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Ph. D. Thesis

TITLE: Theory and Design of Smart RF Components Using the Novel FPMS Technology

Author:

SHAHINSHAH ALI

Supervisor:

Dr. FARHAN ABDUL GHAFAR

Internal committee members:

Dr. YUSHI ZHOU

Dr. CARLOS CHRISTOFFERSEN

Dr. LANGIS ROY

External Examiner:

Dr. YING WANG



A report submitted to the Lakehead University in accordance with the requirements of the degree of DOCTOR OF PHILOSOPHY in the Electrical and Computer Engineering.

ACKNOWLEDGEMENTS

First and foremost, I am deeply thankful to Almighty God for granting me the strength, patience, and perseverance to complete this PhD journey. His blessings gave me the courage to overcome the challenges encountered throughout this research and to continue moving forward during the most demanding stages of my studies. I would like to express my sincere gratitude to my supervisor for his continuous guidance, encouragement, and valuable support throughout the course of this research. His technical insight, constructive feedback, and confidence in my abilities played a significant role in shaping this work and in my development as a researcher. I am equally thankful to my committee members for their valuable comments, suggestions, and support, which helped improve the quality of this thesis. I am especially grateful to my family for their unconditional love, patience, sacrifices, and constant encouragement. Their support has been a source of strength throughout this long journey. Their belief in me, particularly during difficult and uncertain times, gave me the motivation to remain focused and complete this important milestone.

ABSTRACT

The increasing demand for intelligent wireless communication systems has accelerated the development of reconfigurable microwave components capable of adapting their electromagnetic characteristics in real time. Modern applications, including phased-array antennas, satellite communications, radar systems, fifth-generation (5G) and emerging sixth-generation (6G) wireless networks, require compact, low-loss, and electronically reconfigurable microwave devices that can dynamically control the radiation characteristics of antennas. Conventional approaches generally rely on phased arrays with complex feeding networks or transmission-line-based phase shifters, which often suffer from increased circuit complexity, larger physical size, higher insertion loss, and significant power consumption. Consequently, there is a growing need for alternative reconfigurable technologies that simplify system architecture while maintaining high performance.

This thesis investigates the concept of the Field Programmable Microwave Substrate (FPMS) as a new platform for realizing reconfigurable microwave components through dynamic control of the effective permittivity of the propagation medium. The research demonstrates two major applications of the FPMS concept. The first contribution presents a programmable artificial magnetic conductor (AMC) integrated with a single antenna element to achieve electronic beam steering without employing a conventional phased-array architecture. The proposed FPMS-AMC simultaneously functions as a reflective surface that suppresses backside radiation while dynamically controlling the reflection phase to steer the main beam. A beam steering range of approximately $\pm 30^\circ$ is achieved while maintaining a maximum realized gain of 7.2 dBi at 10 GHz. The fabricated prototype validates the proposed concept and demonstrates good agreement between simulated and measured results.

The second contribution introduces a continuously tunable phase shifter based on a programmable dielectric waveguide realized using FPMS technology. Unlike conventional phase shifters that employ switched transmission lines, loaded-line networks, or reflection-type circuits, the proposed design controls the propagation constant by electrically tuning the effective permittivity of the dielectric guiding region. An analytical model based on guided-wave theory is developed for both homogeneous and non-homogeneous dielectric configurations. The analytical predictions are verified through full-wave simulations and experimental measurements. The fabricated prototype demonstrates a continuous phase tuning range of approximately 300° at 2.5 GHz with good agreement between theoretical, simulated, and measured results.

Overall, this thesis establishes the FPMS as a versatile reconfigurable electromagnetic (EM) platform capable of controlling both radiation characteristics and guided-wave propagation through programmable dielectric properties. The proposed analytical framework, validated experimental results, and demonstrated antenna applications provide a foundation for future reconfigurable microwave and antenna systems requiring compact size, continuous tunability, and simplified beam steering architectures.

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Chapter 1

Introduction

The rapid advancement of wireless communication technologies has significantly increased the demand for intelligent and adaptable radio-frequency (RF) systems capable of supporting diverse applications under dynamic operating conditions. Emerging technologies such as fifth- and sixth-generation (5G/6G) mobile communications, satellite communications, Internet of Things (IoT), autonomous vehicles, unmanned aerial vehicles (UAVs), and modern radar systems require high data rates, reliable connectivity, low latency, and efficient spectrum utilization. These stringent requirements have driven the development of advanced antenna systems capable of dynamically adapting their radiation characteristics to varying propagation environments and user demands [1], [2], [3], [4].

For instance, 5G communication systems employ millimeter-wave frequency bands and massive multiple-input multiple-output (MIMO) architectures to achieve significantly higher data throughput than previous generations. However, operation at these high frequencies introduces severe propagation losses and increased susceptibility to blockage, making highly directive and electronically steerable antennas essential for maintaining reliable communication links. Similar challenges are expected to become even more pronounced in future 6G networks, where adaptive beamforming and intelligent radio environments are anticipated to become fundamental enabling technologies [1], [2]. Likewise, satellite communication systems are rapidly evolving from conventional geostationary satellite networks toward large constellations of Low Earth Orbit (LEO) satellites. Unlike geostationary satellites, LEO constellations provide lower communication latency, improved global coverage, and enhanced data throughput. However, maintaining continuous communication with fast-moving satellites requires antennas capable of electronically

steering their radiation beams without relying on bulky mechanical tracking mechanisms. Consequently, compact beam-steerable antennas have become an important area of research for next-generation satellite communication systems [5], [6].

Modern radar systems also impose increasingly demanding requirements on antenna technology. Applications including automotive radar, surveillance, weather monitoring, and military target tracking require rapid beam scanning, high angular resolution, and adaptive radiation control to accurately detect and track multiple moving targets in real time. Traditionally, these capabilities have been achieved using phased-array antennas. Like any other classical solution, phased array antennas have their challenges that include complex feeding networks, numerous phase shifters, and high-power consumption. However, with careful design innovation one can find methods that can reduce the effect of these downsides while providing stable radiation performance that can meet the stringent needs of modern wireless communication. This is the center around which the research of this thesis is built.

1.1 Motivation

One of the promising application areas for beam-steerable antennas is satellite communications, particularly small satellite platforms such as CubeSats [7]. CubeSats have emerged as a cost-effective solution for Earth observation, remote sensing, scientific missions, and defense applications owing to their compact size, reduced development cost, and rapid deployment capability. However, their small physical dimensions impose strict constraints on the antenna subsystem, requiring compact, lightweight, and low-power antenna designs while still maintaining reliable communication links over long distances. Furthermore, antenna and power subsystem failures remain among the leading causes of communication failure in CubeSat missions, highlighting the need for more efficient and adaptable antenna technologies [8].

In recent years, Canada has significantly increased its investment in LEO satellite technologies through initiatives led by the Defence Research and Development Canada (DRDC). One such initiative is the SpaceLEO program, which aims to establish a constellation of LEO satellites capable of supporting the applications pertaining to CubeSats. Compared with conventional geostationary satellites, LEO constellations offer significantly lower communication latency, higher throughput, and improved coverage, particularly in northern and Arctic regions where geostationary satellites experience limited visibility and long propagation delays [8]. However, maintaining reliable communication with rapidly moving LEO satellites requires antenna systems capable of electronically adapting their radiation characteristics without relying on bulky mechanical steering mechanisms.

Similarly, High-Throughput Satellites (HTSs) have become an essential component of next-generation satellite communication networks supporting aviation, maritime, broadband internet, and government communication services [10]. Modern HTS platforms employ multiple spot beams together with frequency reuse techniques to dramatically increase network capacity. Consequently, these systems increasingly rely on electronically steerable and frequency-reconfigurable antennas to dynamically allocate coverage and efficiently respond to changing traffic demands. For example, the Intelsat EpicNG satellite platform utilizes steerable spot beams operating in both X- and Ka-band frequencies to provide reliable, high-capacity connectivity for aviation, maritime, and defense applications while maintaining low latency and high spectral efficiency [11]. Both CubeSat and HTSs clearly demonstrate the need for intelligent antenna technologies that can fit in a small form factor. CubeSats require antennas capable of operating within severe size and power limitations while maintaining reliable long-distance communication. Likewise, HTSs platforms demand antennas that can dynamically modify their radiation

characteristics to optimize coverage, improve spectral efficiency, and support increasing communication capacity. These requirements provide a strong motivation for the development of intelligent antenna systems capable of providing higher gains together with electronic beam steering using compact and energy-efficient architectures. The work presented in this thesis is motivated by these challenges and investigates the use of Field Programmable Microwave Substrate (FPMS) technology to meet the moment and provide viable answers to some of these questions.

1.2 Challenges

Although some progress has been made in the development of reconfigurable microwave components and antenna systems that rely on FPMS technology, several technical challenges still remain that are worth investigating. Along these lines some of the challenges anticipated in this work are outlined below:

Challenge 1: Limited Understanding of the Electromagnetic Behavior of FPMS

FPMS has shown significant potential as a reconfigurable microwave platform with some of its initial RF demonstrations. However, its EM behavior has only been investigated under a limited range of excitation environments. Existing studies primarily focus on transmission-line or lumped-port excitation, whereas its response under periodic boundary conditions and plane-wave excitation remains largely unexplored. For an antenna environment the exploration of this technology in a periodic structure set up is quite important. The lack of investigation in this area shows that not much literature could be referred to when developing such a design. Therefore, this can be enlisted as one of the challenges due to the lack of published knowledge.

Challenge 2: Realization of Reconfigurable Dielectric Phase Control

Precise phase control is one of the fundamental requirements of modern electronically steerable antenna systems. Conventional phase shifters generally rely on switched transmission lines, semiconductor switches, ferrite materials, or MEMS devices, each having limitations in terms of circuit complexity, size, fabrication cost, or integration. FPMS offers an alternative mechanism by dynamically modifying the effective dielectric properties experienced by the propagating EM wave. However, realizing a dielectric phase shifter capable of providing a wide and controllable phase-shift range while maintaining stable microwave performance requires careful design of the programmable dielectric structure and its interaction with guided EM waves.

Challenge 3: Integration of FPMS as an HIS

High Impedance Surfaces (HIS) have been extensively employed by antenna community to enhance the impedance and more commonly the radiation performance of a radiator. A typical example of HIS is an Artificial Magnetic Conductor (AMC). AMCs respond by providing an in-phase reflection to the incident wave and suppress any surface-wave propagation thus resulting in improved antenna gain. This topic is of keen interest for designs pertaining to FPMS. The rationale is the reliance of the FPMS technology on a unit cell that has a distinct correlation with AMC operation. Therefore, it is natural to investigate FPMS as an AMC layer and study its effect on antenna characteristics. This kind of a proposal is laced with a series of challenges among which the integration of FPMS as an AMC with the antenna can be regarded as the primary. This needs to be properly chalked out when using a functional AMC layer with an antenna radiator.

Challenge 4: Beam Steering Using FPMS

Electronic beam steering is conventionally realized using phased-array antennas combined with phase-shifting networks. Although these systems provide excellent beam control, they require numerous RF components, increasing hardware complexity, physical size, power consumption, and system cost. Developing compact beam-steering solutions using programmable EM structures therefore remains a significant challenge. Whether beam manipulation is achieved through dielectric phase shifters or FPMS-based AMCs, the interaction between the programmable structure and the radiating element must be carefully engineered to achieve efficient beam steering while maintaining satisfactory impedance matching, gain, and radiation characteristics.

Challenge 5: Development of an Analytical Model

The design of FPMS-based microwave structures has largely relied on full-wave EM simulations and experimental optimization of individual RF components. The analytical models presented until now deal with the unit cell itself but do not extend to the complete RF component itself. This limits the physical insight into the underlying operating mechanisms. Developing analytical models that accurately describe wave propagation, phase variation, and beam-steering mechanisms is therefore essential not only for the unit cell itself but for the complete RF component under discussion. This would require the theoretical study of the unit cell in combination with the RF component to get the complete picture. Such theoretical models not only improve the physical understanding of FPMS-based devices but also provide efficient design tools for future reconfigurable microwave and antenna systems.

Challenge 6: Practical Implementation and Experimental Validation

The practical realization of FPMS-based microwave components presents additional fabrication and measurement challenges. The simultaneous integration of RF circuitry, DC biasing networks, and varactor diodes increases the complexity of both the hardware implementation and the experimental setup. Furthermore, validating the proposed concepts requires comprehensive characterization of reflection characteristics, antenna gain and beam-steering performance under multiple biasing conditions, demanding carefully designed prototypes and reliable measurement procedures.

1.3 Objectives

Based on the challenges discussed in the previous section, the primary objective of this research is to investigate the feasibility of FPMS as a versatile platform for realizing compact reconfigurable microwave components capable of achieving dynamic phase control and beam steering. To accomplish this objective, the following specific research objectives are established:

- 1) To investigate the EM behavior of FPMS under different excitation environments, including guided-wave propagation and periodic boundary conditions, in order to establish a comprehensive understanding of its programmable dielectric characteristics and suitability for microwave and antenna applications.
- 2) Develop analytical models capable of explaining the operating principles of the proposed FPMS-based microwave structures, including phase variation in dielectric phase shifters and beam-steering mechanisms in AMC-integrated antennas, thereby providing physical insight and an efficient design methodology.

- 3) Develop an FPMS-based AMC by exploiting the programmable EM characteristics of the FPMS unit cell and investigate its reflection properties under periodic excitation. Doing so, study the effect of the FPMS reconfigurability on the AMC performance.
- 4) Design a compact gain-enhanced beam-steerable antenna integrated with the proposed FPMS-based AMC, and evaluate the influence of FPMS on antenna gain, radiation characteristics, and beam steering.
- 5) Design and analyze a reconfigurable dielectric phase shifter based on FPMS, capable of providing controllable phase variation through dynamic modification of the effective dielectric properties of the propagation medium. Also, show the integration of this phase shifter in a phased array environment to show its operation as a smart reconfigurable RF component.
- 6) To fabricate and experimentally validate the proposed microwave components and antenna structures, validating the proposed theoretical models and simulation setups providing evidence of the FPMS technology as a candidate for the implementation of intelligent RF/microwave components and systems.

The remainder of this thesis is organized into four chapters. Chapter 2 presents the theoretical background and literature review, and is divided into three major sections. Section 2.1 reviews the fundamentals and existing research on Field Programmable Microwave Substrates (FPMS), Section 2.2 discusses the operating principles and applications of Artificial Magnetic Conductors (AMCs), and Section 2.3 reviews conventional and reconfigurable phase shifters, highlighting their operating principles, advantages, limitations, and existing research gaps. Chapter 3 presents the design, analysis, fabrication, and experimental validation of the proposed FPMS-based AMC antenna. It comprises Section 3.1 (Introduction), Section 3.2 (Operation of FPMS-Based AMC),

Section 3.3 (Analytical Modeling of the FPMS-Based AMC), Section 3.4 (Simulation of the FPMS-Based AMC Antenna), and Section 3.5 (Fabrication and Measurements). Chapter 4 presents the development of the proposed FPMS-based reconfigurable phase shifter. It consists of Section 4.1 (Introduction), Section 4.2 (Operation of the Reconfigurable Phase Shifter), Section 4.3 (Analytical Modeling of Wave Propagation), Section 4.4 (Phase Shifter Simulations), and Section 4.5 (Fabrication and Measurements). Finally, Chapter 5 summarizes the major contributions of this research, discusses its limitations, and provides recommendations and future research directions for the continued development of FPMS-based reconfigurable microwave and antenna systems

1.4 Thesis Contributions

This thesis investigates FPMS as a versatile reconfigurable EM platform capable of controlling wave propagation through electronically tunable effective material properties. By varying the capacitance of embedded varactor diodes, the electromagnetic response of FPMS can be dynamically modified, enabling adaptive microwave components without mechanical tuning or complex conventional reconfigurable architectures. Two major applications are developed to demonstrate this capability. First, FPMS is employed as a programmable AMC to achieve electronic beam steering of a single radiating element through control of the reflection phase. Second, FPMS is utilized as a programmable dielectric medium to realize a continuously tunable phase shifter through control of guided-wave propagation. Analytical models are developed for both configurations and validated using full-wave EM simulations and experimental measurements. The results demonstrate FPMS as a multifunctional platform for manipulating both reflected and guided EM waves and provide experimentally validated design methodologies for future compact and electronically reconfigurable RF and microwave systems.

Chapter 2

Background and Literature Review

Modern RF and microwave systems are increasingly required to operate under dynamically changing operating conditions imposed by adaptive wireless networks, radar, satellite communications, and next-generation 6G systems. Conventional microwave components are generally designed for fixed operating conditions, where their frequency response, impedance, phase characteristics, polarization, and radiation behavior are determined by their physical geometry. However, many emerging applications require these characteristics to be modified after fabrication without redesigning the hardware. Consequently, significant research has been devoted to reconfigurable RF and microwave components that can electronically alter their EM response through external control signals.

Reconfigurability can be incorporated into various microwave components, including antennas, filters, phase shifters, matching networks, couplers, and amplifiers. Depending on the application, reconfiguration may involve frequency tuning [12], [13], impedance matching [14], polarization switching [15], [16] or radiation-pattern control [17]. For antenna systems, frequency reconfiguration enables multi-band operation, polarization reconfiguration improves communication reliability, while radiation-pattern reconfiguration allows the main beam to be electronically redirected. Similarly, in phased arrays and beamforming networks, phase shifters provide controlled phase differences between antenna elements, enabling electronic beam steering without mechanical movement [18], [19].

Several techniques have been proposed to realize microwave reconfigurability. Semiconductor devices remain the most widely adopted approach due to their ease of integration with planar

circuits. PIN diodes are commonly employed to realize discrete operating states by electrically connecting or isolating sections of a microwave circuit, thereby modifying the resonant length, current distribution, or phase response [20]. In contrast, varactor diodes provide continuous tuning by exploiting their voltage-dependent capacitance, allowing the resonant frequency, impedance, or phase response to be adjusted through reverse-bias control [21]. Although these devices are compact and inexpensive, they often introduce additional insertion loss, nonlinear behavior, limited power handling capability, and biasing complexity.

RF microelectromechanical systems (RF MEMS) provide an alternative reconfiguration mechanism with significantly lower insertion loss, higher isolation, and superior linearity than semiconductor switches, making them attractive for high-performance switches, filters, phase shifters, and reconfigurable antenna arrays [22], [23]. Nevertheless, RF MEMS devices generally require high actuation voltages, specialized fabrication processes, and careful packaging, while switching speed and long-term reliability remain practical challenges.

Besides semiconductor devices, several tunable materials have been investigated for microwave reconfiguration. Ferrite-based components exploit magnetic biasing to modify permeability and have been extensively used in non-reciprocal microwave devices and phase shifters [24]. Ferroelectric materials, particularly barium strontium titanate (BST), offer electrically tunable permittivity and have been successfully applied to tunable filters, phase shifters, and impedance matching circuits [25], [26]. Similarly, liquid crystals provide electrically controlled dielectric properties at microwave and millimeter-wave frequencies [27], whereas graphene has recently emerged as a promising material because of its electrically tunable surface conductivity [28]. Although these materials provide continuous EM tuning, their practical implementation is often

limited by fabrication complexity, dielectric losses, tuning speed, or integration with conventional microwave technologies.

Another important class of reconfigurable microwave structures is based on metamaterials and metasurfaces. These structures consist of periodically arranged unit cells whose effective EM properties can be engineered to control wave propagation. By incorporating active devices or tunable materials into the unit cells, parameters such as surface impedance, reflection phase, transmission phase, polarization, and radiation characteristics can be dynamically controlled [29], [30]. Consequently, metamaterial-based structures have found widespread applications in reflectarrays, transmitarrays, artificial magnetic conductors (AMCs), frequency-selective surfaces, and beam-steering antennas. Their principal advantage lies in providing EM control directly at the unit-cell level, offering considerably greater flexibility than conventional circuit-based tuning approaches. However, their performance strongly depends on unit-cell optimization, bias network design, loss mechanisms, and the EM interaction between the surface and the radiating element.

Overall, the reviewed techniques demonstrate that microwave reconfigurability can be achieved using semiconductor switches, reactive loading, tunable materials, and engineered EM surfaces. Each approach provides distinct advantages but also introduces compromises in insertion loss, tuning range, bandwidth, power handling, fabrication complexity, biasing requirements, and system integration. Therefore, there remains a continued demand for alternative reconfigurable platforms capable of providing efficient EM control while maintaining compatibility with planar microwave technologies.

Recently, a new paradigm has emerged in this direction, known as the FPMS [31]. Unlike conventional substrates that possess fixed dielectric properties, FPMS incorporates electronically controlled unit cells within the substrate, allowing the effective EM properties experienced by the

propagating wave to be modified through electrical biasing. This enables dynamic control of the equivalent substrate characteristics without physically altering the structure. Since FPMS forms the basis of the proposed work in this thesis, the following sections review its operating principle, reported implementations in reconfigurable microwave components, and its application to AMC-based antennas and phase shifters.

2.1 Field Programmable Microwave Substrate (FPMS)

2.1.1 Concept and Operating Principle

After reviewing the broader concept of reconfigurability in RF and microwave systems, this section focuses on a relatively recent programmable EM platform known as the Field Programmable Microwave Substrate (FPMS). The fundamental building block of an FPMS is a reconfigurable unit cell consisting of two orthogonally oriented microstrip lines arranged in a cross-shaped configuration, as shown in Fig. 2.1.

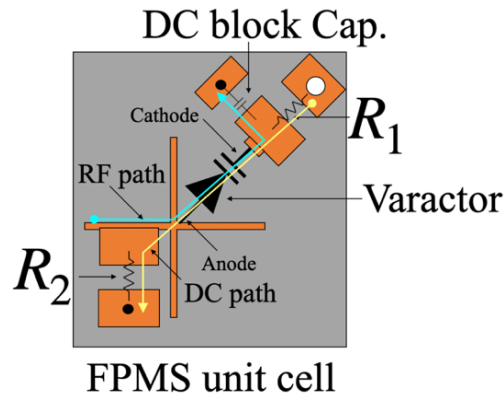


Fig. 2.1. FPMS unit cell (R_1 and R_2 are resistors)

A varactor diode is positioned at the intersection of the metallic strips and acts as a voltage-controlled capacitor. By varying the reverse bias voltage across the varactor diode, its capacitance can be dynamically adjusted, thereby modifying the EM response of the structure and providing a

control to the designer on the unit cell's impedance. The biasing network typically consists of a RF choke resistor and a DC blocking capacitor, allowing independent control of the varactor while minimizing interference between the radio-frequency and biasing circuits. The operation of FPMS is controlled through the voltage-dependent response of the varactor-loaded unit cell. When the reverse-bias voltage is varied, the capacitance of the varactor placed at the intersection of the metallic strips changes. This variation modifies the impedance presented by the unit cell to the incident EM wave. Consequently, the reflected and transmitted fields through the unit cell are altered, and the complete structure exhibits a bias-dependent equivalent EM response. In this sense, the FPMS unit cell behaves like a tunable artificial medium. The extracted effective permittivity therefore represents the combined response of the host substrate, metallic pattern, varactor diode, and biasing network, rather than a physical change in the dielectric constant of the substrate itself.

The bias-dependent response of the FPMS unit cell can be evaluated from its scattering parameters using the Nicholson–Ross–Weir (NRW) method [31]. In this method, the unit cell is modeled as an equivalent homogeneous slab with thickness (d). The simulated or measured reflection coefficient (S_{11}) and transmission coefficient (S_{21}) are then used to retrieve the effective impedance, refractive index, permittivity, and permeability of the equivalent medium. These retrieved parameters are not the physical properties of the dielectric substrate alone; rather, they represent the combined EM response of the complete FPMS unit cell.

The normalized effective impedance is obtained as:

$$Z_{\text{eff}} = \pm \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (2.1)$$

The intermediate parameter (X), which relates the reflection and transmission coefficients to the wave propagation inside the equivalent slab, is calculated as:

$$X = \frac{1 - S_{11}^2 + S_{21}^2}{2S_{21}} \quad (2.2)$$

The propagation factor inside the equivalent slab is then expressed as:

$$e^{-jn_{eff}k_0d} = X \pm \sqrt{(X^2 - 1)} \quad (2.3)$$

Using this propagation factor, the effective refractive index is retrieved as:

$$n_{eff} = \frac{-j}{k_0d} \ln(e^{-jn_{eff}k_0d}) \quad (2.4)$$

where k_0 is the free-space wavenumber. Once z_{eff} and n_{eff} are determined, the effective permittivity and effective permeability are calculated as:

$$\epsilon_{eff} = \frac{n_{eff}}{Z_{eff}} \quad (2.5)$$

$$\mu_{eff} = n_{eff}Z_{eff} \quad (2.6)$$

For the FPMS structures considered in this thesis, the extracted effective permittivity, ϵ_{eff} is the main parameter of interest because it indicates whether the unit cell supports or suppresses EM wave propagation. Depending on the operating frequency and applied bias condition, the unit cell can exhibit both positive and negative effective permittivity regions which is shown in Fig. 2.2.

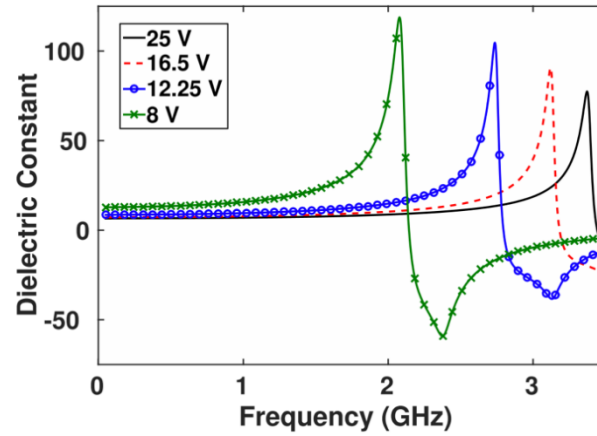


Fig. 2.2 Effective permittivity of Rogers 5870 board [31]

At frequencies below the resonance of the unit cell, the effective permittivity remains positive, indicating that the structure behaves similarly to a conventional dielectric medium and supports normal EM wave propagation. However, as the operating frequency approaches and exceeds the resonant frequency, the effective permittivity transitions from positive to negative values. In this negative permittivity region, EM wave propagation is significantly suppressed, resulting in an evanescent response where the fields rapidly decay within the structure. This ability to switch between positive and negative permittivity regions forms the basis for implementing a wide range of reconfigurable microwave components using FPMS technology.

