b) Isolation [46].

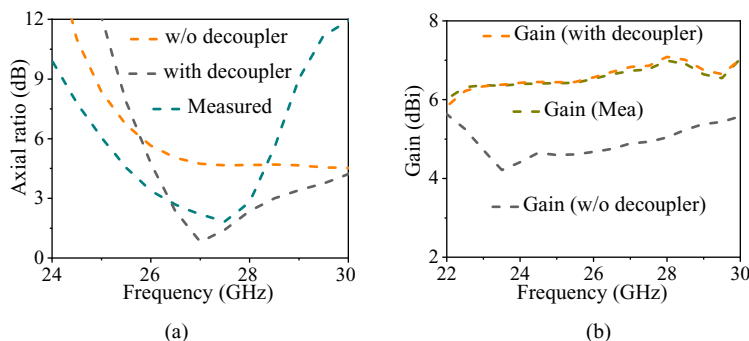


Figure 29. (a) Simulated and measured axial ratio and (b) Simulated and measured gain [46].

indicating good agreement. **Figure 29(b)** shows that the antenna's peak gain is below 4 dBi without a decoupler but rises to 7.2 dBi at 28.2 GHz after adding the JCDG and HRHCP elements, due to the introduction of an additional resonance. The measured gain closely matches the simulated results, confirming the effectiveness of the hybrid decoupling technique.

1.4.4 Multiband MIMO antenna integrates Carrel's log-periodic concept with the planar log-periodic array mechanism

The study indicated that a MIMO antenna can function at elevated frequencies for Ku/K-band applications and possesses significant potential for enhanced wireless communications. Antenna elements arranged orthogonally are constructed in an anti-parallel manner with offset ports to maximize diagonal spacing when implemented, hence minimizing mutual interaction among MIMO antenna components [47]. This approach was founded on the resonant characteristics of creative elements and the function of the MIMO radiator, alongside various roles of the branches, which were suggested to improve antenna diversity and design. Furthermore, the experimental research allegedly employed a generic ground plane for the projective MIMO antenna classification, which consistently maintained uniform voltage across the antenna components and ensured operation for wireless communication. All MIMO performance was experimentally tested and evaluated, and the resulting data were analyzed to yield optimal simulation performance.

Figure 30(a) illustrates the proposed fork-shaped MIMO antenna design, fabricated on a 1.58 mm thick RT-5880 substrate with a permittivity of 2.2 and a loss tangent of 0.0009, thereby minimizing dielectric losses and maximizing efficiency. The overall dimensions of the two-port MIMO prototype are $54 \times 44 \text{ mm}^2$. The

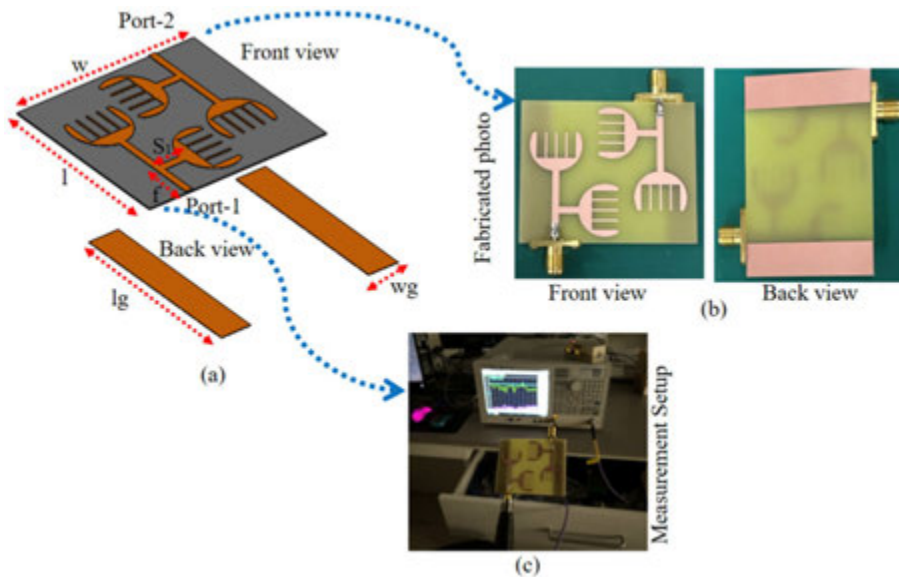


Figure 30. (a) Proposed fork-shaped MIMO antenna 3-D structure, (b) Fabricated front and back view, (c) Measurement setup [47].