It should be emphasized that the dielectric response shown in Fig. 2.2 corresponds to a specific FPMS unit cell geometry. The resonant frequency, tuning range, and the resulting positive and negative permittivity regions are strongly dependent on the physical dimensions of the unit cell, substrate characteristics, and the range of capacitance values employed. Consequently, the EM behavior of the structure can be tailored for different applications through appropriate optimization of these design parameters. For example, a unit cell optimized for operation in a reconfigurable phase shifter at 2.5 GHz would require a significantly different design than a unit cell intended for an AMC operating in the X-band. Therefore, the dielectric response shown in Fig. 2.2 provides an

understanding of the underlying operating principle of FPMS, the actual unit cell dimensions employed in this thesis are optimized according to the requirements of the respective applications. The design procedures and optimization studies associated with the proposed phase shifter and antenna structures are presented in the subsequent chapters.

2.1.2 Use of FPMS in Reconfigurable Microwave Components Design

The first practical demonstration of FPMS technology was reported in [31], where the concept has been introduced for the first time. Using the FPMS unit cells, the authors show the realization of reconfigurable waveguide structures whose performance can be controlled by the dynamic nature of the concept, Fig. 2.3(a) and (b). Multiple FPMS unit cells are arranged in defined patterns to form regions exhibiting positive and negative effective permittivity. EM waves are allowed to propagate through the positive permittivity (highlighted as blue) region, while the surrounding negative permittivity cells (highlighted as white) act as the sidewalls that confined the propagating fields within the positive permittivity region. Consequently, the guiding mechanism is achieved without relying on conventional metallic boundaries and its behavior is similar to parallel plate waveguide response. Furthermore, by electronically reconfiguring the state of the unit cells, the effective width of the waveguide could be modified, resulting in a controllable cutoff frequency as shown in Fig. 2.3 (b) and (c). Interestingly, an upper bound can be seen to the waveguide response which is different from conventional waveguide structures. This upper bound is due to the resonance of the unit cell structures exhibiting positive dielectric constant (blue unit cells). The unit cell structure in combination with the varactor provides a self-resonance response, where the overall impedance of the unit cell changes from capacitive to inductive. This causes the extract permittivity to change from positive to negative values thus blocking any further propagation of

wave. Thus, the waveguide realized using FPMS concepts acts more as a band-pass filter rather than the traditional high-pass filter response.

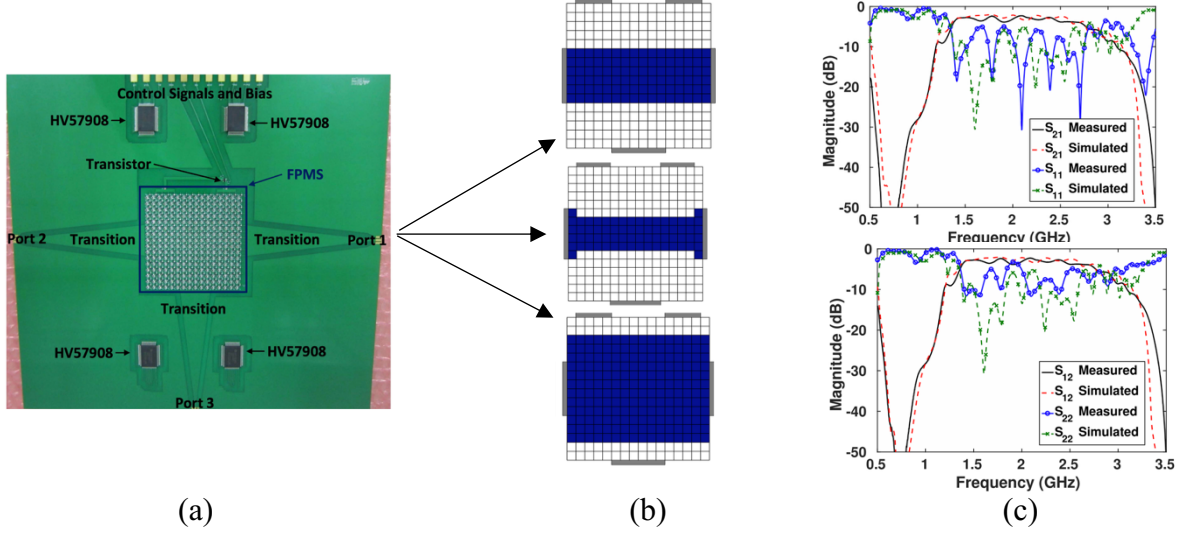


Fig. 2.3. FPMS-based reconfigurable waveguide structure adapted from [20] (a) FPMS board (b) Waveguide implementation using FPMS for various frequency (c) Comparison of simulated and measured waveguide performance [31]

Experimental results demonstrated that the operating characteristics of the waveguide could be dynamically altered through bias control, thereby validating the feasibility of using FPMS as a programmable EM medium. The authors in [32] showed the control on the center frequency of the filter, its bandwidth and even its topology. Thus, the designer can switch a filter from being band-pass to band-stop and vice versa. This type of control on RF components is quite rare and mostly unheard of. Extending the potential of this technology, the authors in [33] used the FPMS concept not only on passive components (such as filters) but also to active microwave circuits, like amplifier [33]. The programmable nature of the substrate is utilized to realize reconfigurable input and output feeding structures for a transistor-based amplifier. By modifying the configuration of the FPMS unit cells surrounding the transistor, the EM characteristics of the feeding network could be altered without physically changing the circuit topology. As a result, which is shown in Fig.

2.4(b), the operating frequency and bandwidth of the amplifier were dynamically controlled using the same hardware platform.

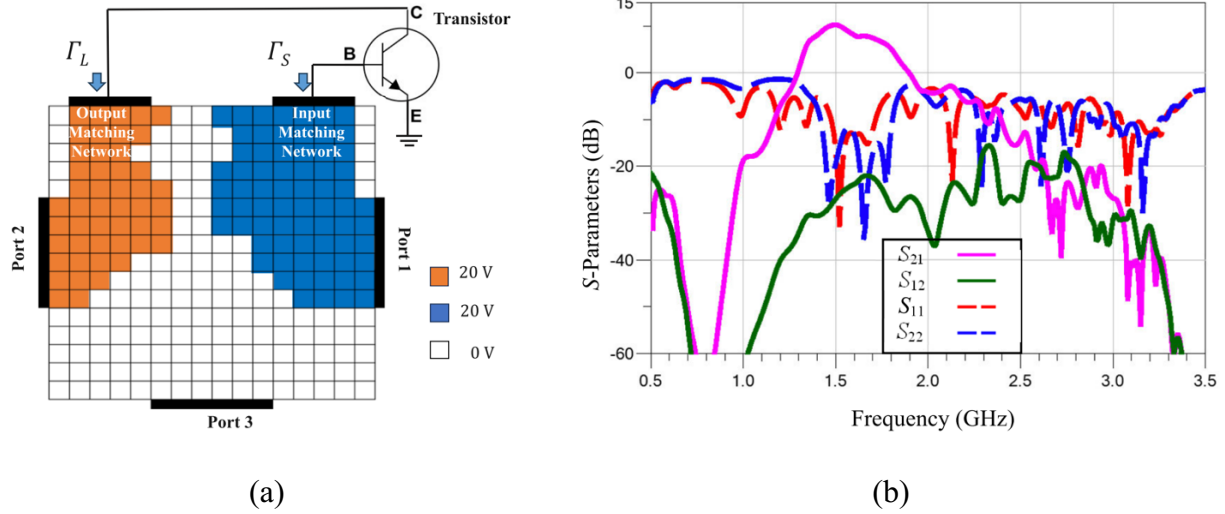


Fig. 2.4 (a) Block diagram of reconfigurable amplifier (b) Simulated results for amplifier centered at 1.5 GHz [33]

In another demonstration, the performance metric of the FPMS concept is further extended [20]. By exploiting the programmable nature of FPMS, the authors in [32] demonstrated that the EM coupling between resonant sections could be dynamically controlled through the biasing of individual unit cells. This capability enabled the realization of tunable bandpass and bandstop filters whose operating characteristics could be modified without altering the physical structure of the circuit. The response of bandpass filter is shown in Fig. 2.5. In particular, the center frequency, bandwidth, and filter response are electronically reconfigured using the same hardware platform. The reported results highlighted the flexibility of FPMS in implementing multifunctional microwave circuits and demonstrated its potential as a reconfigurable filtering technology. Also, authors used same FPMS board for an impedance tuner to cover broad coverage and also worked on the optimization of FPMS unit cell [34] [35]. Like any other technology, the agility of FPMS designs came with its own set of challenges. The most prominent one being the unusually high

insertion loss observed in the filter designs. This kind of response could be a major bottle-neck for a new technology and for it to be made mainstream. Thus, this downside among other left a door wide open for the future works to improve on this performance parameter. After the demonstration of non-radiating RF components that can be tuned using the concept of FPMS, the natural inclination is to check the veracity of it in antenna design.

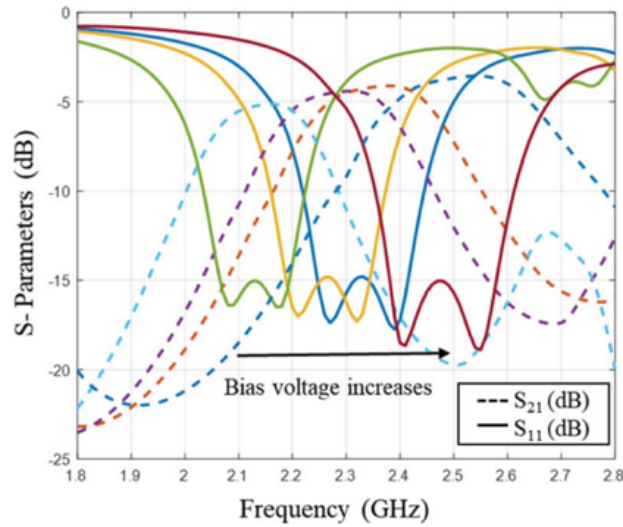


Fig. 2.5 Simulation of tunable bandpass filter [32]

However, at the same time it is imperative that the loss incurred by FPMS unit cells should be lowered. Therefore, in [36], the substrate choice is modified from the original FR-4 to Rogers 5880. The latter being significantly low loss ($\tan\delta = 0.0009$) and at the same time with lower relative permittivity ($\epsilon_r = 2.2$), which is usually conducive to antenna design. A first-ever Half-Mode Substrate Integrated Waveguide (HMSIW) leaky-wave antenna (LWA) is used as the proof-of-concept to achieve beam steering relying on electronic control provided by FPMS. At the same time, the authors chose to increase the frequency band to X-band rather than S-band that is initially shown for filter designs. The antenna design, which is shown in Fig. 2.6, replaces the conventional

metallic sidewalls are replaced by FPMS unit cells, allowing the propagation characteristics of the guided wave to be altered through varactor biasing.

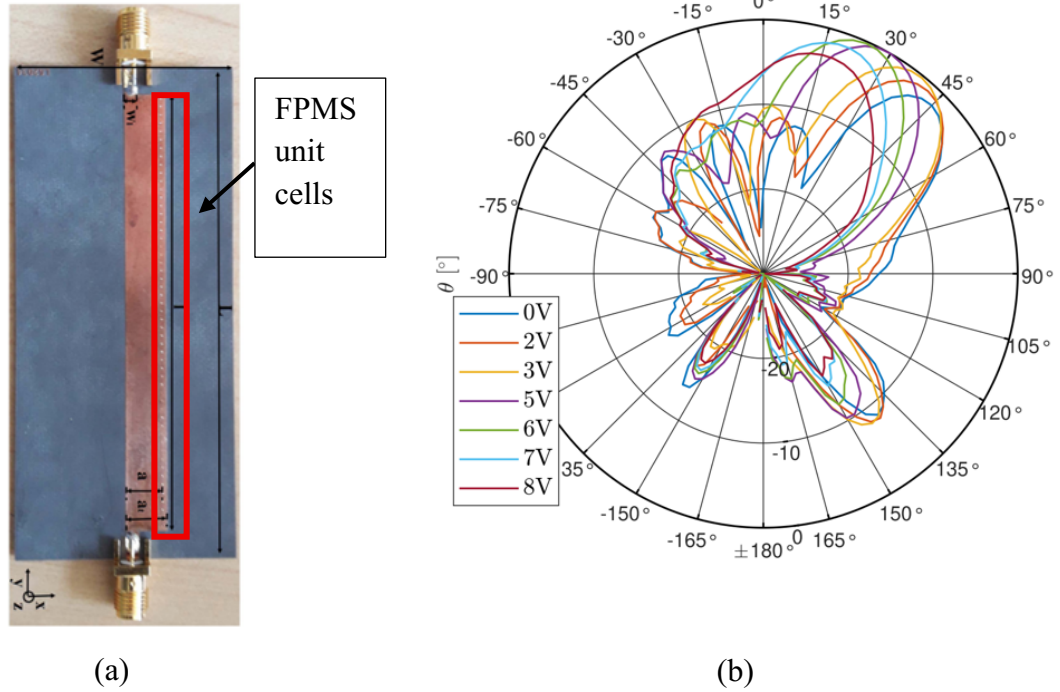


Fig. 2.6 SIW based antenna (b) Radiation pattern at 8 GHz [36]

As a result, the phase constant of the antenna is dynamically altered, leading to electronic beam steering without modifying the operating frequency, which is conventionally done in LWA designs. The reported results demonstrated beam scanning of approximately 27° at 8 GHz, thereby providing the first experimental evidence that FPMS could be employed for radiation pattern reconfigurability in antenna systems. The gain of the antenna is measured to be 7.75 dBi with a maximum variation of 1.5 dB as the beam is steering, Fig. 2.6 (b). These results validate the use of FPMS in radiating structures and at the same time provides adequate results of antenna efficiency thereby answering the concerns of higher loss in the previously reported filter designs.

Building upon this concept, we further extended the application of FPMS in leaky-wave antennas by introducing a beam-steerable Half-Mode Substrate Integrated Waveguide antenna

operating at X-band frequencies [37]. Although the antenna presented in [34] successfully demonstrated the feasibility of using FPMS for beam steering, several limitations remained. The beam steering range was limited, the steering occurred only in the forward direction, and the FPMS unit cells were directly integrated into the antenna without first establishing their EM characteristics through a detailed unit-cell analysis. Furthermore, the relationship between the programmable substrate properties and the propagation characteristics of the LWA was not fully investigated. These limitations motivated further research to better understand the behavior of FPMS at higher frequencies and to improve the beam-steering capability of FPMS-based antennas. To address these challenges, authors in [37], proposed an improved FPMS-integrated HMSIW LWA operating at 12 GHz in the X-band, as shown in Fig. 2.7

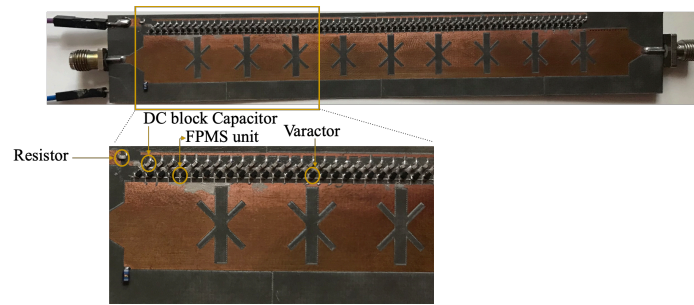


Fig. 2.7. Top view of LWA [37]

Similar to the previous design, the antenna employs FPMS unit cells along one longitudinal edge of the HMSIW to replace the conventional metallic sidewall. Simulation results demonstrated continuous beam steering from -35° to $+37^\circ$, corresponding to a total steering range of approximately 72° , while maintaining a peak realized gain of approximately 6.4 dBi and a reflection coefficient below -10 dB around 12 GHz. The results are shown in Fig. 2.8. Experimental validation using fixed capacitors in place of the varactor diodes achieved a measured gain of 6.2 dBi and a beam steering range of approximately 35° . Although the measured steering was smaller than the simulated result due to the self-resonant frequency limitation of the

commercial varactor diodes, the measurements successfully validated the proposed beam-steering mechanism and demonstrated the practicality of FPMS for reconfigurable antenna applications.

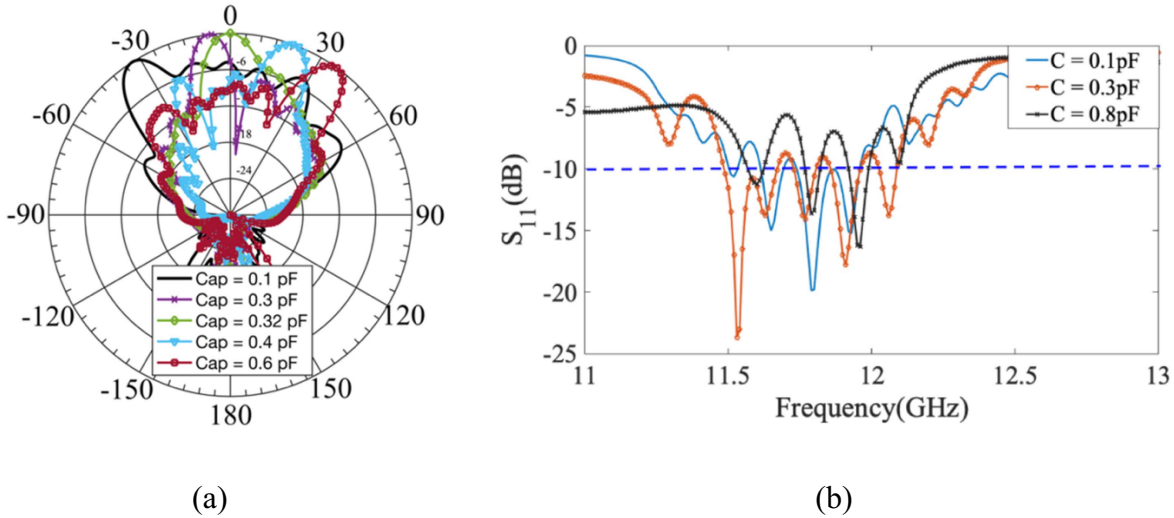


Fig.2.8 (a) Beam steering at 12 GHz (b) Matching of antenna at 12 GHz [37]

2.1.3 Research Gaps in Existing FPMS Literature

The designs discussed in the last section demonstrate that FPMS has evolved from a novel EM concept into a versatile platform for reconfigurable microwave systems. Previous studies have successfully employed FPMS in waveguides, amplifiers, filters, and antenna structures, thereby validating its ability to dynamically control the effective dielectric properties experienced by EM waves [31]-[37]. These investigations have collectively established the feasibility of using programmable substrate characteristics to manipulate wave propagation, resonance behavior, and radiation performance within a single hardware platform to control individual RF components for their behavioral characteristics.

Despite these advancements, several important research gaps naturally remain in this technology because of it being fairly recent. The reported designs based on FPMS generally rely on guided-wave microwave structures and thus used lumped-port excitation for the characterization of individual unit cells. This limits the use of the unit cells in RF components which are conducive

to such excitation. Therefore, it would be lucrative to explore other possible scenarios for studying these unit cells in RF environments that could lead to a variety of other implementations. Thus, one of the focuses of this thesis would be to explore ideas that are different from existing literature. One such example is to study the EM behavior of FPMS unit cells under periodic boundary conditions. The consequent study would be quite useful in many of the popular RF designs such as metasurfaces. The dynamic nature of FPMS would provide an intelligent control over these metasurface structures that could then be used to drive the propagation of EM wave as per designer's need. One of the most common uses of metasurface is as an Artificial Magnetic Conductor (AMC). AMC is a High Impedance Surface (HIS) that is generally used with the antennas to increase the gain and directivity performance. The integration of FPMS unit cells to such elements would allow for an added control. At the same time, it would require a fresh study on FPMS when used as an AMC layer. Thus, this would be one of the focuses of the thesis.

The other absentee in the current FPMS literature is the design of phase shifters to be used in integrated phased array antenna designs. FPMS is envisioned as an all-encompassing technology that can be answer to a variety of RF challenges. Thus, its use in a fundamental RF circuit such as phase shifter needs to be studied to complete the loop. The ability of FPMS to dynamically control the effective permittivity of a propagation medium has a close synergy with phase shifting components. Therefore, it is logical to explore such an RF component that relies on FPMS technology. It is quite obvious that the main purpose of a tunable phase shifter is to be employed in beam steering applications. This area of RF engineering is completely untouched in the FPMS domain.

On the basis of the conclusions drawn above the literature review in this chapter focuses on two areas:

- 1) AMC based antenna designs
- 2) Existing electronically controlled phase shifters.

The consequent section thus forms the foundation of the research presented in this thesis.

2.2 Artificial Magnetic Conductors (AMC)

AMCs also known as HISs, are engineered periodic structures that approximate the behavior of an ideal Perfect Magnetic Conductor (PMC) over a finite frequency range. A typical AMC structure is shown in Fig. 2.10, when it was initially introduced in the antenna engineering domain. Since a PMC does not physically exist, AMC is realized as a substitute to it. This is accomplished artificially by arranging periodic metallic elements on a grounded dielectric substrate. The shape of these metallic elements can be varied as per the designer's need and goals. The concept has been first introduced by Sievenpiper ~30 years ago and has since received significant attention in antenna engineering due to its distinctive reflection phase response and its ability to improve antenna's radiation performance [38].

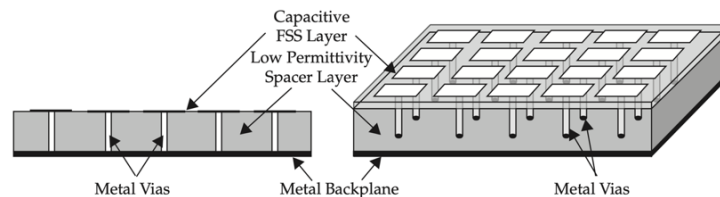


Fig. 2.10. Schematic design of AMC surface [38]

The key feature of an AMC is its reflection phase response. For a conventional PEC, the reflected electric field undergoes a 180° phase reversal. In contrast, an AMC provides an approximately 0° reflection phase at its resonant frequency, allowing the incident and reflected fields to remain in phase above the surface. This in-phase reflection condition produces

constructive interference between the incident and reflected waves and makes AMCs useful for antenna designs requiring gain enhancement, improved radiation efficiency, and reduced antenna profile [39]. The typical reflection phase response of an AMC is shown in Fig. 2.11.

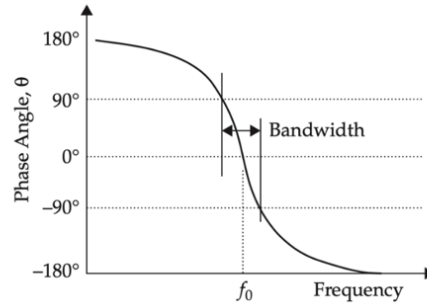


Fig. 2.11 Notional Reflection Phase of AMC [38]

The reflection phase of an AMC is inherently dispersive, as shown in Fig. 2.11. At frequencies below the resonant frequency, the reflection phase is positive and gradually decreases as the operating frequency approaches resonance. At the resonant frequency, the reflection phase crosses 0° , where the AMC exhibits PMC-like behavior and the reflected electric field remains in phase with the incident field. As the operating frequency increases beyond resonance, the reflection phase becomes negative and approaches -180° . The useful operating bandwidth of an AMC is commonly defined as the frequency range over which the reflection phase lies between $+90^\circ$ and -90° . Within this region, the reflected wave remains sufficiently in phase with the incident wave to produce constructive interference, making it suitable for low-profile antenna applications and gain enhancement [38].

The reflective nature of AMC structures makes them a viable candidate in antenna design for gain enhancement. The two primary highlights of the AMC are i) the ability to present high impedance to a localized source or plane wave looking down into the structure, ii) providing a surface-wave band-gap that prohibits the transverse propagation of energy along the surface. For

plane waves at normal incidence, an AMC may be understood as an electrically short length of shorted transmission line with a shunt capacitance placed at the reference plane of the outer surface, Fig. 2.15 [38].

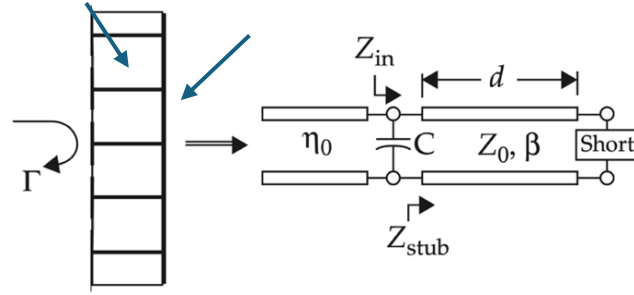


Fig. 2.15. Cross-sectional view of AMC and its equivalent circuit model [38]

The electrically short length of shorted transmission line can be approximated by an inductor. Thus, the equivalent circuit can be effectively described with a parallel LC circuit, whose input impedance is shown in (Eq. (2.7)). At resonance, the short circuit responses transformed to a perfect open circuit response: the so-called perfect magnetic conductor (PMC) boundary with 0° reflection phase and unity magnitude. In the next section, the chapter delves into some of the examples of AMC based antenna designs.

$$\frac{1}{Z_{in}} \approx \frac{j\omega L}{1 - \omega^2 LC} \quad (2.7)$$

2.2.1 Existing Literature on AMC and Metasurface Based Antennas

Early studies on AMC-based antennas mainly employed the AMC surface for gain enhancement [40], [41], [42]. For example, an AMC surface is employed to improve the radiation performance of printed dipole antennas for wideband vehicular wireless communications [40]. The proposed antenna consists of a low-profile printed dipole antenna backed by a 3x3 rhomboid AMC reflector, as shown in Fig. 2.16(a). The AMC surface is introduced to enhance the forward radiation, improve the radiation efficiency, and broaden the impedance bandwidth while maintaining a compact

antenna profile. The results in Fig. 2.16(b) show that the AMC-backed antenna achieves ~ 4 dBi gain enhancement at required frequency range.