antenna incorporates two 50- Ω feed sites offset diagonally to maximize the effective radiation area and employs two orthogonal branches to create a 2×4 MIMO configuration. This arrangement facilitates dual-beam radiation emission across horizontal and vertical planes, enhancing diversity performance by simultaneously supporting horizontally and vertically polarized signals.

The fork-shaped radiators are logarithmically scaled to achieve compact size, wideband behavior, and efficient multi-band operation. The reflection coefficient $|S_{11}|$ and coupling $|S_{21}|$ confirm excellent impedance matching and strong isolation, with $|S_{21}|$ consistently below -20 dB. The antenna also delivers high gain and radiation efficiency above 80%, ensuring reliable transmission and reception with minimal signal loss. This makes the design well-suited for 5 G communications, satellite links, and radar systems that require high gain, multi-band operation, and stable performance. **Figure 30(b, c)** shows the fabricated prototype and the measurement setup of the proposed fork-shaped MIMO antenna.

Figure 31 shows that the proposed two-port MIMO antenna operates effectively across multiple bands: C-band (4.41–5.21 GHz), Ku-band (14.30–17.64 GHz), K-band (20.10–28.10 GHz), Ka-band (28.95–30.76 GHz, 32.38–36.42 GHz), and part of the Q-band (38.34–39.5 GHz). Compared to the single-element antenna, the two-port configuration exhibits small frequency shifts, mainly at lower bands, due to port interactions and inter-element coupling. These shifts, however, are minor, and the MIMO antenna maintains the wide multi-band behavior of the single radiator. Notably, the design achieves excellent isolation, maintaining a minimum of 21.54 dB across all operating bands without requiring additional decoupling structures, demonstrating the efficiency and robustness of the proposed MIMO architecture.

Adding two orthogonal ports enhances diversity and improves the far-field performance of the MIMO configuration. This pattern diversity allows the antenna to handle both horizontal and vertical polarizations while maintaining strong isolation. Consequently, the two-port antenna retains the wide operating bands of the single element and offers improved isolation and better overall MIMO performance. **Figure 32(a, b)** presents the $|S_{11}|$, $|S_{22}|$, $|S_{21}|$, and $|S_{12}|$ responses, showing excellent agreement between simulation and measurement. The close match in $|S_{11}|$ confirms the design and fabrication precision, ensuring reliable performance across all bands.

The $|S_{11}|$ response stays below -10 dB across all operating bands, confirming good impedance matching throughout the C-, Ku-, K-, Ka-, and Q-bands. The $|S_{21}|$

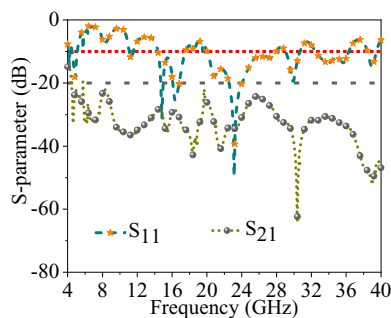


Figure 31.
 Proposed two-port fork-shaped MIMO antenna reflection coefficient and transmission coefficient [47].

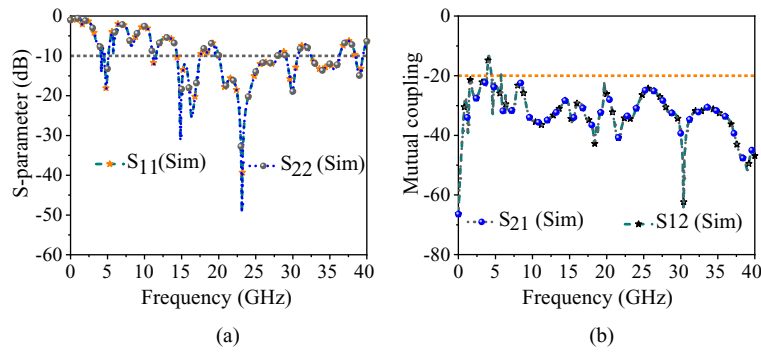


Figure 32.