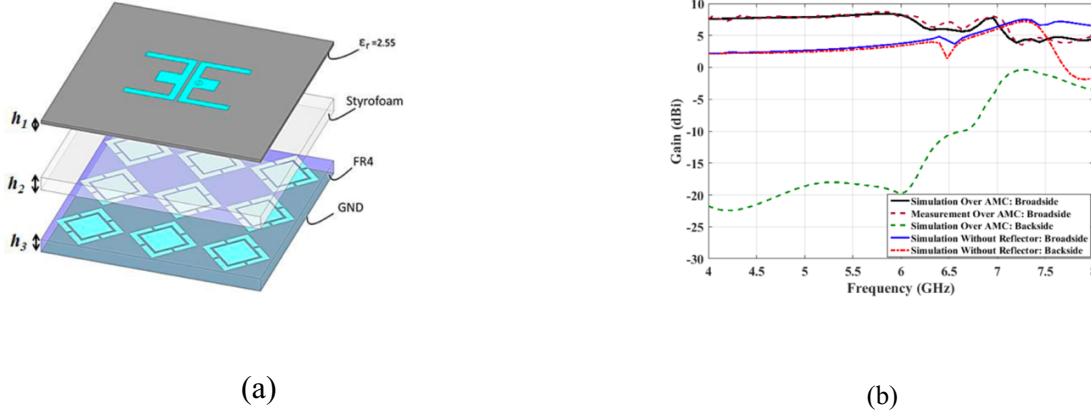


Fig. 2.16 (a) 3D view of antenna on top of AMC surface (b) Simulated and measured gains of design with and without AMC reflector [40]

AMC surfaces have also been used in more application-specific antenna platforms. In [42], authors employed AMC units to improve the performance of an antenna-on-chip operating at 94 GHz. Silicon is not among the best candidates for RF and millimeter-wave frequencies because of its highly lossy nature. To address this issue, the authors introduced coupling enhancement structures to improve the energy transfer between AMC units placed in the near-field region of the antenna, as shown in Fig. 2.17.

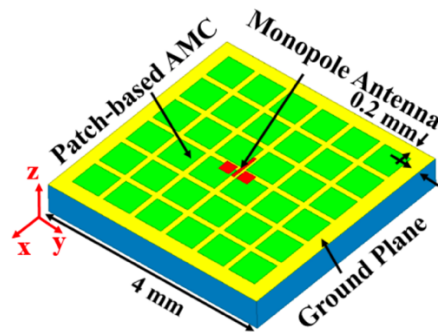


Fig. 2.17. Planar monopole antenna on chip with 6 x 6 patch-based AMC [42]

The proposed design achieved a maximum gain of 4.3 dBi, which is a significant result for a silicon-based on-chip antenna where sometimes the gain values reported are less than 0 dBi (isotropic gain). These studies show that AMC surfaces can be effectively used to improve antenna gain and radiation efficiency across different antenna platforms.

The studies discussed above show that AMC surfaces have been effectively used for gain enhancement, bandwidth improvement, and radiation performance optimization. However, most of these designs are passive and do not provide radiation pattern reconfigurability. Their performance is sublime in terms of radiation characteristics but dynamically limited. A quick glance at the literature shows that not many designs have focused on a tunable AMC design. Some examples exist that combine reconfigurable AMC surfaces with antenna radiators to achieve additional control on antenna's beam. For example, in [43], a cross-slot radiator backed by a (19 x 19) AMC reflector and a 9x9 transmissive coding metasurface (TCMS), as shown in Fig. 2.18(a) and (b), respectively.

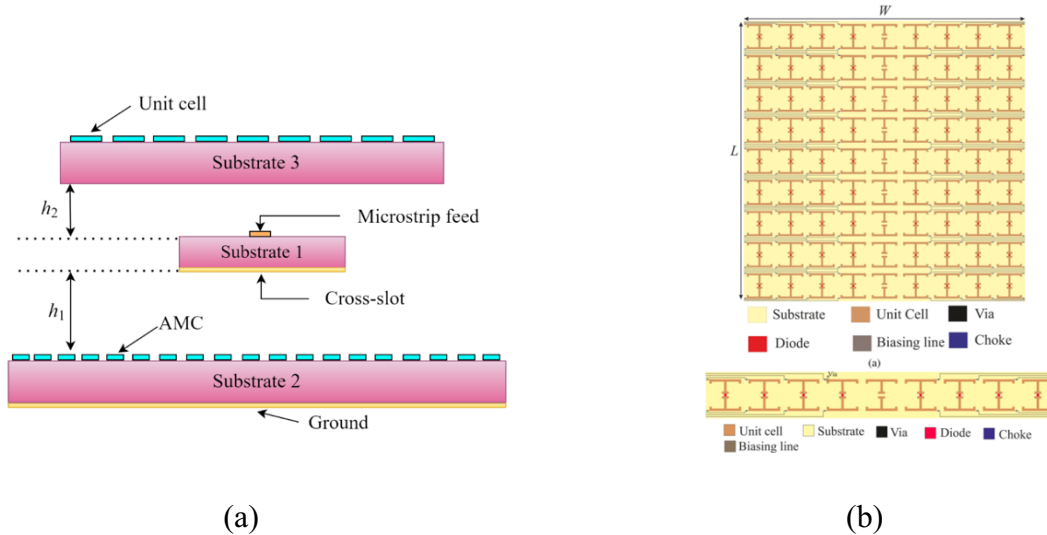


Fig. 2.18 (a) Side view of conceptual design (b) Top view of metasurface [43]

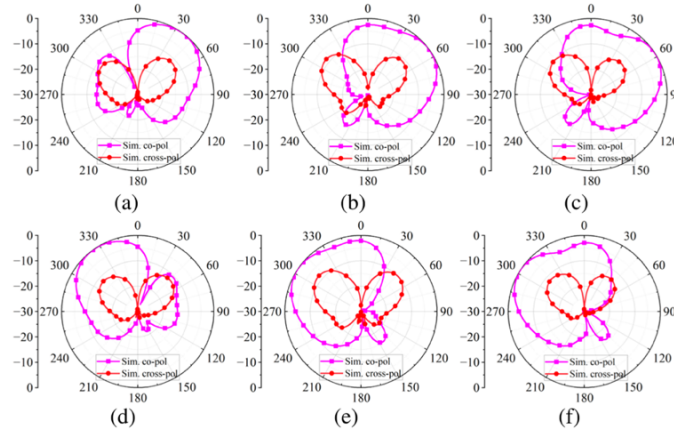


Fig. 2.19 Normalized radiation pattern at an operating frequency of 3.5 GHz in the yz-plane (a) State 7, (b) State 8, (c) State 9, (d) State 10, (e) State 11, and (f) State 12 (States represent different coding of metasurface such as turning ON/OFF PIN diodes) [43]

The cross-slot antenna initially produces a bidirectional radiation pattern with a maximum gain of 3.35 dBi. By incorporating the AMC beneath the radiator, the backward radiation is significantly suppressed, transforming the radiation into a unidirectional pattern while increasing the peak gain to 7.11 dBi. The TCMS is subsequently introduced above the antenna to realize beam steering and beamwidth reconfiguration through PIN-diode-controlled phase manipulation. As illustrated in Fig. 2.19, the complete antenna achieves a beam steering range of $\pm 57^\circ$ in one plane with a peak gain of 8.47 dBi.

Further examples of pattern-reconfigurable AMC-backed antennas are reported in [44] and [45]. In [44], a beam-steerable slot antenna was proposed using a cross-shaped main radiating slot, two parasitic slots, and an AMC reflector (Fig. 2.20). Each parasitic slot is loaded with a PIN diode, which controls the radiation state of the antenna. When both PIN diodes are OFF, the antenna radiates toward the broadside direction. By switching one diode ON and the other OFF, the antenna beam is tilted to either side, resulting in three discrete radiation states. The radiation patterns of antenna are shown in Fig. 2.21.

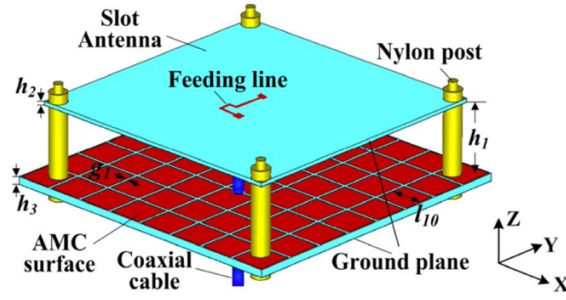


Fig. 2.20. Perspective view of antenna prototype with AMC [44]

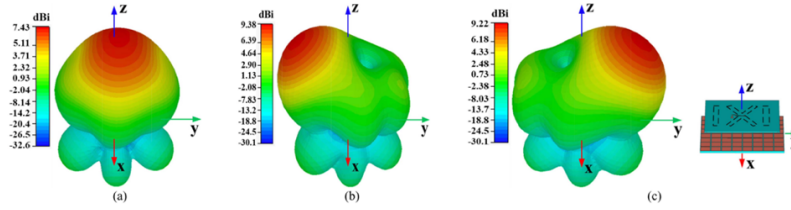


Fig. 2.21. (a) state I, (b) state II, (c) state III [44]

The AMC surface, placed above the ground plane, improves the directional radiation and reduces the back lobe. However, the beam steering is mainly achieved by switching the parasitic slots rather than by reconfiguring the AMC itself.

A similar concept is presented in [45], authors proposed a compact wideband pattern reconfigurable antenna for 5G vehicular communication by integrating a reactive metasurface with a PIN-diode-controlled end-fire radiator. The antenna consists of an active dipole, two passive dipoles loaded with PIN diodes, parasitic patches, and a reactive metasurface placed beneath the radiator to realize a low-profile configuration (Fig. 2.22 (a) and (b)). By switching the ON/OFF states of the PIN diodes, the antenna generates three radiation modes with beam directions of -35° , 0° , and $+35^\circ$ with enhanced gain of 6.5 to 6.9 dBi, which are shown in Fig. 2.23.

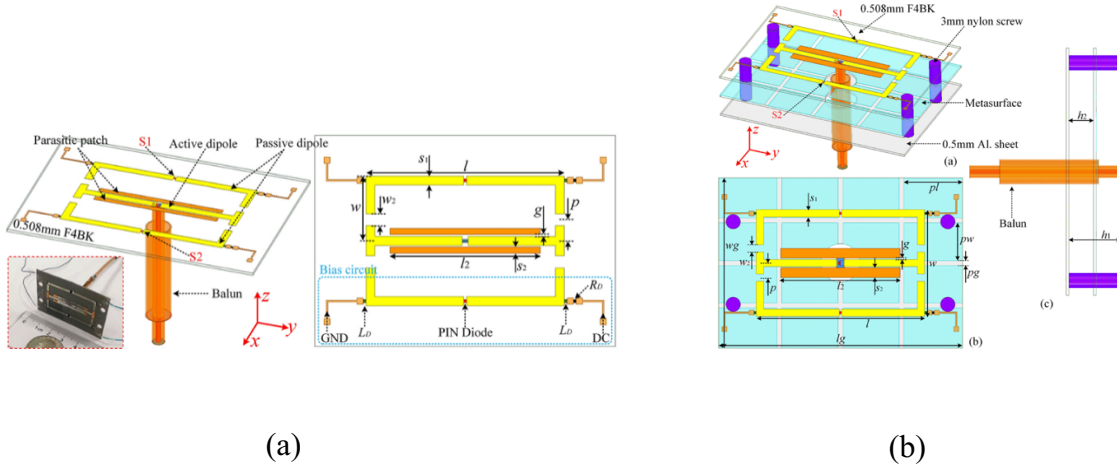


Fig. 2.22. (a) 3D view of antenna without AMC surface (b) 3D view of antenna with AMC surface [45]

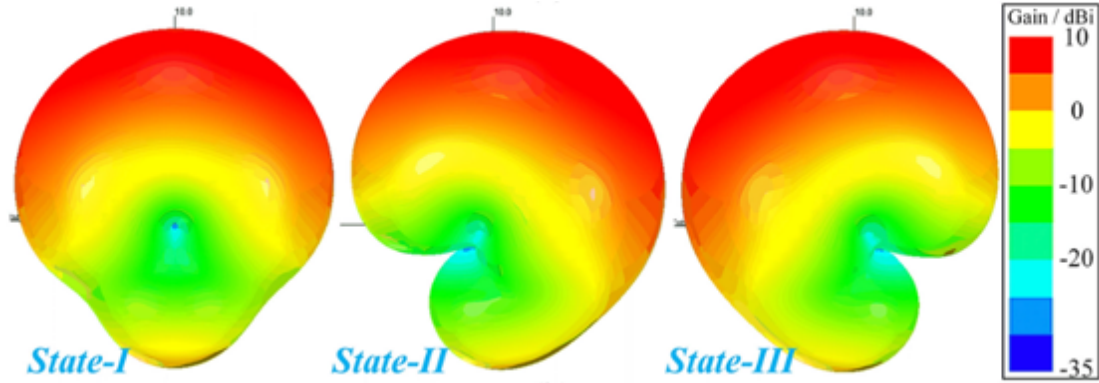


Fig. 2.23. Radiation pattern at 3.6 GHz for three states. (State-I: Both passive dipoles inactive; State-II: Left passive dipole active, right passive dipole inactive; State-III: Right passive dipole active, left passive dipole inactive) [45]

Therefore, in both [44] and [45], the AMC surface mainly acts as a supporting reflecting layer, while the beam steering is produced by modifying the radiating structure.

A slightly different variation of using metasurface is employing it as a superstrate rather than as a traditional AMC [46]. Although this design is slightly different from other antennas reported above, it still has a close synergy with the work that is to be undertaken in this thesis and would be apparent in the chapter to follow. The design consists of a 3×3 square metasurface, a ring-

shaped pixel structure, a square microstrip patch, a ground plane, and a simple feeding network which is shown in Fig. 2.24.

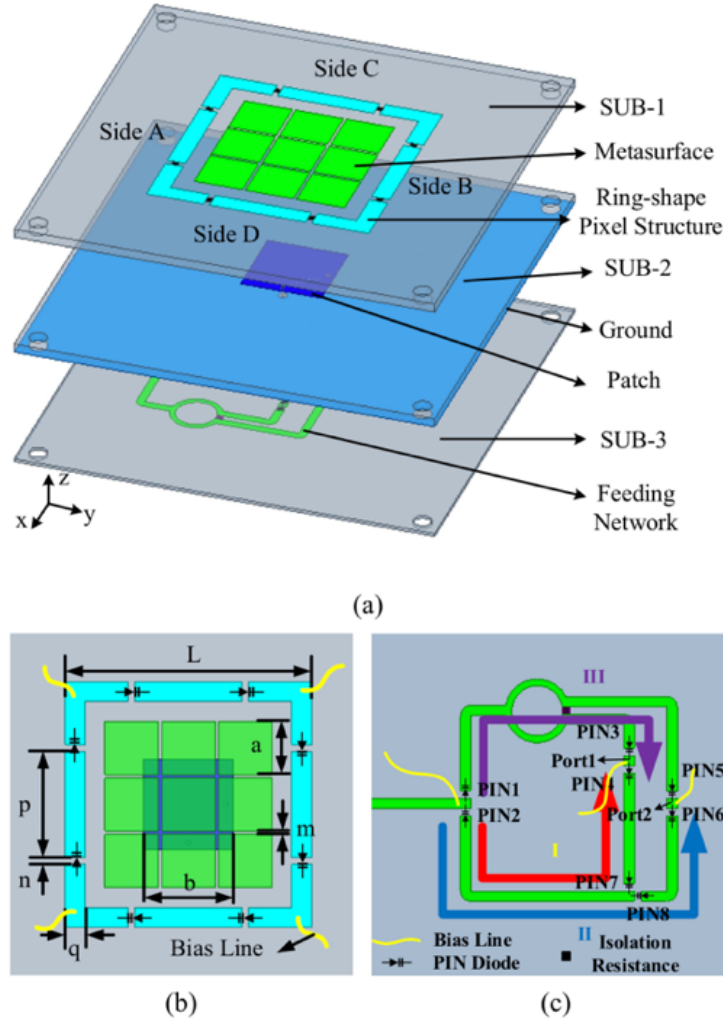


Fig. 2.24. Structure of the MTS antenna (a) Three-dimensional view (b) Top view (c) Bottom view [46]

All three dielectric substrates are stacked directly on each other with no gap between them. By exciting two orthogonal modes of the metasurface, the antenna can switch among horizontal linear polarization, vertical linear polarization, and circular polarization. In addition, the ring-shaped pixel structure loaded around the metasurface is used to control the radiation pattern direction. By changing the states of the PIN diodes on different sides of the pixel ring, the beam can be steered

to a certain degree. The measured results show beam deflections of $\pm 40^\circ$ for horizontal polarization, $\pm 20^\circ$ for vertical polarization, and $\pm 25^\circ$ for circular polarization which are illustrated in Fig. 2.25. One should keep in view, like the previous design this one also shows discrete steps for steering the antenna beam which is a limitation commonly seen in such designs.

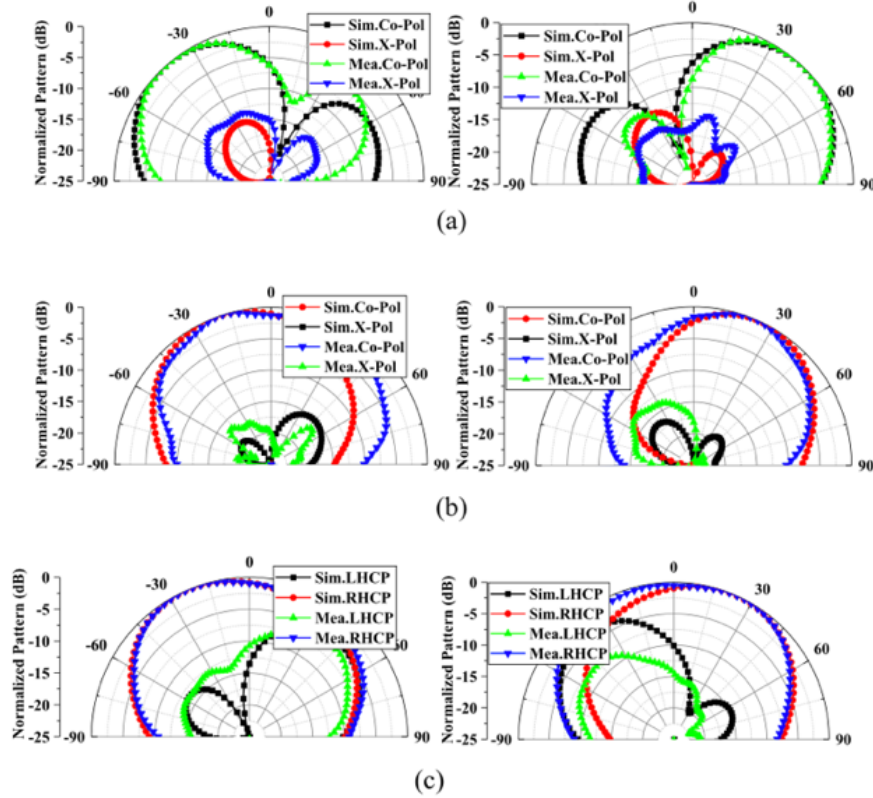


Fig. 2.25. Simulated and measured pattern in the yz plane. (a) HP (b) VP (c) CP [46]

With a series of research papers summarized in the previous section focusing on AMC- and metasurface-based reconfigurable antenna designs, the key features of these works are compiled in Table 2-1. As it is evident, some others have also been included in the table which are not discussed above for the sake of brevity. It can be observed that AMC and metasurface technologies have been extensively employed to enhance the radiation characteristics of antennas while maintaining compact and low-profile structures. Most of the reported designs utilize passive AMC

layers to improve the constructive interference between the radiated and reflected fields, resulting in higher antenna gain, larger impedance bandwidth, improved radiation efficiency, and reduced profile. The focus of this literature review is around AMC or metasurface based designs that are tunable or reconfigurable. Even among reconfigurable AMC based designs, the circle of discussion has been kept around designs that demonstrate beam-steering capabilities [43]–[49]. Different approaches have been adopted to achieve radiation pattern control, including PIN-diode controlled parasitic elements, transmissive coding metasurfaces, reactive metasurfaces, liquid crystal-based high-impedance surfaces, active reflection metasurfaces, and phase-gradient metasurfaces. These techniques successfully demonstrate beam steering ranging from approximately $\pm 20^\circ$ to $\pm 57^\circ$, while simultaneously improving antenna gain, bandwidth, or reducing radar cross section in some cases [43]–[49]. However, most of the reported beam-steering antennas rely on discrete switching states achieved using PIN diodes, which inherently limits the steering resolution and beam continuity. Although liquid crystal and varactor-controlled structures provide continuous tuning capability, their demonstrated steering range remains relatively limited and often requires relatively large or complex multilayer structures [47], [48]. In addition, none of the designs show that a narrow enough beam has been achieved that is steered around one of the principal planes. Therefore, a significant research gap still exists in realizing AMC/metasurface-backed antennas capable of providing wide-angle continuous electronic beam steering while simultaneously maintaining high gain, wide impedance bandwidth, and a simple biasing architecture. This limitation provides the primary motivation for the research presented in this thesis, where a field-programmable metasurface is investigated to achieve continuously steerable radiation characteristics using a compact single-element antenna configuration.

Table 2-1 Summary of Literature Review on AMC based Antenna Designs

(GE: Gain Enhancement, BS: Beam Steering, SR: Size Reduction, HP: Horizontal polarized, VP: Vertical polarized, CP: Circular polarized, BE: Bandwidth Enhancement)

Ref.	Year	Topology	Operating Frequency (GHz)	Scanning Angle (°)	Steering Resolution	Size (W × L × H) (mm ³)	Gain (dBi)	Antenna	Comments
[40]	2021	Rhomboid AMC-Backed Printed Dipole Antenna	4.48–7.00	–	–	48 × 48 × 6.8	8.4	Single Printed Dipole	GE (+4 dB), BE (43.9%), SR (42.3%), LP
[41]	2023	Giuseppe Peano AMC-Loaded Printed Dipole Array (5 × 5 AMC)	3.30–6.02	–	–	60 × 60 × 6.8	8.7	2 × 2 Printed Dipole Array	GE, BE (60%),
[42]	2023	AMC-Backed Antenna-on-Chip with Coupling Enhancement Structure (CES)	94	–	–	4 × 4 × 0.5	9.8	Single	GE (+4.3 dB),
[43]	2024	AMC-Backed Cross-Slot Antenna with Transmissive Coding Metasurface (TCMS)	3.09–3.70	±57°	Discrete (12 States)	144 × 144 × 31	8.47	Single Cross-Slot	BS, Beamwidth Reconfiguration (60°–140°), GE, Wide BW (17.96%),
[44]	2018	AMC-Backed Slot Antenna with PIN-Controlled Parasitic Slots	4.9–5.5	0°, ±36°	Discrete (3 States)	67.2 × 67.2 × 15	9.4 (Sim.), >7 (Meas.)	Single Cross-Shaped Slot	BS, AMC, Wide BW (11.54%), 2 PIN Diodes, Low Back Lobe
[45]	2022	Reactive Metasurface-Loaded Pattern Reconfigurable	3.30–4.20	35°, 0°, –35°	Discrete (3 States)	28.3 × 22.2 × 9.6	6.88 (Measured)	Single End-Fire Dipole	BS, Reactive Metasurface, Wide BW

Ref.	Year	Topology	Operating Frequency (GHz)	Scanning Angle (°)	Steering Resolution	Size (W × L × H) (mm ³)	Gain (dBi)	Antenna	Comments
		End-Fire Antenna							(24%), 2 PIN Diodes
[46]	2021	Compact Low-Profile Reconfigurable Metasurface Antenna with Polarization and Pattern Diversities	5.0	$\pm 40^\circ$ (HP), $\pm 20^\circ$ (VP), $\pm 25^\circ$ (CP)	Discrete	$39 \times 39 \times 4.0$	4.29 (Measured)	Single Metasurface Antenna	BS, LP/CP Switching, Low Profile, Compact, CMA, 16 PIN Diodes
[47]	2020	Tunable Liquid Crystal HIS-Based ESPAR Antenna	27.22–29.24	$\pm 25^\circ$ (Measured)	Continuous	$24 \times 15 \times 0.762$	6.92	Three-Patch ESPAR	BS, Continuous Beam Steering, Low DC Power, Simple Bias Circuit
[48]	2022	Dual-Polarized Antenna with Active Reflection Metasurface	3.30–3.60	$\pm 20^\circ$ (Measured), $\pm 30^\circ$ (Simulated)	Continuous	$91.6 \times 91.6 \times 8.3$	NR	Crossed Dipole	BS, Active Reflection Metasurface, Continuous Steering, Beamwidth Reconfiguration, $\eta > 88\%$, Varactor Controlled
[49]	2019	Reconfigurable Phase Gradient Metasurface (PGM)-Based Low-RCS Beam-Steering Antenna	4.20–4.55	$0^\circ, \pm 7^\circ, \pm 15^\circ, \pm 23^\circ$	Discrete (7 States)	$150 \times 120 \times 35.8$	6.2 (Measured)	Single Microstrip Patch	BS, GE (~3 dB), RCS Reduction (28 dB), Beam Focusing, PIN Diodes

2.3 Phase Shifter

Phase shifters are fundamental components in microwave and RF systems that introduce a controlled phase delay to a propagating EM signal while ideally maintaining a constant signal amplitude. They are extensively employed in phased-array antennas for 5G/6G base stations, satellite beamforming payloads, automotive radar, synthetic aperture radar (SAR), electronically steered airborne and naval radar systems, and electronically scanned imaging systems, where precise phase control is essential for beam steering, beam shaping, and wavefront manipulation. Depending on the system requirements, phase shifters may provide either continuous (analog) or discrete (digital) phase control, offering flexibility for a wide range of microwave and millimeter-wave applications.

The primary function of a phase shifter is to modify the phase of a transmitted or reflected signal without significantly affecting its magnitude. This controlled phase variation can be achieved by altering the electrical length or the effective propagation characteristics of the transmission path through various circuit topologies and implementation technologies. The performance of a phase shifter is typically characterized by parameters such as insertion loss, return loss, phase shift range, phase accuracy, bandwidth, switching speed, power handling capability, and power consumption. Since these parameters often exhibit competing trade-offs, the selection of an appropriate phase shifter architecture depends on the specific requirements of the intended application.

2.3.1 Reconfigurable Phase Shifters

Conventional phase shifters provide a fixed phase response determined during the design stage. However, many modern RF communication systems require dynamic phase control to adapt to changing operating conditions and system requirements. This need has led to the development of

reconfigurable phase shifters, whose phase response can be electronically or mechanically adjusted after fabrication through external control signals [50], [51]. Among these two, electronic tuning is more viable for obvious reasons of easy integration and simpler control. Also, the focus of this work centers around an electronically tunable phase shifter and therefore the literature review is kept around this method.