(a) Proposed two-port fork-shaped MIMO antenna reflection coefficient $|S_{11}|$ and $|S_{22}|$, (b) Transmission coefficients $|S_{12}|$ and $|S_{21}|$ [47].

transmission coefficients also maintain strong isolation, consistently below -20 dB, as shown in **Figure 32(b)**, ensuring reliable MIMO performance. Measured results closely match simulations, with only small deviations more noticeable at higher frequencies due to slight resonance shifts caused by fabrication and environmental factors (**Figure 33**). Overall, the measurements validate the antenna's wideband, multi-band operation and robust isolation characteristics.

Figure 34(a) shows that the antenna gain increases with frequency, featuring distinct peaks at its resonant bands. Gain is moderate at low frequencies (e.g., 5 GHz), rises significantly at mid-band frequencies (16 and 24 GHz), and remains high and stable at 29.8, 34.4, and 38.9 GHz, demonstrating strong suitability for Ku-, K-, Ka-, and Q-band applications. **Figure 34(b)** shows that the total efficiency remains above 75% across most of the operating range, with small drops at lower frequencies due to substrate losses. The high mid- and high-band efficiency confirms the antenna's optimized radiation behavior. Overall, the results validate the design for multi-band applications such as 5G, satellite links, and radar systems that require high gain and robust efficiency.

Figure 35(a) shows that the ECC values, calculated from far-field patterns, remain very low, indicating weak pattern correlation and substantial diversity.

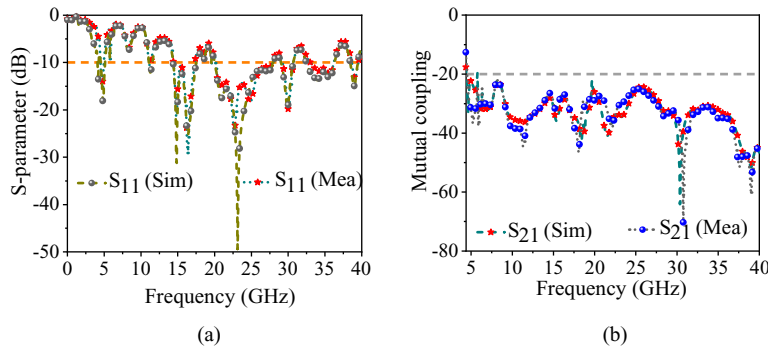


Figure 33.

Proposed two-port fork-shaped MIMO antenna: simulated and measured S -parameters. (a) Reflection coefficient, (b) Transmission coefficient [47].

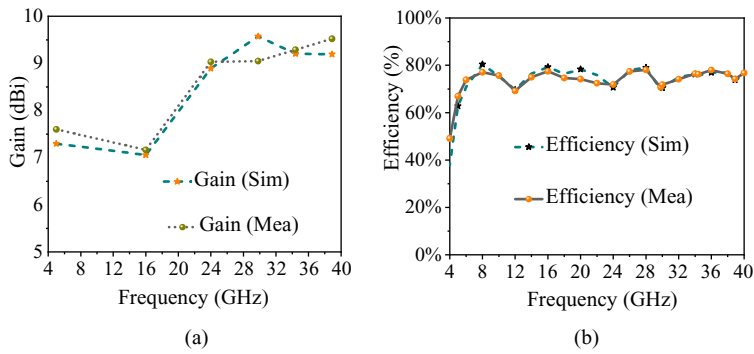


Figure 34.
 Proposed two-port fork-shaped MIMO antenna gain and efficiency, (a) Gain, (b) Total efficiency [47].

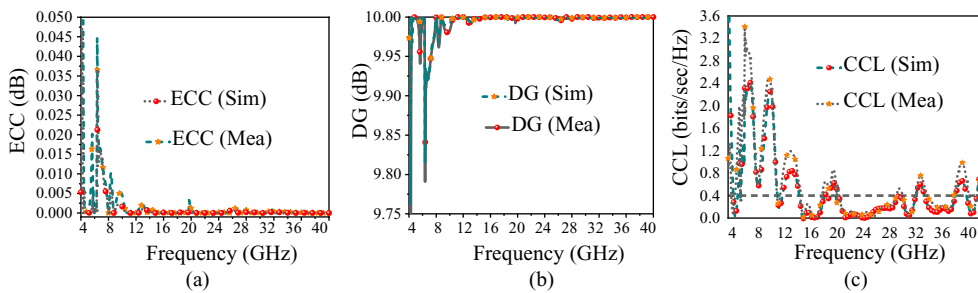


Figure 35.
 Proposed two-port fork-shaped MIMO antenna diversity performance, (a) ECC, (b) DG, (c) CCL [47].

The fork-shaped geometry helps minimize pattern overlap, keeping ECC low in both simulation and measurement. As seen in **Figure 35(b)**, DG exceeds 9.99 dB whenever ECC is minimal, confirming excellent MIMO performance. **Figure 35(c)** shows that CCL remains below 0.3 bits/s/Hz, well below the 0.4 bits/s/Hz threshold, demonstrating high channel efficiency across all resonant bands.

1.4.5 Dual technique MIMO antenna design

This study presents an innovative dual-technique circularly polarized (CP) MIMO antenna for mm-wave IoT applications. The design integrates four rectangular slotted radiators with a hexagonal nested neutralization line (HN-NL) and a circular segmented defected ground (CS-DG), forming a compact $30 \times 30 \times 0.508 \text{ mm}^3$ structure ($2.80\lambda_0 \times 2.80\lambda_0 \times 0.0457\lambda_0$ at 28 GHz). The combined HN-NL and CS-DG network provides effective decoupling, enhancing both axial-ratio bandwidth and inter-port isolation. The antenna achieves a 3-dB AR bandwidth of 27.21–32.83 GHz (20.67%), mutual coupling better than -28 dB , an impedance bandwidth of 25.12–34.5 GHz (30.96%), and peak gains of 7.25 dBi at the lower band and 8.13 dBi at the upper band. With its compact form, low profile, and intense CP and MIMO performance, the antenna is well-suited for modern mm-wave IoT systems [48].

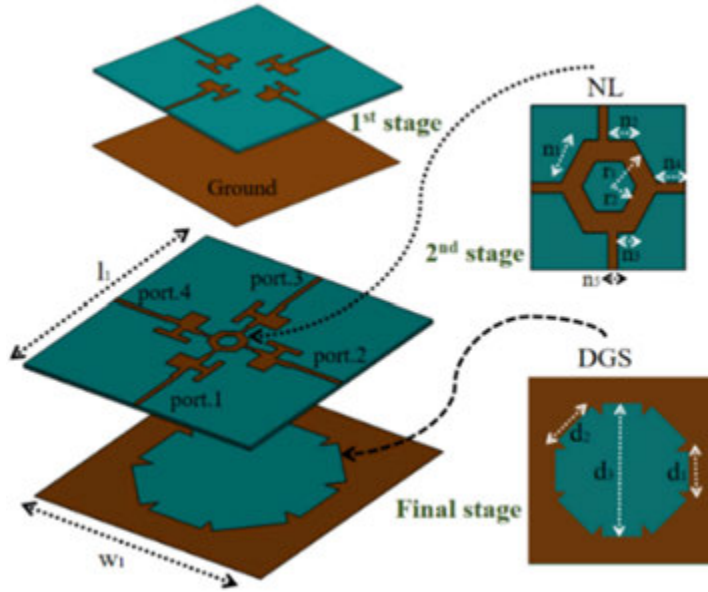


Figure 36.
Proposed 3-D MIMO antenna evaluation [48].