Electronically controlled phase shifters achieve tunability by altering the electrical characteristics of the transmission path using lumped controlled components or tunable materials. Common implementations employ PIN diodes, varactor diodes, RF microelectromechanical systems (MEMS), field-effect transistors (FETs), and other integrated circuits. In other examples, one can see the tunability achieved using functional dielectric materials such as the ones based on liquid crystals and ferroelectrics [50],[51], [52], [53]. By modifying the effective electrical length, characteristic impedance, or reactive loading of the microwave circuit, these devices enable either continuous or discrete phase tuning over a specified range. The performance of reconfigurable phase shifters is commonly evaluated using several key metrics such as, phase tuning range, insertion loss, input and output impedance matching, phase resolution, switching speed, linearity, power consumption, power handling capability, and operating bandwidth. Since improvements in one performance parameter are often accompanied by degradation in another, significant research efforts have focused on developing reconfigurable phase shifters that achieve an optimal balance between electrical performance, implementation complexity, fabrication cost, and reliability [51]–[54]. One easy way to cumulate these parameters into a single performance matrix is by defining a figure of merit (FoM). FoM is usually defined as phase shift per unit insertion loss or phase shift per unit length. These provide a comparison at a glance to the readers and would be used in this work.

2.3.2 Types of Reconfigurable Phase Shifters

It is known that phase shifters can be classified based on their operating principle, circuit topology, or implementation technology. From a circuit topology perspective, the most widely adopted phase shifters include switched-line, loaded-line, reflection-type and substrate integrated waveguide (SIW) based phase shifters. Even these topologies have been short-listed from a wide variety of phase shifters that are available in the literature. To keep this explanation concise and since FPMS based phase shifter relies on varactors, the following sub-section is composed of switched-line and loaded-line based designs because of the obvious synergy. In addition, some examples that rely on SIW or similar designs are also included as the resemblance would become evident during the discussion of the proposed design. In this way, a fair comparison can be built among the reported papers to conclude the direction of research and at the same time the discussion would be kept concise. Each topology achieves the desired phase shift through a different mechanism and exhibits distinct trade-offs in terms of performance parameters by using active elements. Consequently, the selection of a particular topology depends on the performance requirements of the intended RF application.

2.3.2(a) Switched-Line Phase Shifter

The switched-line phase shifter is one of the simplest and most widely adopted phase shifter topologies owing to its straightforward operating principle and ease of implementation. It consists of two or more transmission lines with different electrical lengths connected through RF switches which is shown in Fig. 2.26. By selectively routing the input signal through a transmission line of a particular length, the signal experiences a corresponding propagation delay, thereby producing the desired phase shift. The phase shift can be defined using Eq. (2.8).

$$\phi = \beta \cdot L \quad (2.8)$$

where ϕ is the phase of the RF signal, β is the phase constant and L is the length of line.

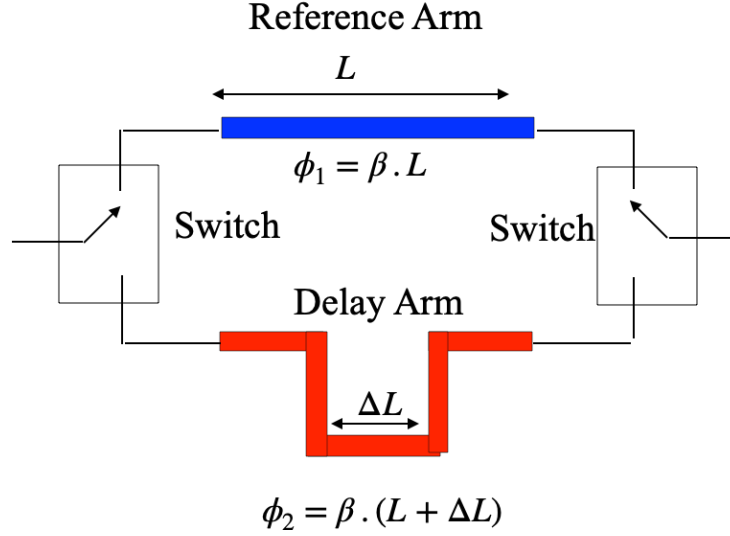


Fig. 2.26. Schematic design of Switched Line Phase Shifter

The differential phase shift is directly proportional to the difference in the electrical lengths of the transmission paths and the phase constant of the transmission medium. Switched-line phase shifters are commonly employed to realize large phase states, such as 90° and 180° , in multi-bit digital phase shifters [55], [56], [57]. For instance, in [58] the authors of the work proposed an improved Inverted-E loaded-line phase shifter to overcome the limited phase shift range and narrow bandwidth of conventional loaded-line phase shifters. The proposed topology replaces the conventional 50Ω transmission line with a line loaded by a fixed open-circuited stub, introducing four independent design variables, Fig. 2.27. This technique provides greater flexibility for impedance matching, bandwidth enhancement, and compact implementation.

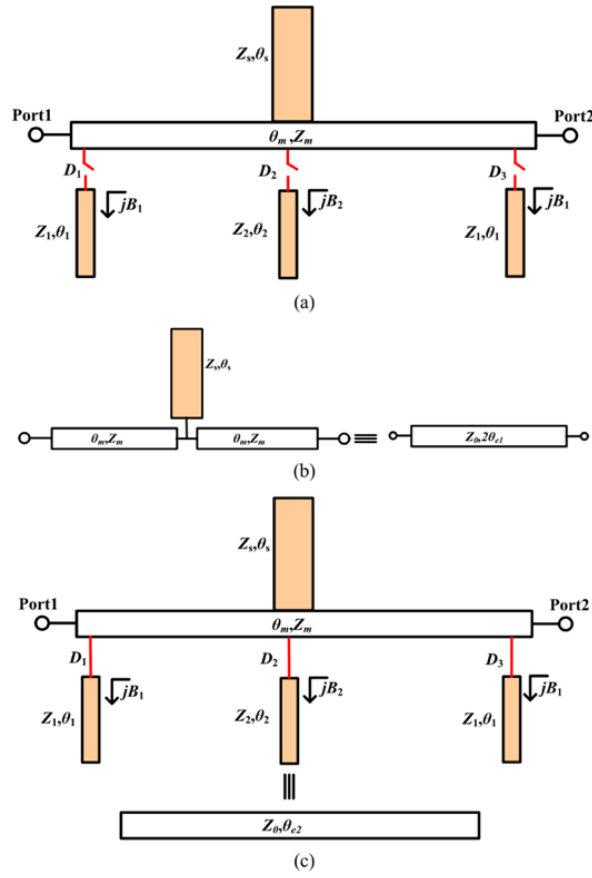


Fig. 2.27. Circuit configuration of the proposed Inverted-E type phase shifter. (a) Main block diagram (b) OFF-state equivalent circuit (c) On-state equivalent circuit [58]

Three symmetric open- or short-circuited stubs each controlled by individual PIN diodes are integrated to generate the required susceptance values for different phase states. The phase shift is produced by switching between the ON and OFF states of the diodes. An analytical design procedure based on even- and odd-mode analysis of the transmission is also proposed in this work to determine the circuit parameters for 22.5° , 45° , 90° , and 180° digital phase shifters. The simulated phase response for 4 different design cases are shown in Fig. 2.28. The authors chose the 90° and 180° phase shifter as examples to provide the proof-of-concept. The measured results show an insertion loss of 1.7 dB and 2.85 dB for the two implemented designs with a phase error of 0.3° and 0.6° at 2.4 GHz, respectively. Compared with conventional loaded-line phase shifters,

the proposed design provides a more compact structure, lower insertion loss, and an extended phase shift range while maintaining good phase accuracy. However, the digital nature of the phase shifter is quite apparent that minimizes the effectiveness of the proposed topology.

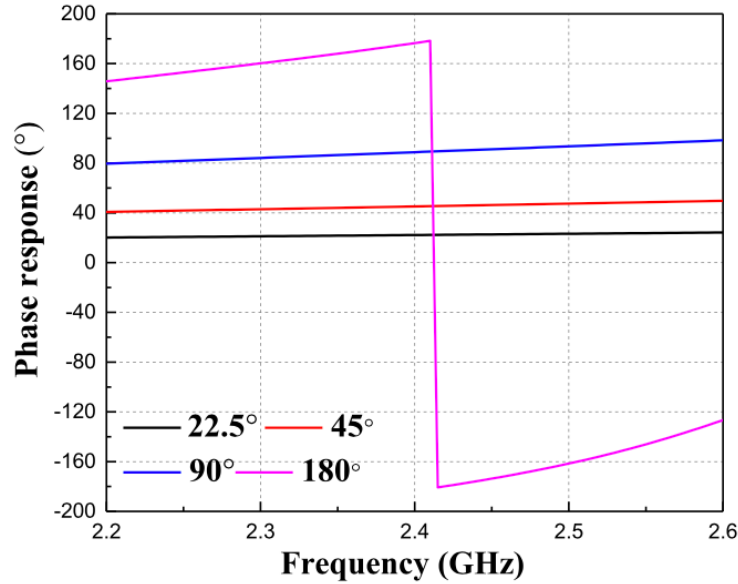


Fig. 2.28. Simulated phase response of proposed Inverted-E type phase shifter [58]

In [59], like [58] a more digital technique is used to tune the phase shifter. It is based on a modified switchable Schiffman phase shifter design principle. The key contribution is the development of a 4-bit digital phase shifter based on an improved single C-section Schiffman structure, where PIN diodes are strategically integrated along the coupled microstrip transmission lines to achieve wide bandwidth and low insertion loss, Fig. 2.29 and Fig. 2.30. To improve impedance matching and phase slope alignment, the authors incorporated $\lambda/8$ open- and short-circuit stubs into each phase bit. The complete phase shifter consists of 22.5°, 45°, 90°, and 180° bits, providing 16 discrete phase states spanning over the range of 0°–360°, while avoiding the tight coupling required in conventional Schiffman phase shifters. The prototypes of the design along the measured results are shown in Fig. 2.31 and Fig. 2.32, respectively. The proposed phase

shifter is designed, fabricated, and experimentally validated on an RO4003 substrate operating over the frequency range of 2.2–3.3 GHz. The downside of the design similar to [58] is the discrete nature of the results which limits its use in certain phased array applications where a continuous phase shifting mechanism may be needed.

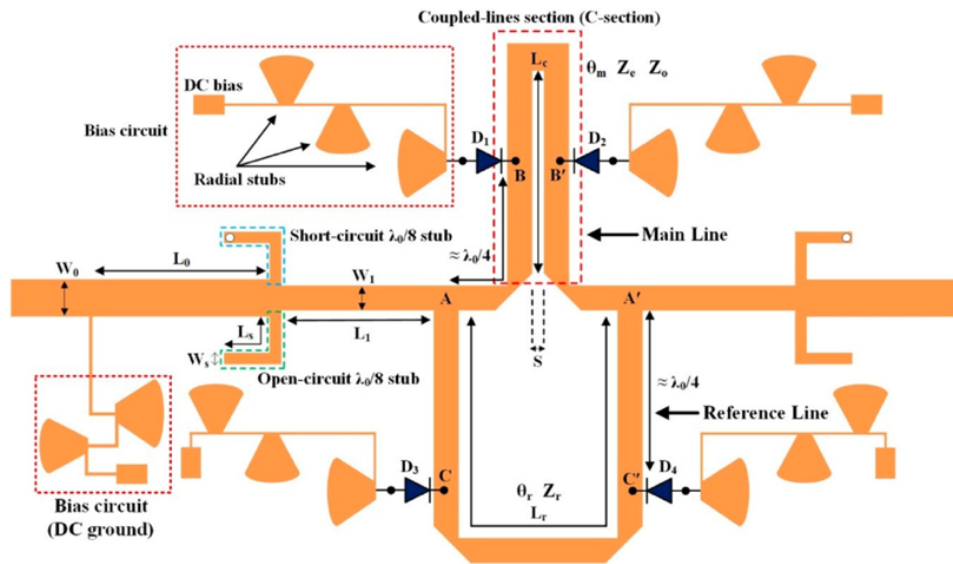


Fig. 2.29. Proposed switchable structure of 22.5°, 45°, and 90° bits [59]

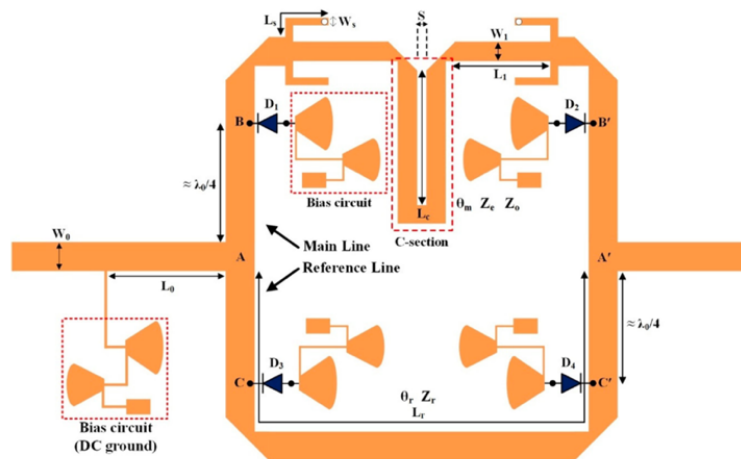


Fig.2.30. Switchable structure of 180° bit [59]

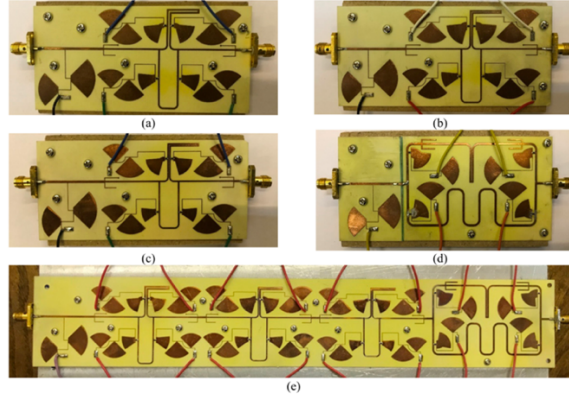


Fig. 2.31. The fabricated prototypes of (a) 22.5° bit (b) 45° bit (c) 90° bit (d) 180° bit and (e) 4-bit phase shifter [59]

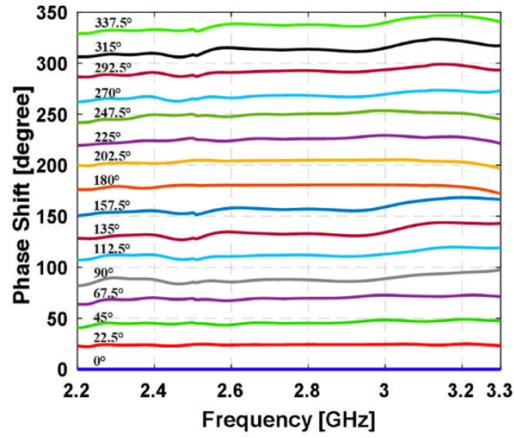


Fig. 2.32. Measured results of phase shifter [59]

2.3.2(b) Loaded-Line Phase Shifter

The other popular category in electronic phase shifters relies on loading an impedance structure with lumped component elements. The structure in-turn modulates the impedance experienced by the propagating wave, hence providing a distinct control on the phase response. The reactive loading elements are usually inductors or capacitors that are integrated along a transmission line to do the job. Compared with switched-line phase shifters, loaded-line phase shifters generally offer a more compact structure, lower insertion loss for small phase shifts, and easier integration into monolithic microwave integrated circuits (MMICs). However, the achievable phase shift is

limited by the amount of reactive loading, and larger phase shifts often require cascading multiple sections, which increases design complexity, size and insertion loss.

A first of these examples is a 5-bit digital loaded-line phase shifter (LLPS) proposed by Liu *et al* [60]. The design relies on the phase-tuning property of an open-/short-circuit microstrip line-loaded slot-line structure to achieve a full 360° phase shift range while maintaining a compact circuit size, Fig. 2.33. Unlike conventional loaded-line phase shifters that employ quarter-wavelength transmission lines and are generally limited to small phase shifts, the proposed design utilizes a slot line as the main transmission path with switchable open- and short-circuit microstrip stubs to modify the effective electrical length. The 5-bit digital phase shifter consisting of 11.25° , 22.5° , 45° , 90° , and 180° phase bits is used for concept demonstration. The measured results of the complete prototype (Fig. 2.34) show more than acceptable matching conditions within the band of operation while providing a phase tuning range of 360° . An insertion loss of 3 dB is observed for most of the measured results around the center frequency of 2 GHz with stable matching condition, satisfying the operational requirements.

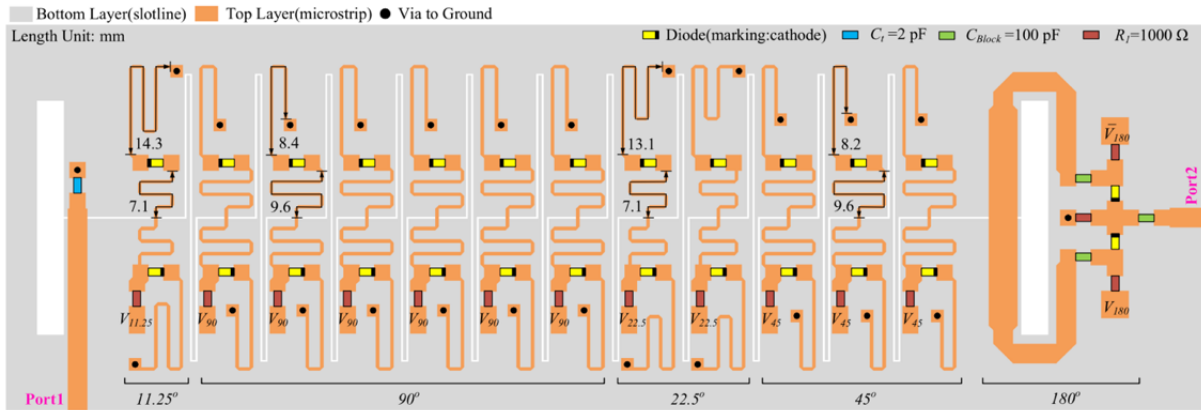


Fig. 2.33. Layout of the proposed 5-bit digital LLPS using the phase-tuning property of open/short circuit microstrip line-loaded slot line structure [60]

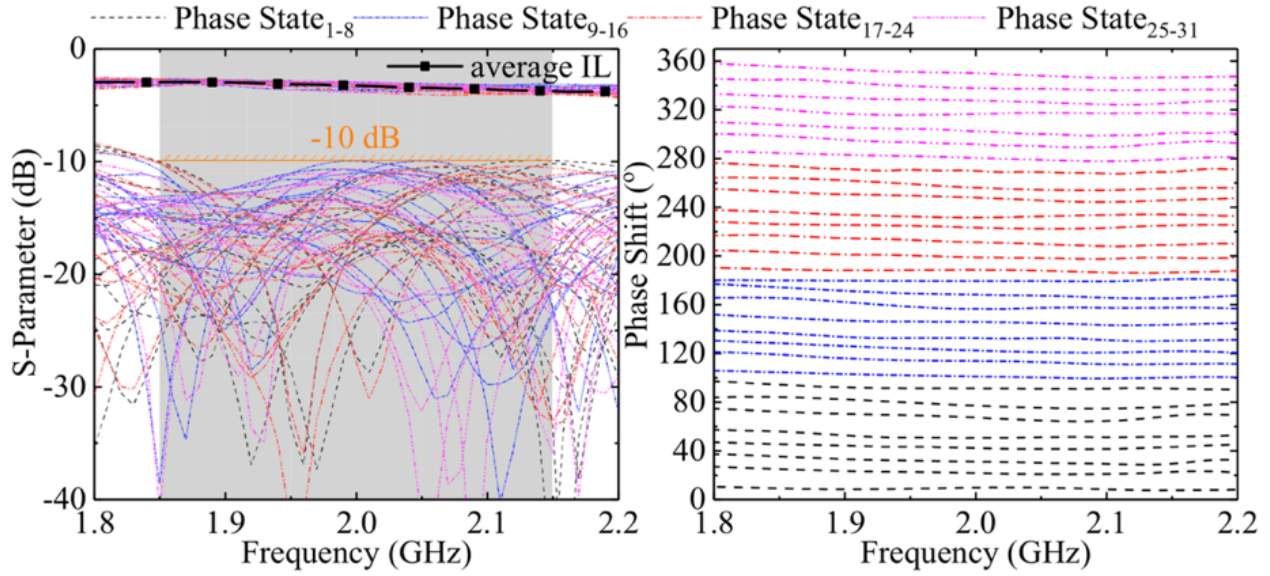


Fig. 2.34 Insertion loss and Phase range [60]

Remarkably, authors in [61], proposed a voltage-controlled tunable loaded-line phase shifter based on few-layer graphene flakes for microwave applications. This is novel and challenging since graphene-based designs have their own set of challenges even at low frequency. The proposed structure consists of a 50Ω microstrip transmission line loaded with a short-circuited stub, where a graphene pad is inserted between the stub and the main transmission line. By varying the applied bias voltage, the resistance of the graphene changes, thereby modifying the input impedance of the stub and producing a continuous phase shift. The authors optimized the stub dimensions using full-wave Ansys HFSS simulations to maximize the input reactance variation while minimizing insertion loss. Fig. 2.35 and 2.36 illustrate the proposed graphene-loaded topology and its equivalent circuit. The proposed phase shifter is fabricated on a Taconic RF-35 substrate and experimentally validated over the 5–6 GHz frequency range with a modest phase shift of 35° – 43° and acceptable insertion loss, Fig. 2.37. The final test setup is presented by the authors as Fig. 2.38. The use of bias-tee is critical in such a design where separate bias lines have

been avoided within the structure. This technique is generally not recommended but simplifies the overall design for implementation.

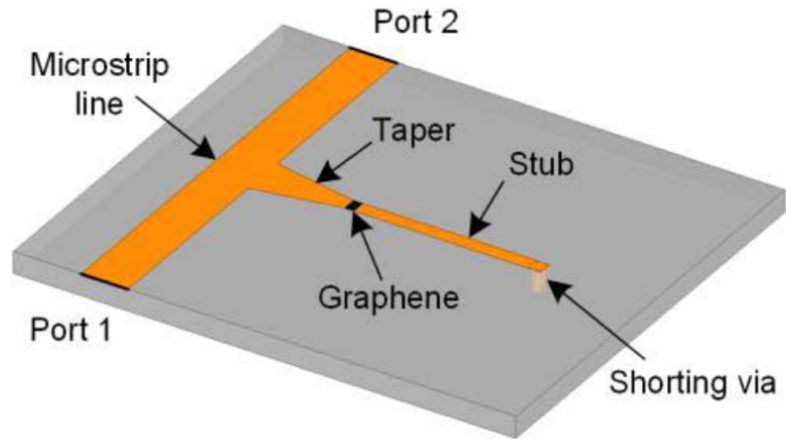


Fig. 2.35. Geometry of graphene-based tunable phase shifter [61]

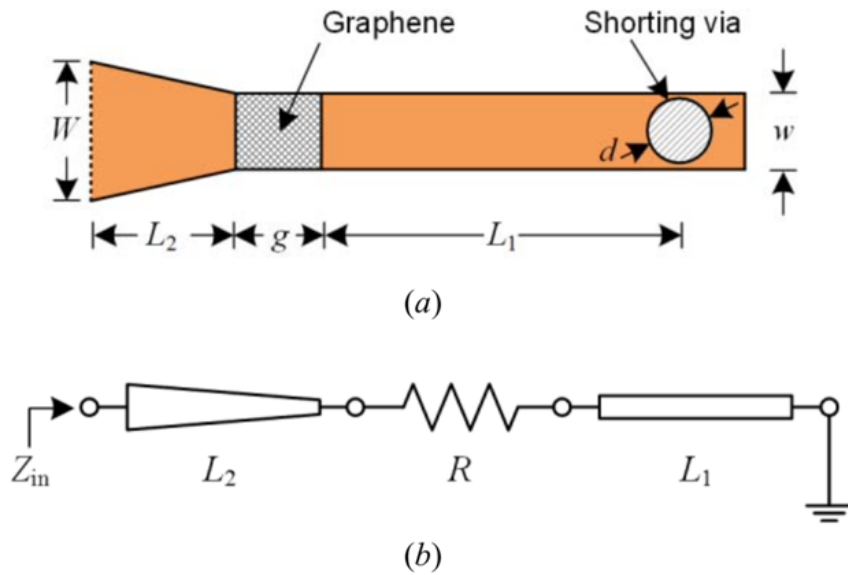


Fig. 2.36. (a) Geometry of stub (b) Equivalent circuit [61]

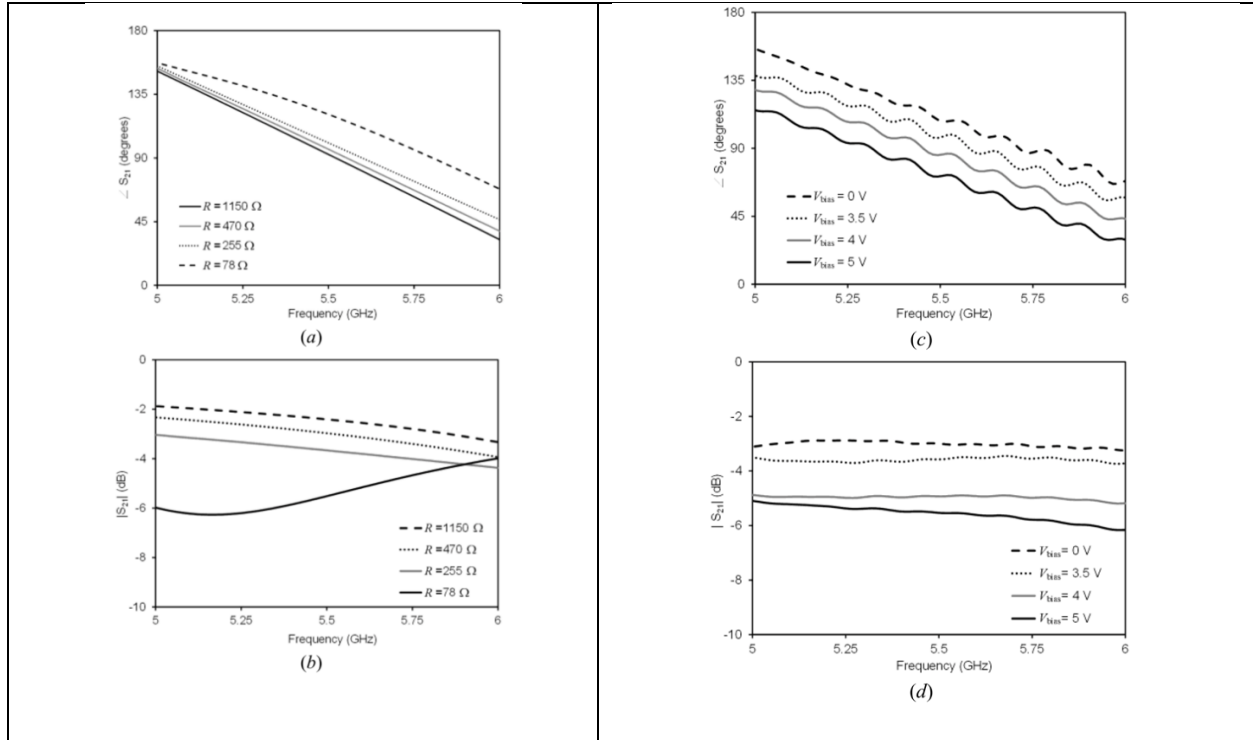


Fig. 2.37. Simulated and measured scattering parameters of the phase shifter. (a) Simulated phase of S_{21} . (b) Simulated amplitude of S_{21} . (c) Measured phase of S_{21} . (d) Measured amplitude of S_{21} [61]

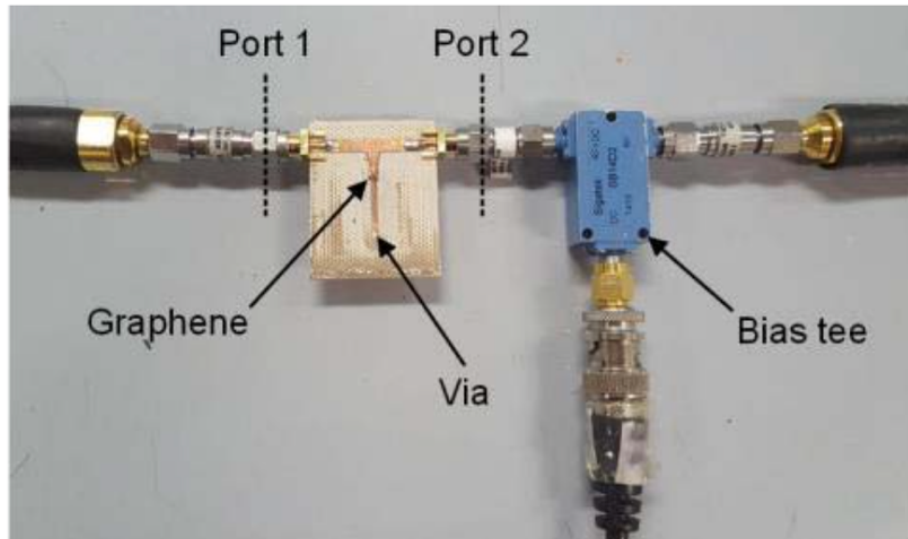


Fig. 2.38. Measurement setup of the tunable phase shifter, with the position of the reference ports [61]

2.3.2(c) SIW based Phase Shifters

The third class of electronically tunable phase shifters included in this discussion are waveguide or SIW based designs. Such components have emerged as an attractive alternative to conventional transmission-line phase shifters because they combine the low insertion loss and high power-handling capability with the compactness and low fabrication cost, qualities that are always desirable. In SIW structures, EM waves propagate inside a synthesized rectangular waveguide-like structure formed by two rows of metallized vias embedded within a dielectric substrate. Since the guided phase depends directly on the propagation constant of the waveguide, the phase response can be controlled by modifying the effective propagation path or by perturbing the EM field distribution using reconfigurable elements such as PIN diodes, varactor diodes, RF MEMS switches, or tunable dielectric materials. Although these designs have low insertion loss, one downside typically associated with these phase shifters is their larger sizes. However, the ease of integration and other above-listed merits easily out-weigh the large size.