Figure 36 illustrates the proposed four-port MIMO antenna, which employs slotted radiators along with HN–NL and CS–DG decoupling structures. Without these mechanisms, the closely spaced radiators exhibit strong mutual coupling, reduced bandwidth, and AR values exceeding 3 dB. The second design stage introduces the HN–NL, whose conductive loops generate countercurrents that cancel coupled fields, significantly improving isolation and bandwidth. In the third stage, a CS–DG is incorporated to suppress surface waves further and redistribute surface currents, providing additional coupling reduction. The CS–DG enhances impedance bandwidth by about 5% and expands the AR bandwidth by roughly 13% below 3 dB. In combination, the HN–NL and CS–DG form an effective dual-stage decoupling and CP-enhancement strategy, enabling a compact and high-performance MIMO-CP antenna.

Figure 37(a) presents the simulated reflection coefficients of the antenna with and without the proposed decoupling structures. As expected from the symmetrical layout, $|S_{11}|$, $|S_{22}|$, $|S_{33}|$, and $|S_{44}|$ exhibit similar behavior. Without any decoupling, the antenna covers 26.46–33.17 GHz. Introducing the HN–NL expands the bandwidth to 25.95–33.83 GHz, and adding the CS–DG further broadens it to 25.14–34.03 GHz, confirming that both structures enhance the overall mm-wave impedance bandwidth. **Figure 37(b)** shows that, without decoupling, the mutual coupling $|S_{ij}|$ remains below –20 dB, indicating moderate isolation. Incorporating the HN–NL improves coupling by about –25 dB on average, creating resonant cancellation paths that suppress inter-element interference.

The integration of the CS–DG enhances performance, resulting in average MCs of –30 dB for $|S_{12}|$, $|S_{13}|$, $|S_{14}|$, $|S_{43}|$, $|S_{42}|$, and $|S_{23}|$. The CS–DG improves performance by interrupting ground plane surface currents, attenuating surface waves, and incorporating bandgap effects. These processes systematically reduce MC and

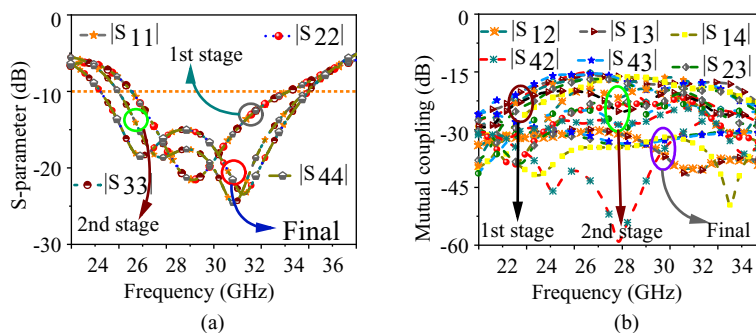


Figure 37. Simulated S -parameters of the MIMO antenna without and with DG. (a) Reflection coefficient, (b) Mutual coupling [48].

increase BW, making the antenna array suitable for high-performance applications. **Figure 38** depicts the fabricated and measured setup of the MIMO-CP antenna. **Figure 39(a)** compares the measured and simulated reflection coefficient, which is nearly identical. The measured MC level is consistently below -25 dB across the operating frequencies, as depicted in **Figure 39(b)**.

Figure 40 shows the antenna vector current distribution, which is key to MIMO-CP antenna performance. When port-1 is excited with $\sqrt{50}$ volts to produce 1 W of incident power, the remaining ports are terminated at the operational frequency, with the ports oriented at 0° , 90° , 180° , and 270° . The current flow, with opposing phase orientation angles and a constant magnitude, demonstrates the CP antenna properties. In **Figure 40**, the LHCP is depicted clockwise at the working frequencies. The AR of the suggested antenna, with and without a hybrid technique, is shown in **Figure 41(a)**. Initially, AR simulations of the MIMO antenna without decoupling are done using LP. With the insertion of the HN-NL decoupling, the alteration is covered from 30.75 to 33.84 under 3 dB. This observation suggests that the antenna preserves its CP properties. With CS-DG insertion, the AR frequency range is from 27.21 to 32.86 GHz, corresponding to a 3 dB ARBW of 5.62 GHz. The AR curves were recorded at 0° in the boresight direction. **Figure 41(b)** shows that the simulated and measured ARBW are almost identical, with slight differences.