There are ample examples of SIW based designs available in the literature. Here, the discussion is more centered around tunable designs. The first of this kind is the digital phase shifter proposed in [62] that employs a reconfigurable waveguide architecture capable of electronically modifying the effective width of the SIW structure. Unlike conventional SIW phase shifters, where the waveguide dimensions remain fixed after fabrication, the proposed design introduces an additional row of metallic vias controlled by 35 pairs of PIN diodes, enabling the waveguide width to switch between two predefined states. As illustrated in Fig. 2.39, the propagation path is divided into multiple sections where the effective width is dynamically controlled through the ON/OFF states of the PIN diodes. Consequently, the EM wave experiences two different propagation (phase) constants, and the accumulated phase delay is obtained by controlling the lengths of the narrow

and wide SIW sections. The measured results demonstrate a 35-bit phase resolution with 36 operating states, providing a total measured phase tuning range of approximately 262° at 4.3 to 5 GHz with a 7.3° phase step between adjacent states, Fig. 2.40. The proposed phase shifter maintains a reflection coefficient below -10 dB over the operating band, while the measured insertion loss varies from approximately 1.2 dB to 5 dB at 4.3 to 5 GHz depending on the selected state due to the increasing number of forward-biased PIN diodes.

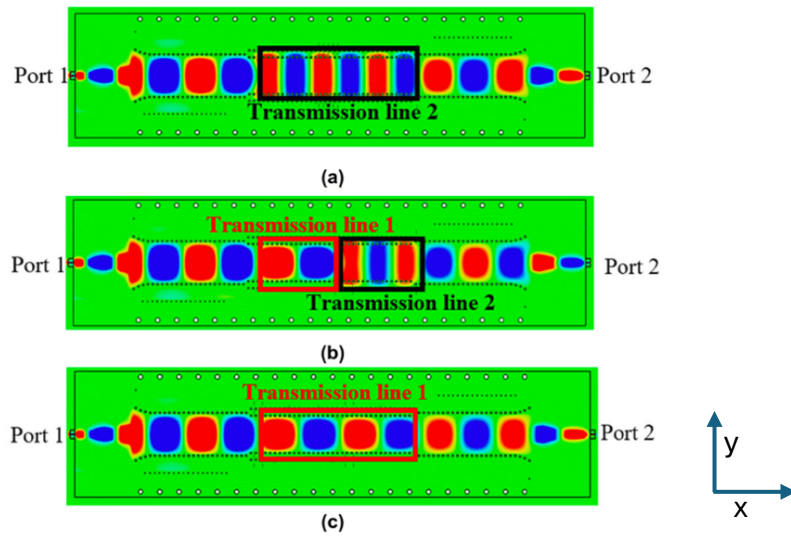


Fig. 2.39. E-field distribution on xy-plane (a) State 1 (b) State 18 (c) State 36. (State is activation status of PIN diodes) [62]

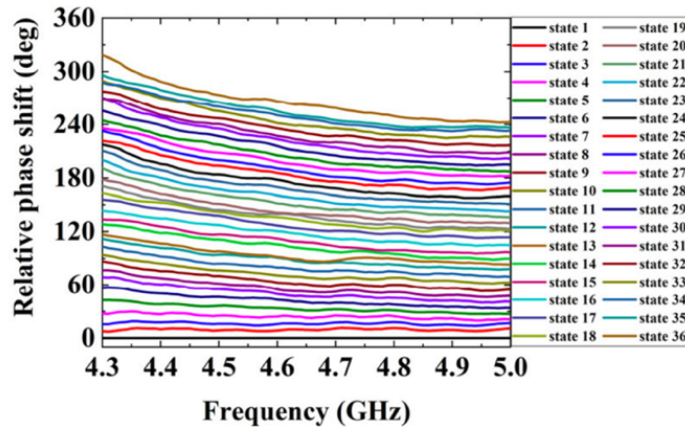


Fig. 2.40 Measured phase shift [62]

Another SIW-based reconfigurable phase shifter is presented by Louati *et al.* [63], where the phase tuning mechanism relies on parallel coplanar waveguide (CPW) stubs loaded with surface-mount PIN diodes onto the waveguide structure. The proposed design uses three diode-loaded CPW stub sections along the SIW component, allowing different phase states to be obtained by changing the ON/OFF combinations of the PIN diodes. As shown in Fig. 2.41, each CPW stub is connected to the SIW through a metallized via, forming a shunt RLC loading network.

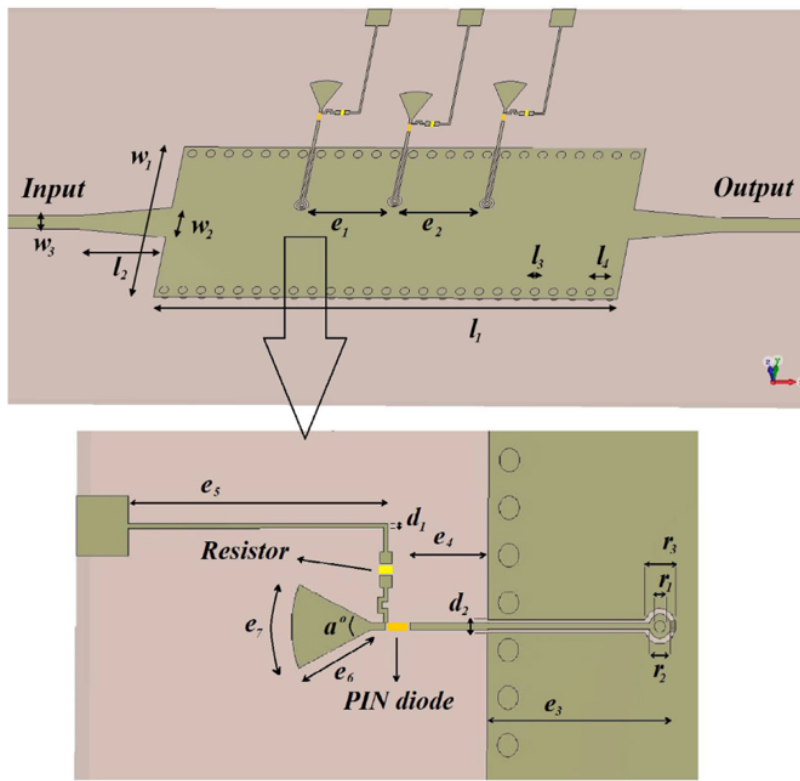


Fig. 2.41. Geometrical topology of design [63]

When the diode state changes, the electrical loading introduced by the stub also changes, thereby altering the current distribution and the effective propagation (phase) constant of the guided wave. Consequently, the transmitted signal accumulates a different phase delay while propagating through the SIW. The prototype provides eight discrete phase states corresponding to the different switching combinations of the three PIN diodes (Fig. 2.42). The measured results

demonstrate a reflection coefficient better than -10 dB and an insertion loss below 1.5 dB across the operating frequency range. Furthermore, the phase shifter achieves discrete phase shifts ranging from 8° to 60° with a phase error below 5° , confirming stable phase performance over the entire bandwidth.

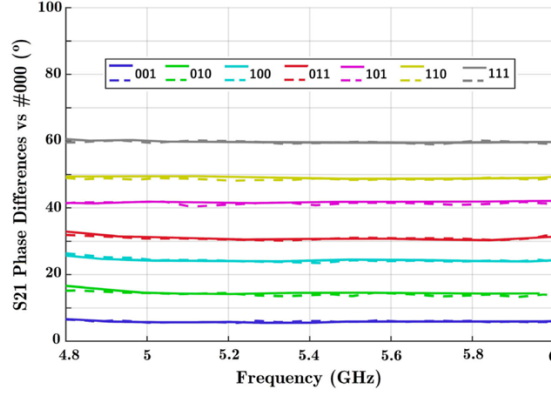


Fig. 2.42 Simulated and measured phase differences of phase shifter [63]

Similarly, SIW-based phase shifter is proposed in [64], where the phase shift is achieved using a half-mode substrate integrated waveguide (HMSIW) operating in the evanescent mode. Instead of varying the effective waveguide width or introducing tunable loading elements, the proposed design employs complementary series and shunt capacitive loading to reduce the cutoff frequency of the HMSIW while maintaining a compact footprint. As shown in Fig. 2.43, an interdigital capacitor is introduced along the reference transmission line, whereas blind vias loaded with metallic patches are incorporated into the main transmission line. These capacitive loading elements modify the propagation characteristics of the guided wave and produce complementary phase responses. For validation, the authors fabricated 45° and 90° phase-shifter prototypes, as shown in Fig. 2.44. The measured results demonstrate phase shifts of $43.5^\circ \pm 2.5^\circ$ and $86.5^\circ \pm 2.5^\circ$, with insertion losses of 0.40 – 0.63 dB and 0.76 – 1.32 dB, respectively. Furthermore, all channels exhibit a return loss better than -10 dB over the operating bandwidth.

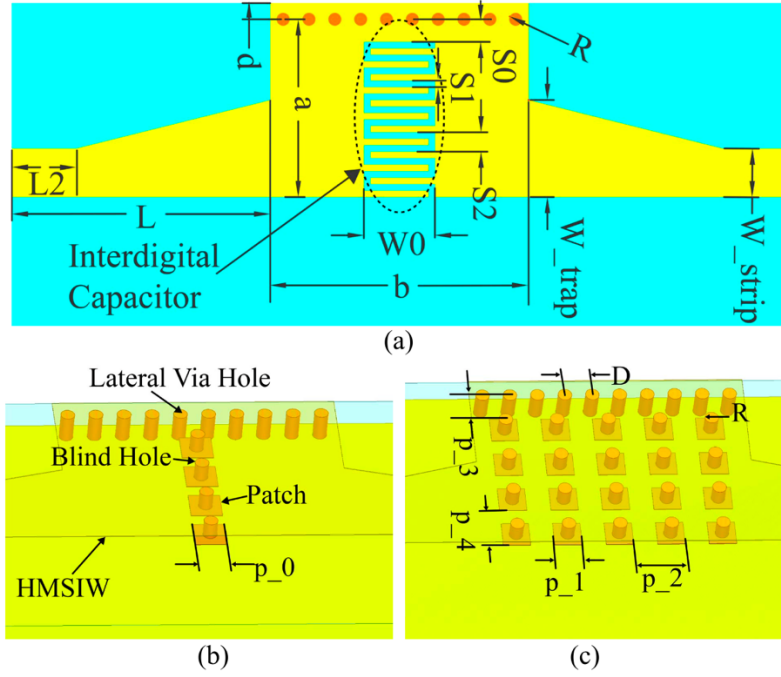


Fig. 2.43. Topology of modified HMSIWs. (a) HMSIW for reference line (b) HMSIW with four patches for 45° branch (c) HMSIW with 20 patches for 90° branch [64]

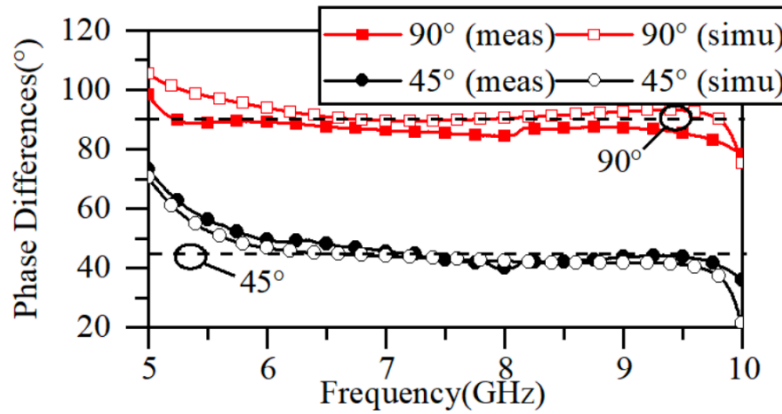


Fig. 2.44. Measured and simulated phase shift [64]

All papers related phase shifter are summarized in Table 2-2, which also includes the reflection-type phase shifter as well. The discussion about reflection type phase shifter is excluded because that kind of phase shifter doesn't synergize with proposed phase shifter in terms of design.

Table 2-2 Summary of Literature Review on Phase Shifters

Ref.	Topology	Operating Frequency (GHz)	Phase Shift Range	Insertion Loss	Return Loss	Bandwidth	Control	FoM (°/dB)
[58]	Switch-line	28 & 82	4-bit, 360°	11.6 dB / 11.9 dB	>10 dB	53.6%, 45.1%	CMOS switches	30.3
[59]	Switch-line	2	90° & 180°	1.7 dB / 2.85 dB	>10 dB	400 MHz / 100 MHz	PIN diodes	63.2 (180° case)
[60]	Loaded-line	1.85–2.15	5-bit, 360°	3.26 dB	>10 dB	15%	PIN diodes	110.4
[61]	Loaded-line	5–6	35°–43°	2–6 dB	-	20%	Graphene resistance tuning	≈7.2
[62]	SIW-based	4.65	262° (7.3° step, measured)	1.2–5 dB	>10 dB	4.3–5.0 GHz (~15%)	PIN diodes	52.4
[63]	SIW-based	4.8–6.0	60° (8°, 16°, 25°, 32°, 40°, 50°, 60° states)	<1.5 dB	>10 dB	4.8–6.0 GHz (~22%)	PIN diodes	40
[64]	SIW-based	5.8	45° and 90° (two prototypes)	0.40–0.63 dB (45°), 0.76–1.32 dB (90°)	>10 dB	5.15–6.15 GHz (≈17.7%)	(≈17.7%) Fixed (Passive Capacitive loading)	≈71°/dB (90° prototype)
[65]	Reflection-type	3.5	132.7°	8.9–13.7 dB	>10 dB	200 MHz (~5.7%)	Varactor + MOSFET	9.7
[66]	Reflection-type	1.88 & 2.44	114.1° / 114.0°	<1.86 dB / <1.89 dB	>17 dB	100 MHz	Varactor diodes	≈60.7

Ref.	Topology	Operating Frequency (GHz)	Phase Shift Range	Insertion Loss	Return Loss	Bandwidth	Control	FoM (°/dB)
[67]	Reflection-type	9–11	360°	2.1 ± 1.3 dB	>10 dB	20%	Varactor diodes	≈ 105
[68]	Switch-line	2.2–3.3	4-bit, 360°	<4 dB	>9.5 dB	40%	PIN diodes	>90
[69]	Loaded-line	6–10 (GaAs), 8–12 (GaN)	40°–165°	<6 dB / <8 dB	>10 dB	40–50%	Reverse-biased transistors	≈ 20.6

Overall, the reviewed literature demonstrates that both reconfigurable AMC and phase shifters have been extensively investigated for electronic beam steering in modern antenna systems. Various AMC configurations have been reported to manipulate the reflection phase for controlling the radiation characteristics, while switched-line, loaded-line, and reflection-type phase shifters have been developed to achieve the desired phase control with different trade-offs in phase range, insertion loss, bandwidth, and implementation complexity. These studies provide a comprehensive understanding of the existing beam-steering techniques and establish the foundation for the research presented in the subsequent chapters. Building on the concepts reviewed in this chapter, the next chapter introduces the proposed FPMS. The design methodology, analytical modeling, fabrication, and measurement results of the proposed structure are presented in detail.

Chapter 3

Beam Steerable Patch Antenna Integrated with FPMS layer

This chapter presents the design and implementation of a beam-steerable patch antenna integrated with an FPMS layer operating as a reconfigurable AMC. It begins with the characterization and analytical modelling of the FPMS unit cell under plane-wave excitation, followed by its integration with a semi-circular patch antenna. The chapter then describes the biasing strategy, fabrication, and experimental characterization of the proposed prototype. Finally, the simulated, analytical, and measured results are compared to validate continuous beam steering in both the elevation and azimuthal planes.

3.1 Design Concept

The three-dimensional view of the proposed antenna design is presented in Fig. 3.1. A brief overview of the complete antenna structure is illustrated herewith. It consists of 2 layers of substrates that are vertically stacked in the defined sequence. Rogers 5870 has been chosen for the antenna realization due to its low dielectric constant ($\epsilon_r = 2.3$) and loss tangent ($\tan\delta = 0.0009$) values. The top substrate layer is allocated for the semi-circle shaped patch antenna. The design is not restricted to a particular antenna shape, and similar behavior can be achieved with other radiator geometries, although the resulting performance depends on their field distribution and coupling with the FPMS. Therefore, some optimization would be required if a different shape of antenna is used. FPMS layer consisting of the unit cells are sandwiched between the antenna and the ground plane. The unit cell consists of a plus-shaped metallic arrangement with a capacitor loaded at the point of intersection, Fig. 3.1(b). The capacitor is realized in the actual scenario with the help of varactor diodes to provide tunability.

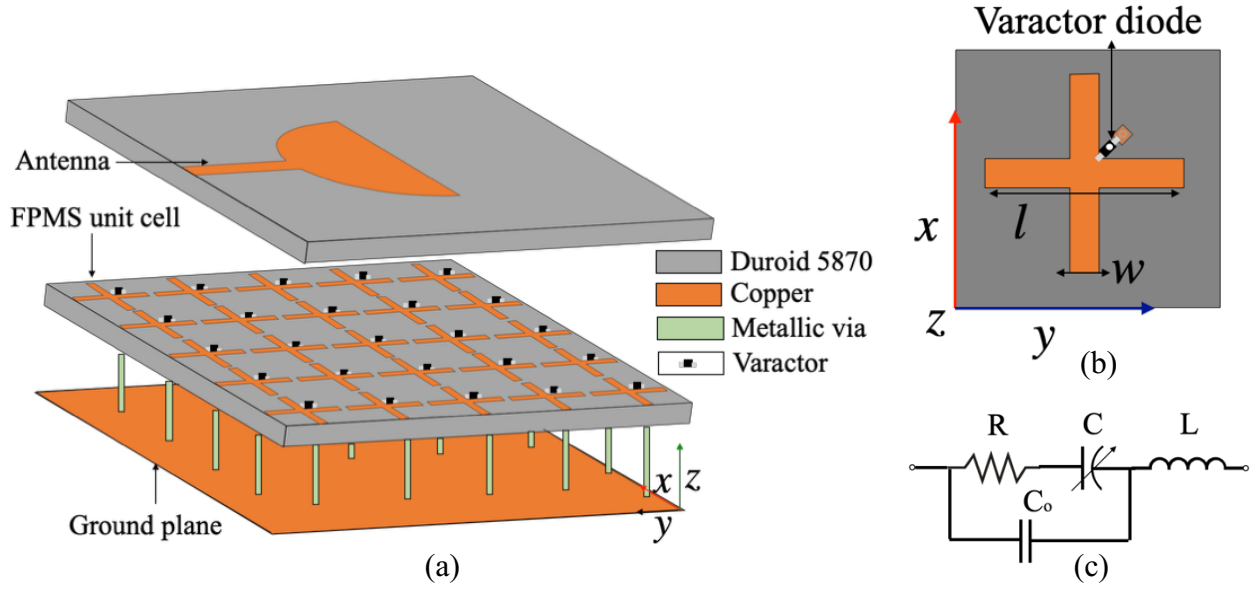


Fig. 3.1. Schematic design of HMSIW LWA (a) Perspective view of whole prototype (Vias are extended for better understanding) (b) Top view of FPMS unit cell ($l = 9$ mm, $w = 2$ mm) (c) Varactor diode equivalent RLC circuit

MACOM 46580-1209 is short-listed for this design because of their successful demonstration around X-band frequencies [34]. The equivalent circuit of the varactor comprises an inductor, $L = 0.105$ nH, connected in series with a parallel combination of two branches. One branch consists of a series resistor, $R = 2.2 \Omega$, depicting the varactor loss in conjunction with the variable capacitor (Fig. 3.1(c)), providing the main functionality [70]. The other shunt arm shows the nominal capacitance value of the varactor, $C_0 = 1.56$ pF. This nominal value occurs at no bias voltage. The total varactor capacitance is shown in Eq. (3.1) where $V_0 = 5.755$ V and $\psi = 1.96$. By controlling the bias voltage V , one can alter the capacitance of the varactor diode.

$$C(V) = \frac{C_0}{\left(1 + \frac{V}{V_0}\right)^\psi} \quad (3.1)$$

For RF modeling in the full-wave simulator, the varactor is realized by using its lumped circuit model with the DC bias being modeled through a via to RF ground. These vias are visible in the

isometric view of Fig. 1, extending from the FPMS layer to the ground plane underneath. The substrate thicknesses of layers 1 and 2 are 2.381 mm and 1.854 mm, respectively, which are intentionally preferred, keeping in view the commercial size available from the supplier. These thickness values play a significant role in the AMC design and have been optimized with the help of full-wave simulations. A key benefit of this design strategy is that the FPMS unit cells require DC biasing for the varactors, while the RF wave has to be injected into the antenna element. These two structures are physically separated from each other, resulting in an inherent isolation between the two types of excitations. This aspect of the design is further explained in the planning of the antenna prototype. At this stage, for readers' benefit, it is better to rely on the simpler structure presented in Fig. 1 and the same is used in the antenna simulator. The functioning of the unit cell is critical to understand the antenna performance and is therefore studied separately from the antenna in the next section before its integration in the proposed vertical stack up which is then followed by the antenna design and its integration with the FPMS layer.

3.2 FPMS Unit Cell Characterization and Dielectric Profiling

The analysis of FPMS in the existing literature has predominantly relied on simulation framework based on lumped port excitation, both in a circuit simulator (Advanced Design System (ADS)) and full-wave simulator (HFSS). However, no prior work has explored the simulation of these unit cells under the influence of a plane wave with a periodic boundary condition around the substrate. This aspect is critical for this design as the authors intend to use it as an AMC and needs to be evaluated before its use with the antenna system.

3.2.1 FPMS Unit Cell as an AMC

AMC layers are deemed to be infinitely large in area, at least in theory. Practically, the notion of a boundless structure is not possible. Thus, a modeling technique is imperative that could properly translate the behavior of an AMC layer from simulations to practice. This is accomplished by employing periodic boundary conditions around the substrate with a plane wave excitation on the structure from the top. The idea is to record the reflective abilities of the AMC layer as it provides high impedance to the incoming wave. This methodology is very well documented in literature and does not merit a detailed discussion here [31]-[37]. The unit cell simulated for this case is shown in a magnified image in Fig. 3.1(b) and follows the same process as the literature. At resonance, the AMC reflects part of the incident wave whose magnitude and phase are both of significance to the design. Ideally, all of the incident wave should be reflected back; however, this is highly unlikely in an actual scenario. More importantly, the focus is on the phase of the reflected wave. The closer the reflected phase of the wave is to 0° , the better it is for the gain enhancement of the antenna. With this pretext, the simulation results of the unit cells under the plane wave excitation are shown in Fig. 3.2. At first, the structure is simulated without any varactor or capacitor to optimize the design dimensions. It can be seen that the cell works well at 10 GHz with its dimensions captioned in Fig. 3.1, reflecting a portion of the incoming wave with a phase of 0° . These results can be further optimized to reduce the magnitude of transmission, but the values achieved here show ~65% of wave reflection, which is suitable as the starting point.

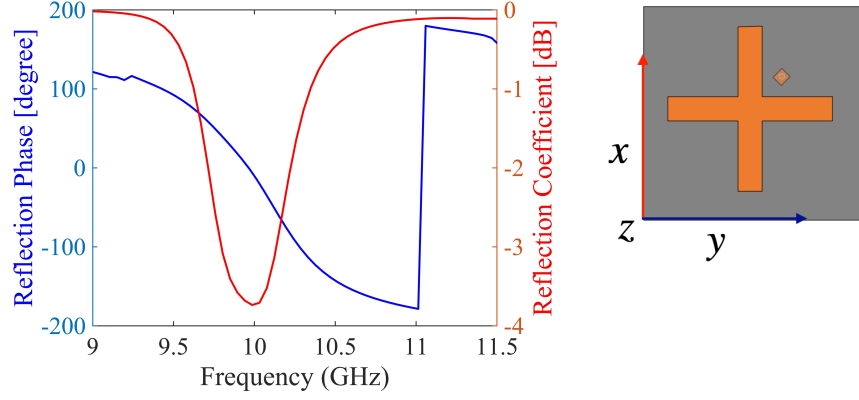


Fig. 3.2. Simulated reflection magnitude and phase of FPMS unit cell without capacitance

As the varactor diode is integrated onto the proposed structure, its resonance frequency alters. This allows for the reconfigurability of the structure, effectively altering the behavior of the FPMS as an AMC to achieve dynamic functionality. To study this, the capacitance of the diode is varied from 0.1 pF to 1 pF, Fig. 3.3.

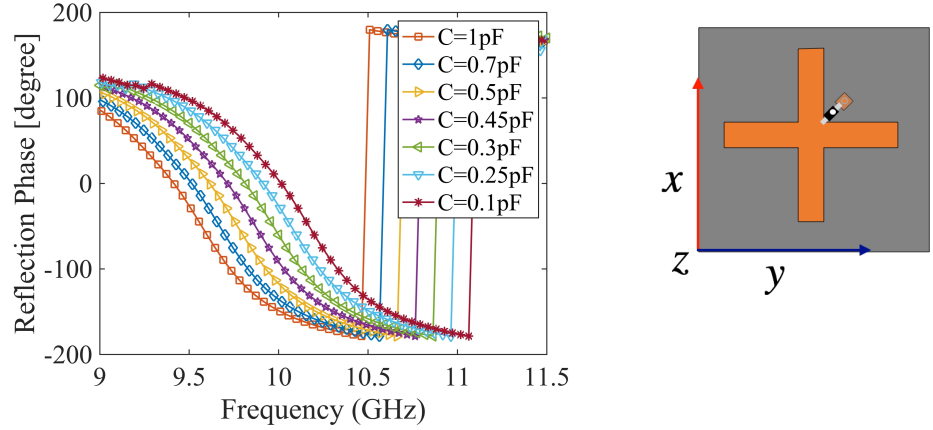


Fig. 3.3. Simulated reflection phase of FPMS unit cell with varying capacitances

As the capacitance of the varactor diode increases, the resonance frequency of the AMC shifts downwards to lower frequency bands. This shift occurs because a higher capacitance increases the overall reactive impedance of the AMC structure, lowering the frequency at which the reflection phase crosses 0° . Essentially, increasing the varactor's capacitance makes the AMC structure more responsive at lower frequencies, allowing for dynamic tuning of its operational range. It is

noteworthy to mention it here that the range of the capacitor is chosen on the basis of the practical varactor diodes that are available at 10 GHz [70]. Higher variation of capacitance can be used; however, as a starting point, this range is considered to be good enough to provide the proof-of-concept while keeping the discussion concise. These results are quite encouraging as they show that the behavior of the AMC structure can be controlled by integrating varactor diodes on it. Having written this, the impact of this change on the antenna design is not very clear from these simulations. The reason is that the individual unit cells under the antenna can be assigned different capacitor values, which means the AMC layer can't be modeled as an infinite or even a finite sheet of the same unit cells. Thus, the authors believe that the unit cell should be studied for its impedance behavior alongside these simulations to better understand its operation as an FPMS and to translate its performance onto the antenna.

3.2.2 Unit Cell Simulations under Plane Wave Excitation

The unit cell used for the FPMS topology has been studied before for impedance characterization [31], however, its study under a plane wave Floquet port excitation is on the shallow end. Thus, the natural inclination for this work is to study the proposed unit cell in such a setup where the unit cell is bounded by periodic conditions with a Floquet port at the top, Fig. 3.4(a). The structure under investigation consists of a dielectric slab of thickness d loaded with a surface reactance formed by the FPMS unit cell as shown in Fig. 3.4(b). Similar to the FPMS based designs previously reported, it is important to model this unit cell using its lumped circuit model. For this purpose, the starting point is taken from Jerusalem-cross based AMC structure presented in [71] (Fig. 3.4(c)). The capacitance C_{TL} represents the distributed shunt capacitance of the transmission line section of the unit cell relative to the ground plane. For the large impedance of the line, this component's contribution to the impedance can be safely ignored, consistent with [71].

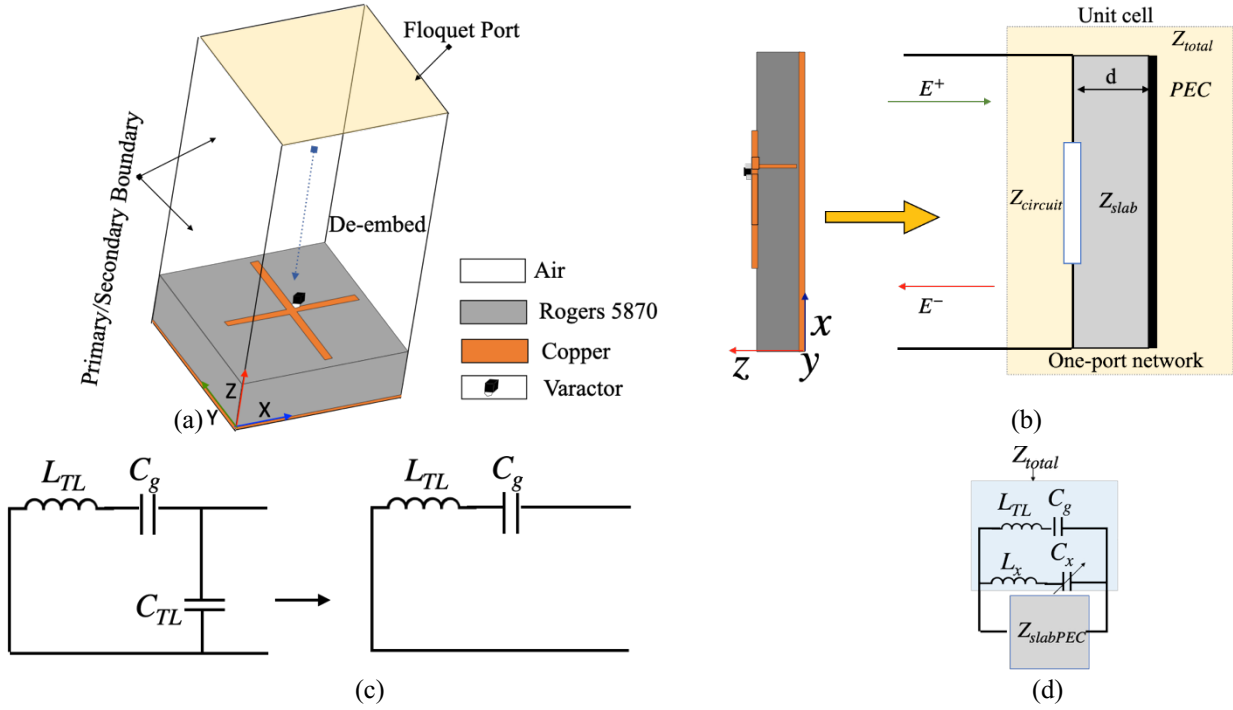


Fig. 3.4. (a) Primary/secondary boundary condition setup of AMC unit cell (b) Equivalent circuit model (side of FPMS unit cell is shown) (c) lumped circuit of transmission line (d) Transmission line model with varactor

Consequently, the equivalent circuit can be reduced to a series combination of the microstrip line inductance, L_{TL} and the gap capacitance, C_g , as shown in Fig. 3.4(c), with minimal loss of accuracy in predicting the surface impedance behavior of the structure. This circuit model is combined with the inductive impedance of the grounded substrate slab, $Z_{slabPEC}$ in parallel with the lossless lumped model of the varactor diode represented by L_x and C_x , shown Fig. 3.4(d).

The transmission line parameter L_{TL} can be evaluated using the well-known techniques [72]. Similar is the case for the impedance of the dielectric slab backed by a PEC, $Z_{slabPEC}$ [71]. These variables are evaluated using Eq. (3.2) and Eq. (3.3), respectively. The term C_g is the capacitance created due to the gap between the two adjacent unit cells and is unique when compared to the basic transmission line theory. The detailed understanding of this capacitance and its impact on the RF signal is explained by Eq. (3.4) [73] and is employed here directly. The incorporation of the

gap capacitance onto the unit cell has considerable impact on the overall impedance and cannot be overlooked. This parameter alongside the $Z_{slab_{PEC}}$ constitutes a significant difference between the proposed model and the one initially presented for the unit cells providing a new angle of explanation to the FPMS concept.

$$L_{TL} = \frac{\sqrt{\epsilon_{eff}}}{c} Z_0 \quad (3.2)$$

$$Z_{slab_{PEC}} = jZ_{slab} \tan(\beta d) \quad (3.3)$$

$$C_g = \frac{2w}{\pi} \epsilon_0 \epsilon_{eff} \cosh^{-1} \left(\frac{a}{g} \right) \quad (3.4)$$

where ϵ_{eff} is the effective dielectric constant for the microstrip line feed, c is the speed of light, β is the phase constant of the impinging wave, Z_0 is the impedance of free space and d is the thickness of the slab, a is the line length between the centers of the two-unit cell, g signifies the physical gap between the cells while w is the width of the line.

To realize tunability due to the varactor, a shunt branch composed of an inductance L_x and a variable capacitance C_x is incorporated [32]. Here, L_x corresponds to the parasitic inductance of the varactor and the metallic via connecting the one terminal of the varactor on top of substrate to the ground plane, while C_x represents the variable capacitance of the integrated varactor diode. By tuning C_x , the impedance of the unit cell can be dynamically controlled, thereby enabling reconfigurable operation. The parasitic resistance of the varactor is neglected in this model to keep the model simple and free of any ohmic losses. The impedance of the circuit of Fig. 3.4(d) excluding the slab impedance can thus be defined as:

$$Z_{circuit} = \frac{(j\omega(L_{TL} * l) + \frac{1}{j\omega C_g})(j\omega L_x + \frac{1}{j\omega C_x})}{(j\omega(L_{TL} * l) + \frac{1}{j\omega C_g}) + (j\omega L_x + \frac{1}{j\omega C_x})} \quad (3.5)$$

where l is the length of transmission line and ω is the angular frequency. Thus, the final input impedance Z_{total} of the unit cell backed by the dielectric slab is calculated by cascading in parallel the surface impedance of the circuit from Eq. (3.5) with the impedance of the slab backed by PEC Eq. (3.3):

$$Z_{total} = \frac{Z_{slab_PEC} * Z_{circuit}}{Z_{slab_PEC} + Z_{circuit}} \quad (3.6)$$

The impedance curves (Z_{total}) in Fig. 3.5 show the comparison between the theoretical and simulated surface impedances for two capacitance values of the varactor diode. Largely there is good synergy between the analytical and the full-wave simulation of HFSS with some discrepancies which are expected because of the assumptions made in the analytical model. The change in varactor's capacitance produces a controlled change in the surface impedance response both in the theory and HFSS simulations. This shift alters both the resonance location and the reactive slope of Z_{total} , indicating that the magnitude and phase response of the surface can be tailored through capacitance tuning. Such control over the surface impedance forms the basis for imposing a spatial phase variation, which is later exploited for beam steering.

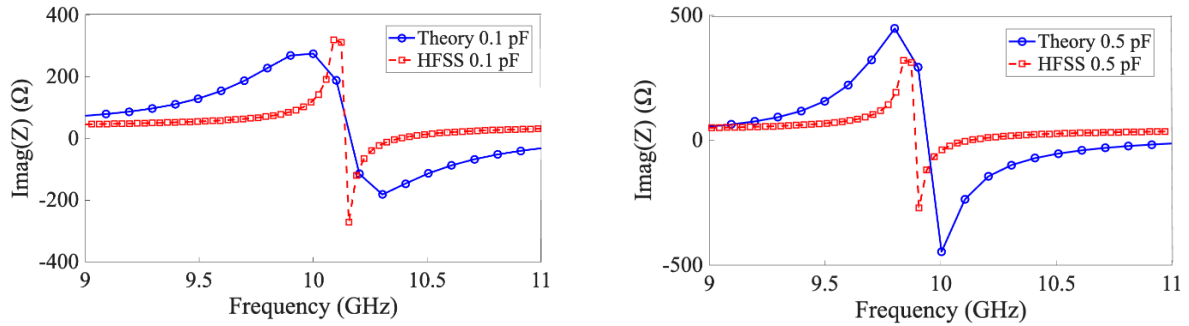


Fig. 3.5. Comparison of surface impedance (a) Reactance at 0.1 pF (b) Reactance at 0.5 pF

3.2.3 Spatial Phase Profiling of the FPMS Unit Cells

To understand the beam steering behavior of the proposed antenna integrated with the AMC based FPMS surface, it is necessary to examine how the incident near field interacts with the spatially varying surface impedance of the FPMS layer. In earlier AMC-based studies [74], the illumination was treated as a plane wave, even though the antenna was operating in the near field of the structure. In the present configuration, the radiator is placed directly above the FPMS surface; therefore, the analysis must begin from the near-field expressions of the antenna itself. This will result in a more accurate approach of modeling the antenna behavior. In [75], the field distribution inside a circular microstrip patch is derived by modeling the patch cavity as a thin resonant region bounded by perfect electric conductors (PEC) at the top and bottom surfaces and an equivalent perfect magnetic conductor (PMC) boundary along the circular rim. The PMC boundary forces the tangential magnetic field to vanish at $\rho = a$, which enforces specific modal conditions within the cavity. Pictorial image of a semi-circular patch is shown in Fig. 3.6, with the specific boundary condition on each surface.

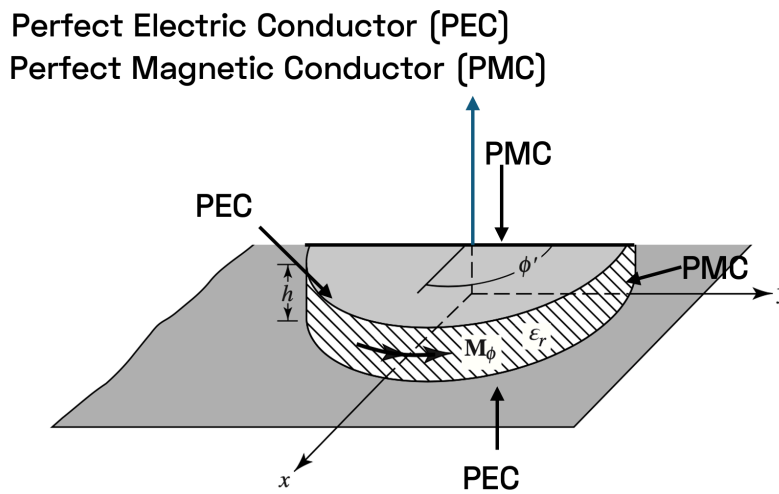


Fig. 3.6. Cavity model of semi-circular patch

For the dominant TM_{110}^z mode, the electromagnetic fields inside the cavity region exhibit purely axial electric behavior, with all transverse electric and longitudinal magnetic components being identically zero, leading to Eq. (3.7) and Eq. (3.8) for a complete circular metallic patch [74]

$$E_\rho = E_\phi = H_z = 0 \quad (3.7)$$

and the axial electric field and magnetic field

$$E_z(\rho', \phi') = E_0 J_1(k_\rho \rho') \sin \phi' \quad (3.8)$$

$$H_\rho = j \frac{E_0}{\omega \mu_0 \rho} J_1(k_\rho \rho') \cos \phi' \quad (3.9)$$

$$H_\phi = j \frac{E_0}{\omega \mu_0 \rho} J_1(k_\rho \rho') \sin \phi' \quad (3.10)$$

Here, $J_1(\cdot)$ denotes the first-order Bessel function of the first kind, k_ρ is the radial wavenumber associated with the TM_{110}^z eigenvalue, and (ρ', ϕ') describe the local cylindrical coordinates under the patch. This modal solution forms the basis for deriving the equivalent magnetic currents at the patch periphery, which subsequently govern both the near-field and far-field radiation characteristics.

For the semi-circular microstrip patch, the same cylindrical cavity formulation applies, except that the cavity is restricted to half of the angular domain and a PMC boundary is enforced along the diameter. Using cylindrical coordinates (ρ, ϕ, z) , the cavity can be defined by the region $0 \leq \rho \leq a$, $-\pi/2 \leq \phi \leq +\pi/2$, and $0 \leq z \leq h$. The metallized patch at $z = h$ and the ground plane at $z = 0$ behave as PEC boundaries, while the curved rim at $\rho = a$ and the straight diameter at $\phi = \pm\pi/2$ are modeled as PMC boundaries. As a result, the dominant TM-like mode inside the semi-circular cavity satisfies

$$E_\rho = E_\phi = H_z = 0 \quad (3.11)$$

with the axial electric field given by

$$E_z(\rho, \phi) = E_0 J_1(k_\rho \rho) \sin \phi \quad (3.12)$$

This expression represents the complete modal structure of the semi-circular cavity and serves as the starting point for constructing the equivalent aperture fields that determine the near- and far-field radiation. Using Eq. (3.12), the incident electric field expressed in Cartesian coordinates at the FPMS plane, can be written as Eq. (3.13):

$$\vec{E}_i(x, y) = E_0 J_1 \left(k_\rho \sqrt{x^2 + y^2} \right) \frac{y}{\sqrt{x^2 + y^2}} \hat{z} \quad (3.13)$$

where $\rho = \sqrt{x^2 + y^2}$ and $\sin \phi = y/\rho$

Above equation determines the amplitude and phase of the incident field at each point on the FPMS plane. The reflected electric field at the FPMS plane is therefore expressed as Eq. (3.14) which is defined in [74] and is used as the basis for the derivation of the proposed theory.

$$\vec{E}_r(x, y) = \Gamma(x, y) \vec{E}_i(x, y) \quad (3.14)$$

where the local reflection coefficient is given by Eq. (3.15):

$$\Gamma(x, y) = \frac{Z_{\text{total}}(x, y) - Z_0}{Z_{\text{total}}(x, y) + Z_0} \quad (3.15)$$

Here, $Z_{\text{total}}(x, y)$ is the impedance presented by the FPMS layer and depends on the distributed parameters L_{TL}, C_g , the varactor diode lumped circuit elements and the grounded-slab impedance Z_{slabPEC} (given by Eq. (3.3). It is worthy of mentioning it here that the value of $Z_{\text{total}}(x, y)$ is extracted from HFSS simulation of Fig. 3.5 for different values of capacitance as it presents a more

accurate value for the modeling purposes. For a lossless structure, $Z_{\text{total}}(x, y)$ is dominated by its reactive part,

$$Z_{\text{total}}(x, y) = jX_{\text{total}}(x, y) \quad (3.16)$$

Under this condition, the magnitude of $\Gamma(x, y)$ is theoretically unity, and its phase reduces to Eq. (3.17):

$$\phi_r(x, y) = \arg\{\Gamma(x, y)\} = -2 \tan^{-1} \left(\frac{X_{\text{total}}(x, y)}{Z_0} \right) \quad (3.17)$$

The reflected field can therefore be written as Eq. (3.18):

$$\vec{E}_r(x, y) = \vec{E}_i(x, y) \exp \left[-j 2 \tan^{-1} \left(\frac{X_{\text{total}}(x, y)}{Z_0} \right) \right] \quad (3.18)$$

When the FPMS surface is discretized into unit cells, the incident field interacts independently with each cell. Evaluated at the center (x_m, y_n) of the $(m, n)^{\text{th}}$ unit cell, the assigned reflection phase becomes Eq. (3.19):

$$\phi_r(x_m, y_n) = -2 \tan^{-1} \left(\frac{X_{\text{total}}(x_m, y_n)}{Z_0} \right) \quad (3.19)$$

For this analysis, the reflected field at the FPMS plane is identified with the discrete aperture excitation which is added back into the incident field resulting in the total radiated field of the AMC-backed antenna as:

$$\begin{aligned} E_{\text{ap}}(x_m, y_n) &= \vec{E}_r(x_m, y_n) + \vec{E}_i(x_m, y_n) \\ &= E_i(x_m, y_n) \exp \left[-j 2 \tan^{-1} \left(\frac{X_{\text{total}}(x, y)}{Z_0} \right) \right] \\ &\quad + E_0 J_1 \left(k_\rho \sqrt{x^2 + y^2} \right) \frac{y}{\sqrt{x^2 + y^2}} \hat{z} \end{aligned} \quad (3.20)$$

The reflected portion of the field in Eq. (3.20) represents the synthesis of all locally reflected contributions, and the imposed spatial variation of $X_{\text{total}}(x_m, y_n)$ directly sets the reflection phase distribution across the FPMS. The resulting phase gradient governs the direction of constructive

interference in the radiated field, thereby enabling controlled beam steering. Finally, the far-field radiation pattern corresponding to this aperture excitation follows directly derived from the standard aperture-radiation theory and is given by Eq. (3.21):

$$E_{\infty}(\theta, \phi) \propto \sum_{m=1}^M \sum_{n=1}^N E_{ap}(x_m, y_n) \exp\{jk_0[x_m \sin \theta \cos \phi + y_n \sin \theta \sin \phi]\} \quad (3.21)$$

This representation highlights that the FPMS aperture behaves as a programmable reflecting surface whose far-field characteristics are directly determined by the spatial distribution of the reflected component of $E_{ap}(x_m, y_n)$. The incident component is largely constant as it represents the radiated field of a simple patch antenna for this concept. To validate this, a comparison between the theoretical and simulated beam steering is included in the subsequent section. This discussion is left for a later point in the manuscript as it is useful to validate the theoretical model by placing it right next to the simulated results.

3.3 Antenna Integration with FPMS

As discussed previously, a simple semi-circular patch antenna is envisioned to be used with the reconfigurable FPMS based AMC layer. Keeping the antenna simple, eases the conclusions that can be drawn from the design concept. The microstrip fed semi-circle shaped patch with its top is shown in Fig. 3.7(a). At first, to realize the omnidirectional pattern, the ground plane is only extended under the antenna feed. This results in a peak gain of 2.17 dBi, which is typical for this class of antenna (monopole). The reflection coefficient remains below -10 dB throughout the X-band frequency range and beyond; however, this large bandwidth is not the main focus here and is expected to significantly diminish due to the AMC integration. The simulated results of the antenna are quite commonplace and, therefore, omitted from inclusion in this section. After studying the

antenna and the FPMS unit cells individually, the next logical step is to integrate them in a vertical fashion to realize the proposed FPMS-AMC based antenna element, as shown in Fig. 3.7(a). The integration is to be done carefully while keeping in mind the antenna structure and the middle FPMS layer. A total of 25-unit cells is used between the antenna and the ground plane. Each unit cell is kept isolated from the others to provide individual control within the region.

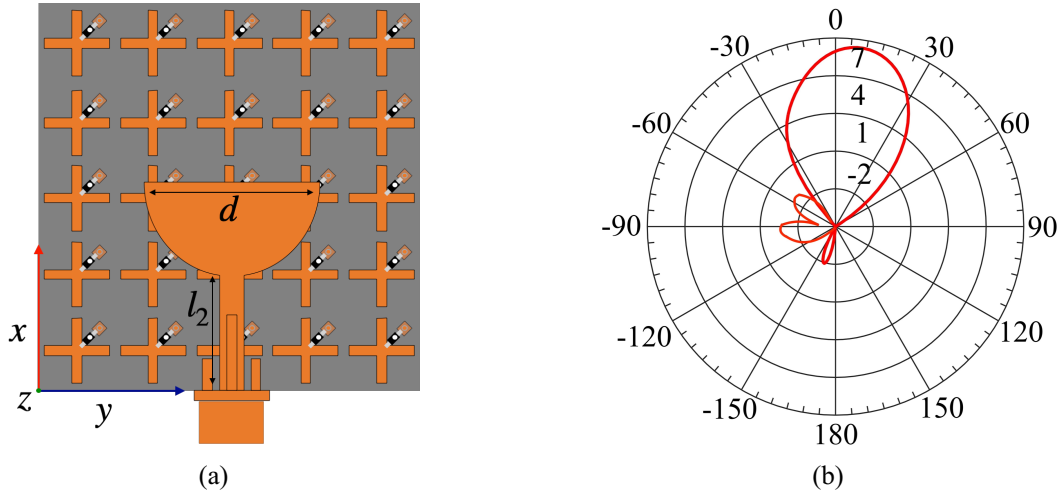
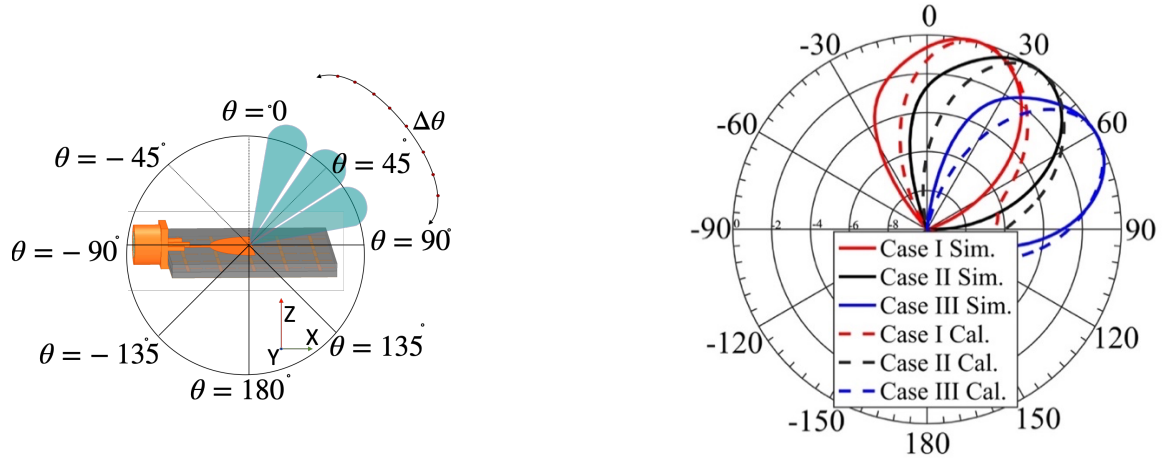


Fig. 3.7. (a) Top view of antenna ($l_2 = 13.08$ mm, $d = 26$ mm, $h_1 = 1.575$ mm) (b) Radiation pattern of patch antenna ($\phi = 0^\circ$) (all FPMS unit cells have 0.15 pF capacitance)

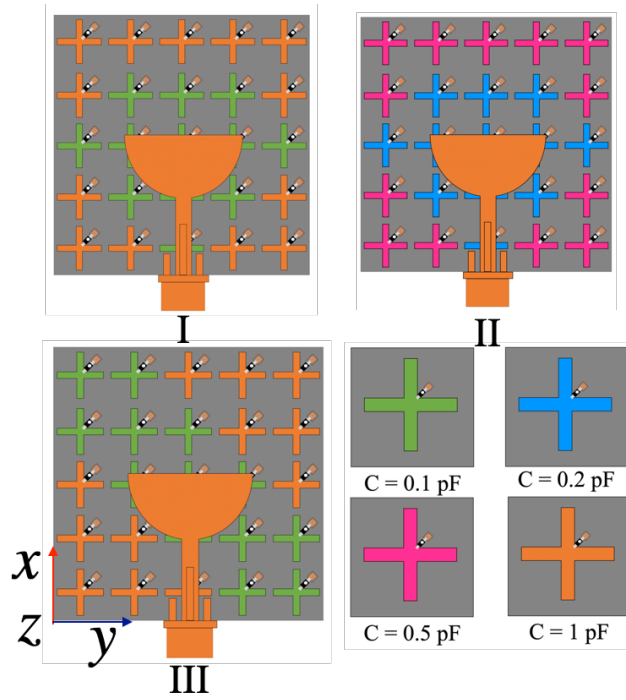
This strategy would complicate the biasing of the unit cells but provides better control to the designer. As expected, carefully stacking the antenna on the FPMS board results in an increased gain, due to the AMC's reflective behavior. This has been achieved by optimizing the gap between the antenna and the FPMS layer. The antenna's gain increases to almost 8.2 dBi with this modification, which demonstrates an increase of ~ 6 dB as compared to the standalone antenna gain at 10 GHz. The reflection coefficient magnitude of the FPMS unit cell for 0.15 pF capacitance is $\sim 70\%$. This magnitude is not optimized further due to the inhomogeneity of the FPMS layer with varying capacitance values which is different from classical AMC based designs. The radiation efficiency of the antenna reduces to 90 % from 95% due to the addition of the unit cells which is quite acceptable. The gain value may well be increased further by adding more unit cells,

but that comes at the cost of increased complexity. Therefore, at this stage, the authors deem this no. of unit cells to be sufficient. Furthermore, the fact that these unit cells are to be biased, supports the idea of keeping the number limited at this point. Initially, the capacitance value assigned to the unit cells is 0.15 pF and the radiation pattern in XZ-plane thus obtained is shown in Fig. 3.7(b). These results perfectly align with the traditional AMC-based antenna designs where reasonable gain enhancement is achieved due to the conversion of an omnidirectional pattern to a directive one with optimal beamwidth.