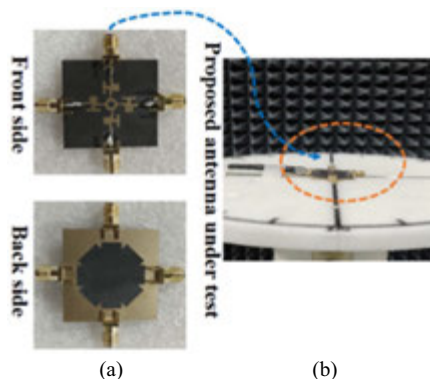


Figure 38. (a) Fabricated photo of the suggested antenna, (b) Measurement [48].

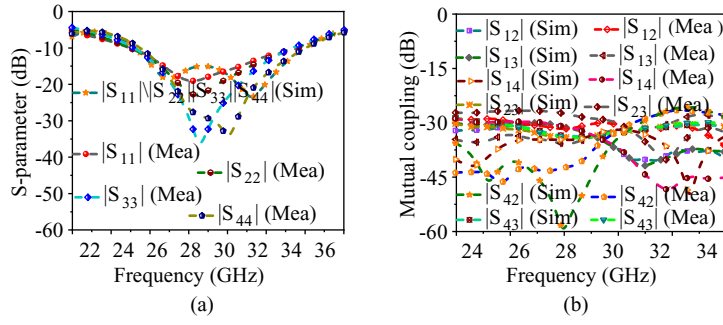


Figure 39. Simulated and measured S -parameters of the MIMO antenna. (a) Reflection coefficient. (b) MC [48].

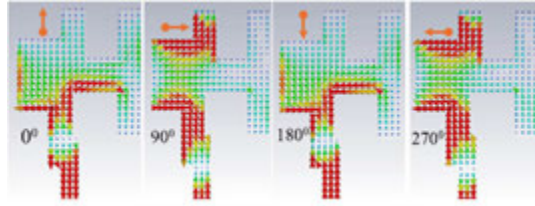


Figure 40. CP current vector distributions at different time phases [48].

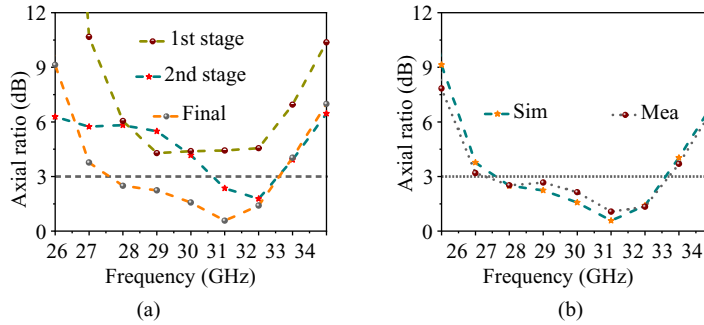


Figure 41. (a) Simulated AR with and without a decoupler. (b) Measured and simulated AR [48].

Figure 42(a, b) illustrates the RHCP and LHCP radiation of the proposed antenna, along with the analysis of radiation patterns for the $\phi = 0^\circ$ and 90° planes. The patterns exhibit quasi-omnidirectional behavior, with some concavity at 28 GHz, covering all principal planes. LHCP radiation exhibits greater boresight intensity than RHCP radiation, thereby enhancing CP. In boresight, LHCP radiation exceeds RHCP by more than 20 dB. **Figure 42(c)** shows the MIMO antenna gain with decoupling. When various ports are excited, the decoupled MIMO antenna's peak gain is 7.25 dBi at lower frequencies and 8.13 dBi at higher frequencies. **Figure 42(c)** also shows the measured and simulated efficiencies of the MIMO antenna, with an average above 85%.