After achieving the intended gain enhancement while using a single capacitor value for all the unit cells, it is pertinent to study the antenna performance with varying degrees of biasing for different unit cells. There are numerous combinations that can be applied to bias the FPMS-AMC unit cells. To validate continuous beam steering capability, multiple simulations are conducted, offering a more comprehensive assessment of the antenna's reconfigurable performance. For conciseness, a subset of results is presented, showcasing the antenna's remarkable ability to continuously steer its beam within a range of $+15^\circ$ to $+60^\circ$ in the XZ-plane (elevation) while maintaining a gain of around 8 dBi. The radiation patterns for states I to III are shown in Fig. 3.8(a) with the capacitance values elaborated in Fig. 3.8(b). This beam steering is achieved through a specific configuration of FPMS unit cells, where the capacitance values of the unit cells are dynamically adjusted to manipulate the phase distribution across the surface. A point of interest that should be kept in mind is that using a mix of capacitance values through the FPMS layer allows for the control on the dielectric constant experienced by the antenna. The resulting effect is the steering in the elevation plane achieved from the radiating structure.



(a)



(b)

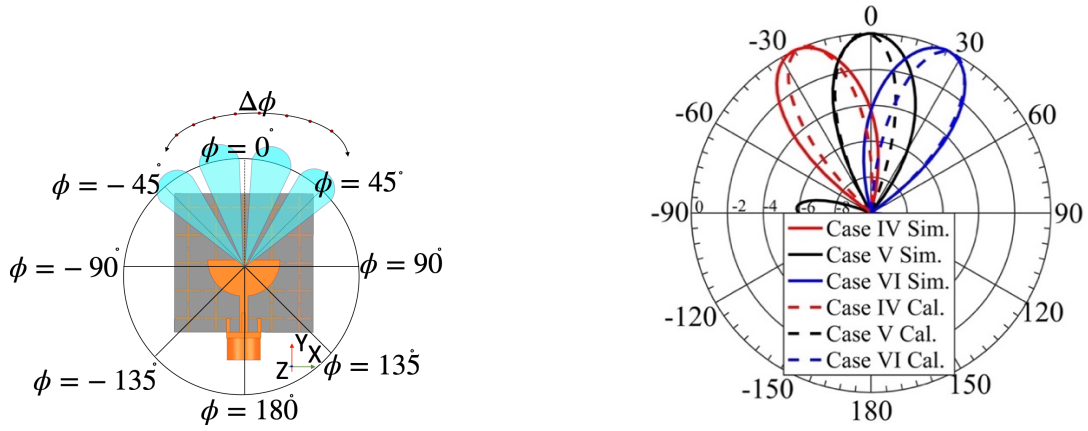
Fig 3.8. FPMS unit cells configurations are represented by different color codes (Substrate of antenna is hidden for better illustration) (a) Beam steering with normalized gain in the elevation plane ($\phi = 0^\circ$) (b) FPMS configuration states for elevation beam steering

To further strengthen this observation, the theoretically calculated beam steering angles are also evaluated and compared with the simulation results. For this purpose, (3.21) is evaluated using the reflection phase of (3.19) which depends on the varying capacitance of the varactor. It should be

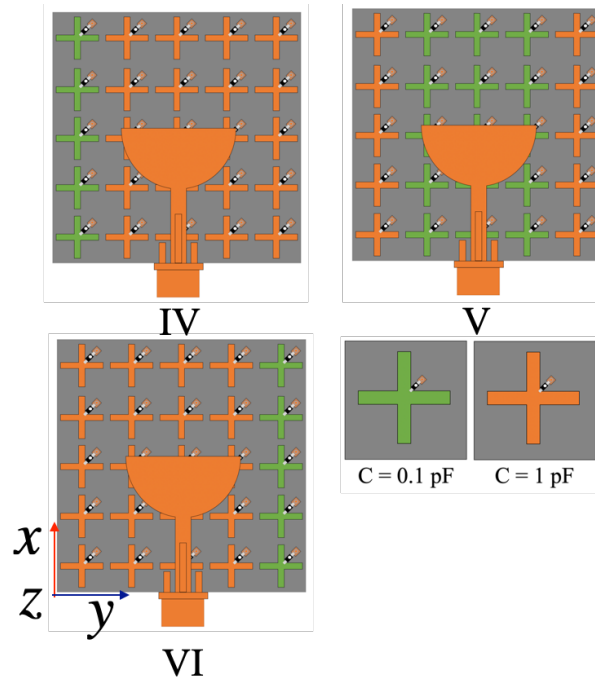
mentioned here that for the analytical modeling the reflection phase of (3.21) is evaluated individually for each cell and then superimposed on the incident field to get the reflected field of each cell, (3.18). Finally, (3.21) is used to combine the effect of 25-unit cells (5×5 FPMS layer), resulting in the curves shown in Fig. 3.8(a). The predicted steering directions closely follow the simulated beam positions for the same biasing states, confirming that the proposed theoretical approach captures the underlying steering mechanism. This agreement indicates that the spatial phase variation introduced by the biased FPMS-AMC surface is sufficient to explain the change in the main radiation direction. Some discrepancy can be observed in the beamwidth of the two sets of results. This deviation primarily arises from a simplifying assumption made in the theoretical analysis, where all unit cells are considered to be excited by an identical incident field. In practice, the antenna operates in the near field of the FPMS surface, and therefore the excitation is inherently non-uniform. Furthermore, the unit cell is considered to be completely lossless in this design which is not true for the simulation environment. All-in-all the results of Fig. 3.8(a), provide a reasonable validation of the proposed theory.

After observing the elevation scanning, it is intriguing to see, if the antenna beam can be revolved around the azimuthal plane. It is already established that the maximum of the radiation can be steered to $\theta = 60^\circ$. Using this point in the elevation, the study is now extended to evaluate the beam control in the azimuthal plane (XY-plane). By employing the states of Fig. 3.9(b), the beam is steered from $\phi = -25^\circ$ to $+25^\circ$, shown in Fig. 3.9(b). The reconfigurability in this plane is also realized by tailoring the FPMS unit cell configurations to alter the substrate properties locally and achieve the desired radiation direction. In the case of azimuthal steering, the two capacitance values used in the unit cells are 0.1 pF and 1 pF. It is evident that beam reconfiguration can be realized not only through different configurations of FPMS unit cells but also by varying the

capacitance values within the same structure, allowing effective tuning of the antenna's radiation characteristics.



(a)



(b)

Fig 3.9. FPMS unit cells configurations are represented by different color codes (Substrate of antenna is hidden for better illustration) (a) Beam steering with normalized gain in the azimuthal plane i.e. ($\theta = 60^\circ$) (b) FPMS configuration states for azimuthal beam steering

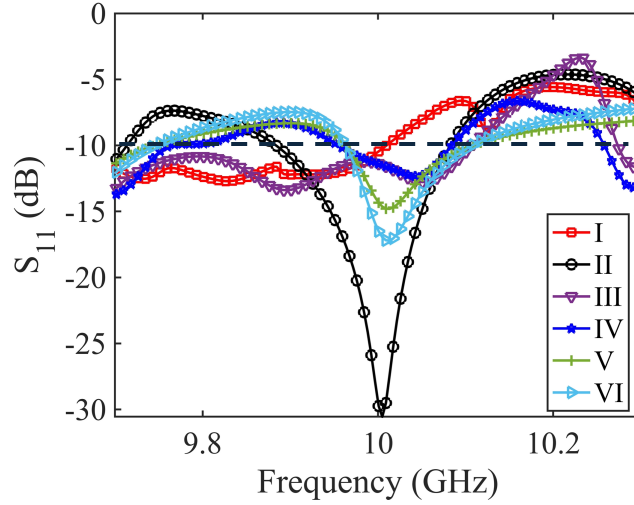


Fig. 3.10. Antenna reflection coefficient for all cases

Just like for the elevation plane, the direction of beams for the case of azimuthal steering are also calculated using the theoretical model. Again, a good agreement can be seen between the analytical and simulated antenna radiation in Fig. 3.9(a). This reinforces that the local modification of the substrate properties through the FPMS unit cell configuration can also govern the azimuthal beam control. Notably, for both the steering scenarios, i.e., elevation and azimuthal, the reflection coefficients remain well-matched below -10 dB, ensuring efficient impedance matching and minimal signal loss (Fig. 3.10).

Acceptably, as the beam is steered, some reduction in the maximum antenna gain is anticipated. The variation observed here is less than 1 dB, validating the impedance to be close to the matched condition. The trend of the antenna reflection coefficient trace is not quite as smooth as usually seen in antenna designs. This is due to the presence of metallic unit cells underneath the feed line of the antenna. However, since minimum matching conditions are met, these results can be considered acceptable for proceeding to proof-of-concept. In the future, it would be prudent to avoid the use of metallic cells underneath a section of the microstrip feed. This can help in improving the antenna's impedance.

3.4 Biasing Strategy for Antenna Prototype

In the RF modeling of the FPMS based antenna, the varactor diode is modeled as a simple capacitor with varying values alongside other lumped elements of Fig. 3.1(c). Although the actual antenna prototype needs to be provided with DC biasing to control the capacitance value. Therefore, it is crucial to understand the biasing mechanism, as it plays a fundamental role in achieving reconfigurability. The 3D isometric view of the final antenna structure is represented in Fig. 3.11(a). The first two layers are the same as the ones shown in Fig. 3.1; however, an additional layer is added underneath the antenna ground plane, referred to as layer 3. This layer is provided with the DC bias lines for the varactor diode on the top (Trace 1) as well as the bottom surface (Trace 2), Fig. 3.11(b) and (c). SPDT switches are integrated at the bottom of layer 3 to toggle between the two sets of DC bias lines (Traces 1 and 2). Interestingly, two sets of bias lines are used because at any given point, the unit cells are biased with only two distinct voltages, as explained in the last section. Thus, if the bottom set of lines is applied with one voltage value, the top can be assigned a different one. With the assistance of toggling the switch, the designer can choose the bias value for an individual unit cell. The varactor diode is to be mounted in a fashion that the cathode terminal is directly connected to the center of the unit cell, which is exposed to the DC biasing layer with the help of via A. The anode of the diode is shorted to the RF ground, which acts as a dual ground for both the RF and DC sources, in Fig. 3.11(d). The DC voltage is applied through via A, which serves as the positive biasing point. To join via A to the DC biasing lines, the authors have used headers, 732-5315-ND and socket 1212-1180-ND to act as an external connector in tandem with the internal vias (via A), shown in Fig. 3.11(e). The voltage on the top biasing lines of layer 3 is referred to as V1, while the value on the bottom lines is defined as V2.

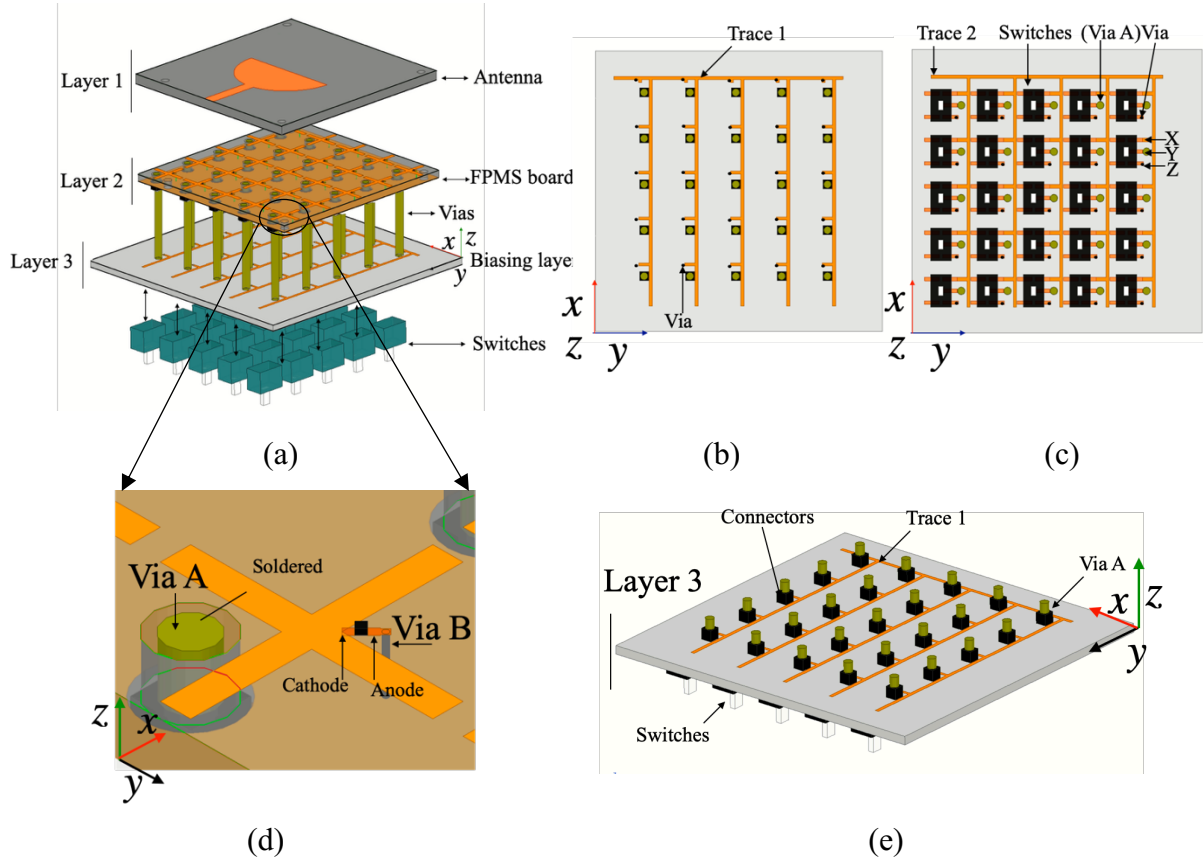


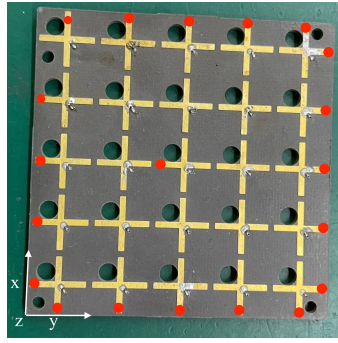
Fig 3.11. (a) Exploded view of antenna structure, showing Layer 1, Layer 2 and Layer 3 separately. Layer 1 is Antenna, Layer 2 is FPMS board and Layer 3 is Biasing layer (b) Top view of Layer 3 (c) Bottom view of Layer 3 (d) close view of varactor placed on Layer 2 (e) Prospective view of biasing layer

The values of V_1 and V_2 would be changing as needed; however, they are given fixed attribution for simplicity. The mounting strategy for the SPDT switch is also critical for the planned biasing scheme. The center terminal labelled as 'Y' in Fig. 3.11(c), is responsible for supplying the DC voltage to the FPMS unit cells, while the terminals 'X' and 'Z' choose the value between the two voltage levels. Trace 1 and 2 of the biasing lines are connected to the terminals Z and X, respectively. Since Trace 2 is on a different level as the SPDT switch, a metallic via is used to connect the two together (visible on terminal Z line). When the switch is moved towards X, it

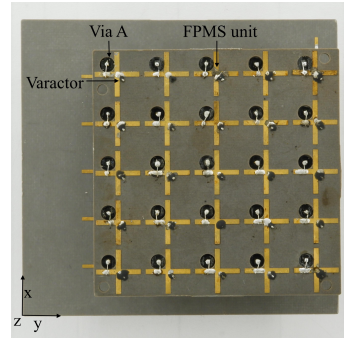
connects the unit cell to voltage V_1 , and likewise can be interpreted for the other scenario, i.e., Z . In this way, the designer can control the bias value applied to each unit cell with complete freedom.

3.5 Fabrication and Measurements

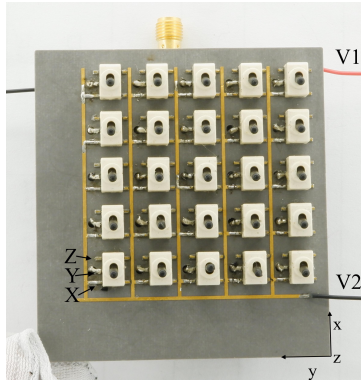
The realization of the proposed antenna system involved several practical challenges during fabrication and characterization. These included the precise mounting of varactor diodes, alignment and stacking of the substrate layers, and integration of RF connectors. Among these, the placement of varactor diodes presented the most critical fabrication difficulty, as the components were required to be embedded between the upper substrate layers with limited accessibility. MACOM provides these diodes in a beam-lead package, as specified in the datasheet, which highlights the need for careful handling and precise mounting procedures. Eventually, the optimum solution devised in this case is the use of dummy SMT resistors on the FPMS-AMC unit cell layer. The highlighted red dots, shown in Fig. 3.12(a), are the points where the resistors are mounted. These components are smartly placed onto the metallic strip so as to avoid any modifications in the antenna response due to their presence (Fig. 3.12(b)). Obviously, this created a small gap between layers 1 and 2 of the prototype that has been since modeled in the full-wave simulator to study its effect. Fortunately, no major change is seen due to this change, as is validated by the simulations. For the stacking of top two layers, four holes are drilled on the substrate edges that are inserted with ribbits to keep the substrates in their place. After which, the tapes are clamped on the four edges to provide further mechanical support. The SMA connector is then mounted onto the antenna feed while connecting it to the ground plane using conductive epoxy. These steps prepare the top two layers, consisting of the antenna, the unit cells, and the ground plane.



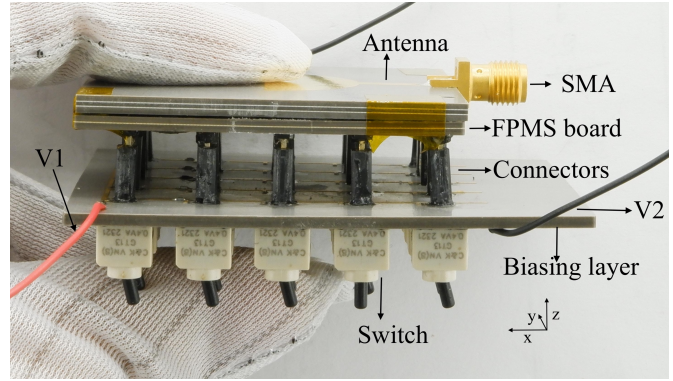
(a)



(b)



(c)



(d)

Fig 3.12 (a)Top view of FPMS board (Layer 2) in on top of layer 3 (red dots represent dummy resistors for protection) (b) Top view of whole prototype (c) side view of whole prototype (d) bottom view of layer 3

The bottom-most layer is provided with the DC bias lines as explained in the last section. SPDT switches mounted on the bottom of this layer to toggle between two different voltage levels are photographed in Fig.3.12(c). Finally, vertical connectors are used to attach the antenna layers/substrates to the DC bias layer as represented in Fig. 3.12(d). The side view of the final prototype is visible here. The prototype antenna is measured for its impedance and radiation characteristics. The measured reflection coefficient of the antenna for a variety of biasing states is shown in Fig. 3.13. For the reader's convenience, 6 different cases are added to the plot. In all these cases, the antenna is well-matched at 10 GHz, showing minimum reflection at the input.

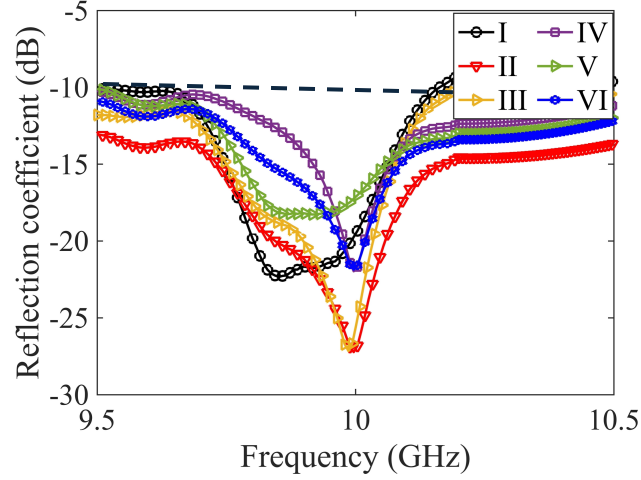


Fig. 3.13. Reflection coefficient of antenna for the cases I – VI

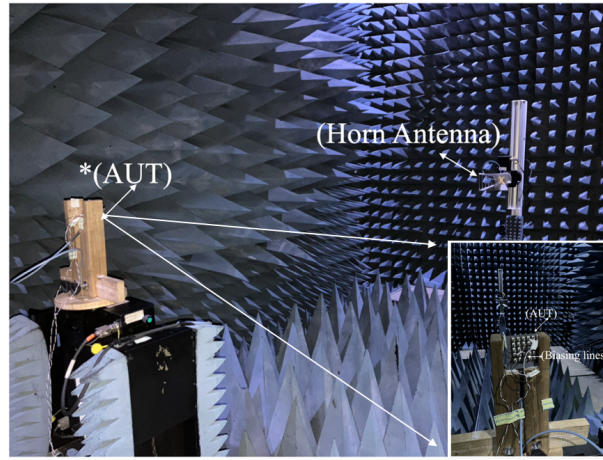


Fig. 3.14. Anechoic chamber with the AUT

This signifies that the simulation model employed for this design is robust and establishes confidence for the next steps of measurements. For the gain and radiation measurement of the prototype, it is mounted in an anechoic chamber, Fig. 3.14. The simulations of this antenna show that it can be steered in the elevation plane from almost the bore-sight radiation to 60° . This steering is achieved in the XZ plane ($\phi = 0^\circ$) and is reasonably straightforward to measure. The anechoic chamber facility available to the authors can only measure in a single plane at a time. Therefore, setting the antenna to be measured in $\phi = 0^\circ$ plane is simple. The steering for this case is drawn in Fig. 3.15(a) with a reasonable match between the simulated and measured data.

However, the challenge is to measure the antenna in the azimuthal plane while tilting the maximum of the antenna to 60° . This required appropriate antenna alignment in the chamber. Nevertheless, eventually the authors are able to set the antenna in the correct alignment and bias it in the desired configurations. The steering in the azimuthal plane is shown in Fig. 3.15(b), again validating the simulation model. A reasonable match is observed in Fig. 3.15 between the simulated and measured radiation pattern of the antenna.