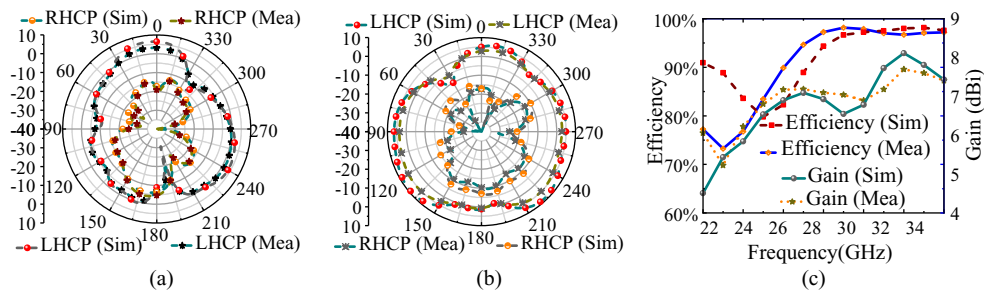


Figure 42. Simulated and measured RHCP and LHCP radiation patterns. (a) 0° , (b) 90° . (c) Simulated and measured efficiency and gain [48].

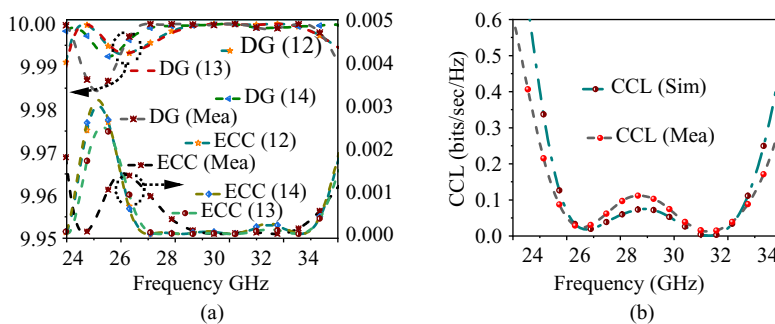


Figure 43. (a) Simulated ECC and DG of MIMO antenna. (b) Measured and simulated CCL [48].

MIMO performance was evaluated using ECC, DG, and CCL. The ECC, computed from measured far-field patterns, remains extremely low, with a maximum value of 0.001 (**Figure 43a**), indicating excellent pattern diversity. The corresponding DG reaches 9.99 dB, confirming strong fading resistance. Based on measured S-parameters, the CCL stays below the 0.4 bits/s/Hz limit (**Figure 43b**), demonstrating minimal channel degradation and high MIMO efficiency.

2. Summary

This chapter presents a comprehensive overview of mutual coupling suppression in MIMO antenna systems using artificial electromagnetic structures (AEMs). As modern wireless platforms, ranging from sub-6 GHz to WLAN and mmWave 5G, continue to demand higher data rates, greater spectral efficiency, and compact integration, effective decoupling techniques have become essential. Mutual coupling arising from near-field interactions, surface-wave propagation, ground-plane currents, and space-wave overlap degrades key MIMO metrics, including ECC, DG, CCL, radiation efficiency, and overall link reliability.

The chapter first establishes the electromagnetic and mathematical foundations for analyzing MC, including impedance and scattering matrices, far-field correlation

integrals, and surface-wave models. These tools emphasize the importance of identifying and suppressing dominant coupling pathways. AEM structures such as FSS, metamaterials, parasitic resonators, neutralization lines, and DGS are shown to be highly effective at controlling these paths through engineered electromagnetic responses.

A comparison with existing work demonstrates that the proposed designs provide an excellent balance of isolation, fabrication simplicity, radiation stability, and PCB compatibility. Unlike many complex multilayer EBG, SIW, or metasurface-based solutions, the AEM-driven approaches remain planar, cost-effective, and suitable for mm-wave fabrication tolerances.

Looking ahead, AEMs will play an increasingly important role in emerging technologies such as 6 G, terahertz communication, reconfigurable intelligent surfaces (RIS), ultra-massive MIMO, and integrated sensing-and-communication (ISAC). Concepts such as cavity resonance control, bandgap formation, destructive current cancellation, and hybrid dual-domain decoupling offer valuable foundations for future research.

In conclusion, this chapter has shown that well-designed artificial electromagnetic structures provide a practical and versatile solution for suppressing MC in compact MIMO antennas. Through theory, simulation, and measurement, AEM-based techniques were demonstrated to significantly improve isolation and overall system performance across multiple bands and environments, supporting the continued evolution of next-generation wireless systems.

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