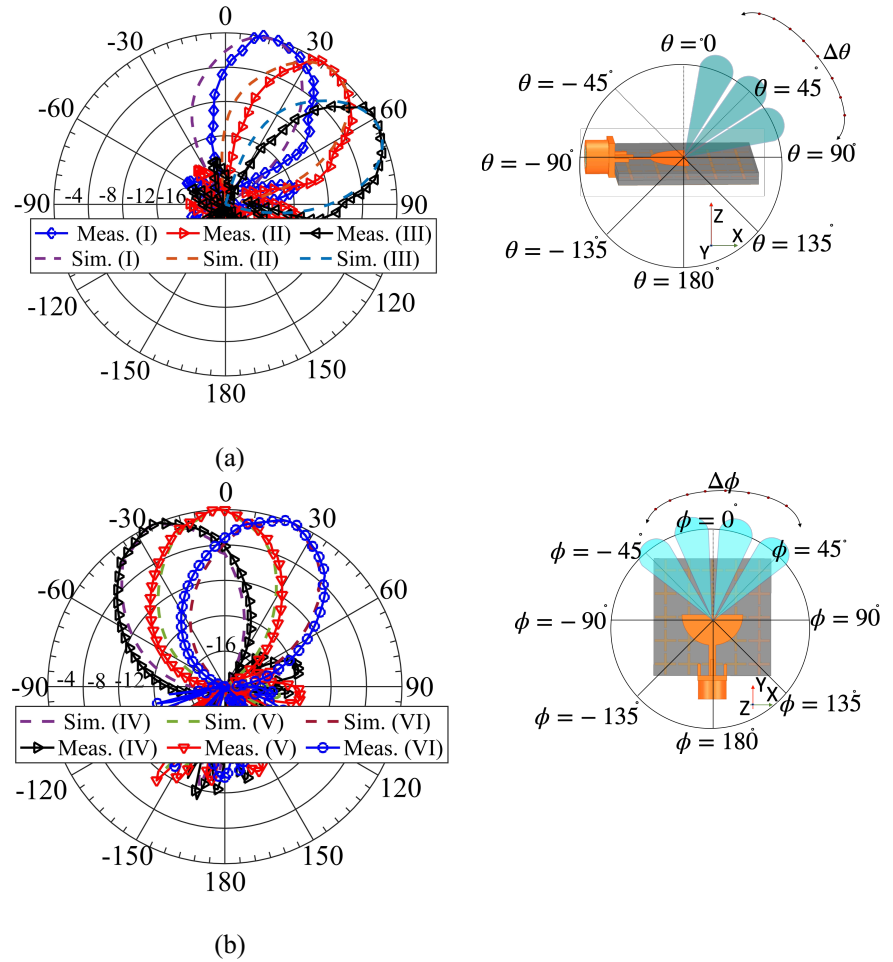


Fig. 3.15. Simulated vs. measured radiation pattern in the (a) elevation plane (b) azimuthal plane ($\theta = 60^\circ$)

The maximum measured gain of the antenna comes out to be 7.2 dBi with some variation in its magnitude while steering the beam. These values compare well with the simulations. A 1 dB difference between the simulated and measured gain values can be attributed to the practical

aspects such as connector mounting, substrate stack up, loss coming from varactor diodes, etc. In any case, this discrepancy is within the acceptable range. Finally, to demonstrate continuous beam steering, the authors measured the antenna in finer angular steps both in the elevation and the azimuthal planes. The biasing configurations reported alongside the radiation plots of Fig. 3.16, where (a) shows the elevation plane, while (b) is used for the azimuthal plane.

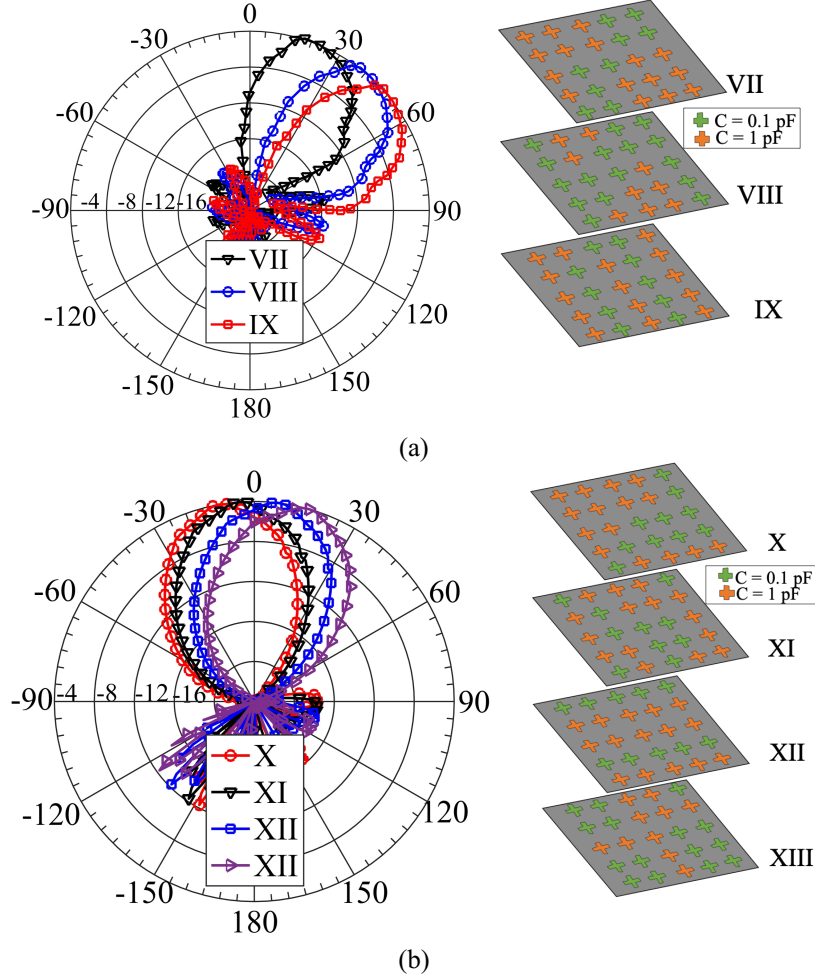


Fig. 3.16. (a) Continuous beam steering in XZ-plane ($\phi = 0^\circ$) (b) Continuous beam steering in XY-plane ($\theta = 60^\circ$)

Table I shows the summary of the steering achieved from a variety of these states, elaborating on how to bias the unit cells as per the steering requirements. This summary acts as a look-up sheet for the users relying on this design to achieve the desired goals.

Table I. Continuous beam steering for different cases

Elevation ($\phi = 0^\circ$)	Cases for ($\phi = 0^\circ$)	Azimuthal ($\theta = 60^\circ$)	Cases ($\theta = 60^\circ$)
+15°	I	-25°	IV
+20°	VII	-10°	X
+32°	II	0°	V
+35°	VIII	+10°	XII
+42°	IX	+15°	XIII
+60°	III	+25°	VI

Table II. Comparison with State of the Art

(*S.E = Single Element, *M.E = Multiple Elements, *Azi = Azimuthal plane, *Elev = Elevation plane)

Ref.	Aperture ($W/\lambda_0 \times H/\lambda_0$)	Scan Range (Azi/Elev)	Gain Reduction	Control Mech.	Operational Bandwidth	Aperture Efficiency	S.E or M.E
[76]	1.05 x 1.05	+20°	-	Varactor diodes	3.3-3.6 GHz	>~15%	S. E
[77]	0.65x0.65	-20°,0°,20°/ -40°,0°,40°	~1 dB	PIN diodes + feeding network	~16% -30%	38%	S.E
[78]	~1.7x1.36	+23	<- 0.96 dB	PIN diodes	4.2-4.55 (~8%)	11.5%	S. E
[79]	4.2 x 0.96	-/ 72°	0.9 dB	Phase shifter	10.04- 10.58GHz (~5.2%)	-	M.E
[80]	2.4 x 2.4	$\pm 30^\circ/\pm 60^\circ$	5 dB	Phase shifters	X-band: ~46% L-band: ~21%	-	M.E
[81]	4.6 x 4.6	+45°	2 dB	Frequency Dependent	X-band: ~16.35% Ka— band:18%	64 % for X-band	M.E
[82]	5.4 x 2.7	$\pm 50^\circ/\pm 50^\circ$	~5dB	Phase shifter	S- band:11.7% X-band: 11.1%	-	M.E
This work	1.67 x 1.67	$\pm 25^\circ / 45^\circ$	~1 dB	varactors	X-band: 7.2%	~19 %	S. E

3.6 Comparison with the State of the Art

A comparison of this work with other beam steering antennas is summarized in Table II. It can be observed that a few single-element designs, such as those reported in [76]–[78], achieve beam steering using metasurfaces in combination with varactors or PIN diodes, demonstrating comparable gain stability and compact implementation. However, these approaches are generally limited in one or more aspects, such as reduced or discrete steering range and the requirement of additional feeding structures for excitation and control. In particular, some designs achieve steering only over a limited angular range (discretely) or along a single plane, while others rely on biasing schemes that increase implementation complexity. In comparison, the proposed antenna operates in the X-band and enables continuous two-dimensional beam steering ($\pm 25^\circ / 45^\circ$) using 25 varactor elements, without requiring a complex feeding network. Although the achieved gain variation (~ 1 dB) is comparable to existing single-element designs, the proposed approach provides enhanced flexibility in beam control through continuous and simultaneous steering in both azimuth and elevation planes, while maintaining a compact aperture ($1.67 \times 1.67 \lambda_0$) and low direct-current power consumption (< 3 mW). Other beam steering antenna designs rely on phased array configurations [79]–[82], which inherently involve multiple radiating elements, phase shifters, and distribution networks. While such systems can achieve high directivity and wide scanning ranges, these designs typically result in increased system complexity, larger physical size, and higher power consumption. In contrast, the proposed antenna offers a simplified architecture by shifting the complexity from the RF domain to low-power control elements, thereby providing a practical trade-off between performance and implementation complexity. Overall, the proposed antenna demonstrates a balanced performance in terms of steering capability, gain stability, and

structural simplicity, making it a competitive alternative to both single-element antenna using metasurface and phased-array-based beam steering solutions.

3.7 Conclusion

The study demonstrates the innovative use of FPMS inspired AMC to enable dynamic beam steering and gain enhancement in a simple antenna structure. By integrating a patch antenna with the FPMS unit cells, continuous beam steering is achieved in both the elevation and azimuthal planes, with angular ranges of 15° to 60° and -25° to 25° , respectively. The versatility of the FPMS is evident, as varying configurations and capacitance values effectively reconfigure the antenna's radiation characteristics. Simulations confirm consistent performance, including a gain improvement of ~ 6 dB and reflection coefficients maintained below -10 dB across all configurations. A novel theoretical model has also been proposed to explain the antenna steering due to the presence of FPMS unit cells. The analytical results are validated by simulations as well as with the help of an actual prototype. Therefore, it can be safely ascertained that a good foundation has been laid down by this work for exploring more designs that can build on this proof-of-concept. At present the design explores the use of a 5×5 array of FPMS unit cells to explore this concept. By increasing the number of unit cell, one may achieve better beam steering range from the antenna. However, this adds to the complexity of the DC biasing strategy. Thus, a trade-off exists between increasing the complexity and enhancing the beam steering performance of the antenna using this approach. The designer can decide on a compromise on the basis of the application at hand and its requirements.

Chapter 4

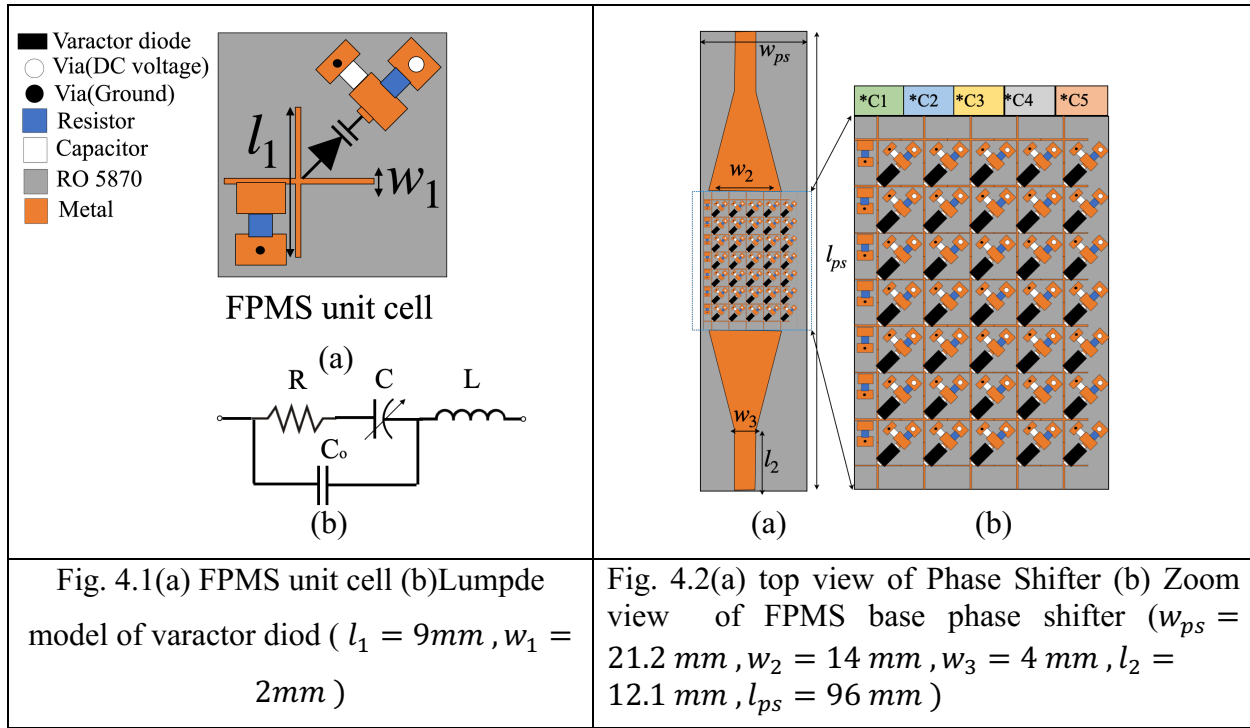
FPMS Based Agile Phase Shifter Design

Phase shifters are essential components in modern microwave and millimeter-wave systems because they provide precise control over the phase of propagating EM waves without significantly affecting their amplitude. Their ability to introduce a controllable phase delay makes them indispensable in phased-array antennas, beamforming networks, radar systems, satellite communications, and next-generation wireless technologies. Among these applications, phased-array antennas represent one of the most important uses of phase shifters, where a progressive phase difference between adjacent radiating elements enables electronic beam steering without mechanical movement. Consequently, the performance of a phase shifter is primarily determined by its achievable phase range, insertion loss, impedance matching, and implementation complexity. This chapter presents the design, analytical modeling, and implementation of a reconfigurable phase shifter based on the FPMS. Furthermore, the proposed phase shifter is integrated with a printed dipole antenna and subsequently extended to a two-element antenna array to demonstrate electronic beam steering through capacitance-controlled phase progression.

4.1 Design Concept of FPMS Phase Shifter

The phase shifter proposed herewith is composed of multiple FPMS unit cells providing a path for the wave to flow but with an added control with the designer. The structure is depicted in Fig. 4.1 (a) with the active part of the design illustrated in Fig. 4.1 (b). The targeted frequency is opted to be 2.5 GHz as opposed to 10 GHz. The rationale for this change is the cost of the varactor diodes that are available at 10 GHz. To reduce the cost and still provide a proof-of-concept, the design is

operated within S-band. The active part consists of cascaded FPMS unit cells that connect the input and output ports of the design. Since each unit cell can be independently controlled through the applied bias, the overall phase response of the structure can be precisely tailored. However, for the purpose of this design, the author plans to provide a single bias to one column, thus maintaining the properties along a line. This can be discussed in detail at a later stage. Like the antenna in the last chapter, this design is also implemented on a Rogers 5870 substrate with a thickness of 1.575 mm. A total of 35-unit cells arranged in a 5×7 grid make up the active part of the phase shifter corresponding to 5 columns and 7 rows.



To emulate the waveguide-like behavior, it is important to re-visit the unit cell first as the manner in which the concept is used in the phase shifter is quite different from the AMC based antenna. The geometry of a single unit cell is shown in Fig. 4.1(b), consisting of two microstrip lines arranged orthogonally to form a cross-patch structure, consistent with what has been previously reported in this thesis. There is one major difference in this case as compared to the

AMC design. Here, the unit cell is to be directly excited with the DC bias and RF signal at the same time. The meaning of which is that both types of signals are physically excited on the same metal layer. Therefore, the strategy of bias is similar to the works reported before [31],[32]. To avoid repetition, the design details are skipped here and only the component values used in the unit cell are shown in the figure.

4.2 Effective Permittivity Control Through Varactor Tuning

The reconfigurable behavior of the proposed phase shifter is based on the programmable unit cell that provides direct control over the effective permittivity of the substrate by modulating the impedance of the propagating wave. These details have been covered in Chapter 2 of the thesis. By simulating the unit cell in a full-wave simulation set-up and employing NRW algorithm, once can extract the substrate properties experienced by the wave. For this design, when the structure is simulated through a capacitor range of 0.1 pF to 1 pF, the resulting dielectric constant values are shown in Fig. 4.3.

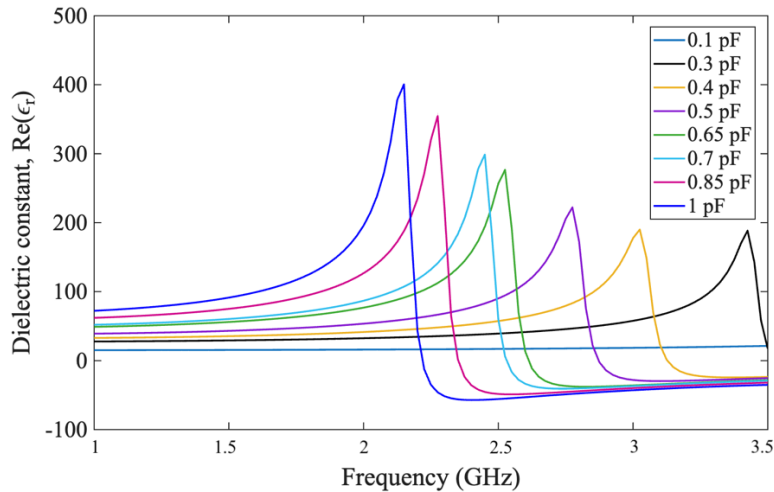


Fig. 4.3 Effective permittivity of unit cell

It is evident that around 2.5 GHz the ϵ_r values go through a change from -75 to 110. This data is quite useful in planning the phase shifter design as it can help in deciding the bias values to be used. As expected, the unit cell resonates at different frequencies depending on the assigned capacitance value. Beyond this resonance, the impedance of the structure switches from capacitive to inductive thus resulting in negative ϵ_r values. These results are useful to conclude that between approximately 0.1 pF and 0.6 pF the cell exhibits positive effective permittivity, whereas from 0.8 pF and 1.0 pF the same structure provides negative effective permittivity to the incoming wave. Therefore, the same programmable unit cell can realize two distinct EM functions depending on the applied bias condition. On one hand, it can form a dielectric region with continuously tunable permittivity. On the other hand, it can emulate a negative permittivity medium thus causing the wave to be evanescent within this region. This dual functionality provides the physical mechanism on which the phase shifter design is built.

4.3 Operation of Reconfigurable Phase Shifter

The operation of the proposed phase shifter can be understood by drawing an analogy with a SIW. An SIW consists of a dielectric-filled waveguide bounded by two rows of metallic vias that emulate the sidewalls of a conventional rectangular waveguide. EM waves propagate within the dielectric region as guided modes, and the accumulated phase depends on the propagation constant of the supported mode. Since the proposed structure also forms a dielectric-guided propagation channel with synthesized electromagnetic boundaries, it follows the same guided-wave propagation mechanism. Therefore, the phase delay can be analyzed using the propagation characteristics of a dielectric waveguide. For a dielectric-filled waveguide, the phase accumulated over a propagation length L is given by Eq. (4.1):

$$\phi = \beta L \quad (4.1)$$

where β is the propagation constant of the dominant guided mode. For the dominant mode, the propagation constant is expressed as Eq. (4.2):

$$\beta = \sqrt{k_0^2 \epsilon_{\text{eff}} - k_c^2} \quad (4.2)$$

where $k_0 = 2\pi/\lambda_0$ is the free-space wavenumber, ϵ_{eff} is the effective permittivity of the dielectric guiding region, and k_c is the cutoff wavenumber determined by the waveguide geometry. This relationship indicates that wave propagation depends not only on the dielectric properties of the guiding medium but also on the confinement introduced by the waveguide boundaries. Since the waveguide geometry remains fixed after fabrication, the cutoff wavenumber remains constant. Consequently, the propagation constant can be controlled by modifying the effective permittivity of the guiding region.

In the proposed phase shifter, the effective permittivity is electrically tuned through the FPMS unit cells. A change in capacitance modifies the effective permittivity of the guiding region, thereby changing the propagation constant of the guided mode. Consequently, different capacitance states produce different phase delays over the same physical length. The phase difference between two tuning states can therefore be expressed as

$$\Delta\phi = (\beta_2 - \beta_1)L \quad (4.3)$$

where β_1 and β_2 are the propagation constants corresponding to two different capacitance states.

This expression Eq. (4.3) demonstrates that the achievable phase range depends directly on the variation of the propagation constant across the tuning states. Therefore, maximizing the change in effective permittivity leads to a larger differential phase shift. Although the structure is implemented using discrete programmable cells, it operates as a distributed electromagnetic

medium. Consequently, the phase constant is determined by the collective response of the entire dielectric region rather than by the characteristics of any individual unit cell. This behavior captures the underlying mechanism responsible for smooth and nearly continuous phase tuning. Furthermore, because the phase shift is achieved by modifying the propagation characteristics of the medium itself, the design avoids the need for multiple cascaded delay sections. This approach provides a practical trade-off between phase tuning range, insertion loss, and implementation complexity while maintaining stable performance within a compact footprint. Therefore, the operating principle of the proposed phase shifter is based on direct control of the effective permittivity and, consequently, the propagation constant of a guided wave. Since the accumulated phase delay is proportional to the propagation constant, precise adjustment of the varactor capacitance enables continuous tuning of the output phase. This principle forms the basis for the analytical modeling and numerical validation presented in the subsequent sections of this chapter.

4.3.1 Waveguide-Like Behavior Using Negative-Permittivity Sidewalls

One of the most distinctive features of the proposed phase shifter is its ability to emulate waveguide-like behavior using a fully planar and electronically programmable structure. In a conventional rectangular waveguide, EM energy is confined by metallic sidewalls that impose a perfect electric conductor boundary condition. These boundaries force the tangential electric field to vanish at the walls, thereby restricting the propagation of the EM wave to the interior dielectric region. In the proposed design, an analogous confinement mechanism is achieved without employing physical metallic walls. Instead, the required boundary condition is synthesized by programming selected unit cells to exhibit negative effective permittivity. When the effective permittivity of a material becomes negative, the propagation constant becomes predominantly

imaginary and the electromagnetic field decays exponentially inside the medium resulting in evanescent wave. This behavior can be expressed as Eq. (4.4)

$$E(z) = E_0 e^{-\alpha z} \quad (4.4)$$

where α denotes the attenuation constant. Since the wave is unable to propagate through such a region, the medium effectively reflects the incident field and suppresses energy penetration. Consequently, the negative-permittivity region behaves in a manner analogous to a metallic boundary over the frequency range of interest. This phenomenon can also be interpreted through the wave impedance of the medium. For a non-magnetic material, the intrinsic impedance is given by Eq. (4.5)

$$\eta = \sqrt{\frac{\mu_0}{\epsilon_0 \epsilon_{\text{eff}}}} \quad (4.5)$$

When $\epsilon_{\text{eff}} < 0$, the impedance becomes purely imaginary, indicating that the medium stores reactive energy rather than supporting traveling waves. Therefore, the boundary presents a high impedance to the incident field and prevents significant power transmission through the region. The complete phase shifter consists of five columns of programmable unit cells. The first and fifth columns are biased at approximately 1 pF, which corresponds to a negative effective permittivity at the operating frequency of 2.5 GHz which can be observed from Fig. 4.4. As a result, these outer columns suppress lateral field leakage and serve as electronically synthesized sidewalls. In contrast, the three intermediate columns are assigned capacitance values that produce positive effective permittivity. These columns form the central dielectric region through which the dominant guided mode propagates.

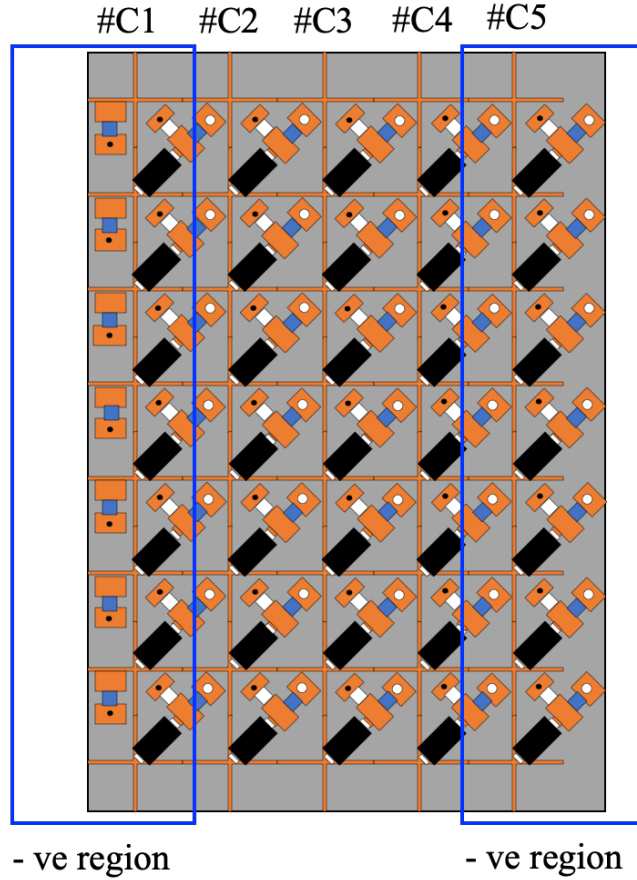


Fig. 4.4. Proposed phase shifter with highlighted negative permittivity regions

This arrangement establishes a planar equivalent of a dielectric-filled rectangular waveguide. The outer columns define the electromagnetic boundaries, while the intermediate columns determine the propagation characteristics of the guided wave. Since the effective permittivity of the central region can be adjusted continuously, the propagation constant of the dominant mode can be dynamically modified without altering the physical dimensions of the structure. Consequently, the same architecture simultaneously provides field confinement and phase control. Although the synthesized boundaries are not ideal perfect conductors, they exhibit sufficiently high attenuation to prevent substantial energy penetration. Therefore, the dominant portion of the electromagnetic field remains concentrated within the central propagation region. This confinement enables the structure to support a well-defined guided mode whose phase response

can be accurately described using dielectric slab and transfer-matrix models presented in the subsequent sections. Another important advantage of this approach is its reconfigurability. Since the boundary behavior is established through electronic biasing rather than fixed metallic walls, the same programmable substrate can be reconfigured to realize different electromagnetic functions. For example, selected columns may serve as wave-confining boundaries in one configuration and as part of the propagation region in another. This capability offers considerable flexibility for the realization of adaptive microwave components. Therefore, the use of negative-permittivity sidewalls provides an elegant and physically intuitive mechanism for synthesizing waveguide-like behavior within a planar programmable substrate. By electronically imposing effective boundary conditions, the proposed structure confines the guided wave to a tunable dielectric core, thereby forming the foundation for the continuously reconfigurable phase shifter developed in this chapter.

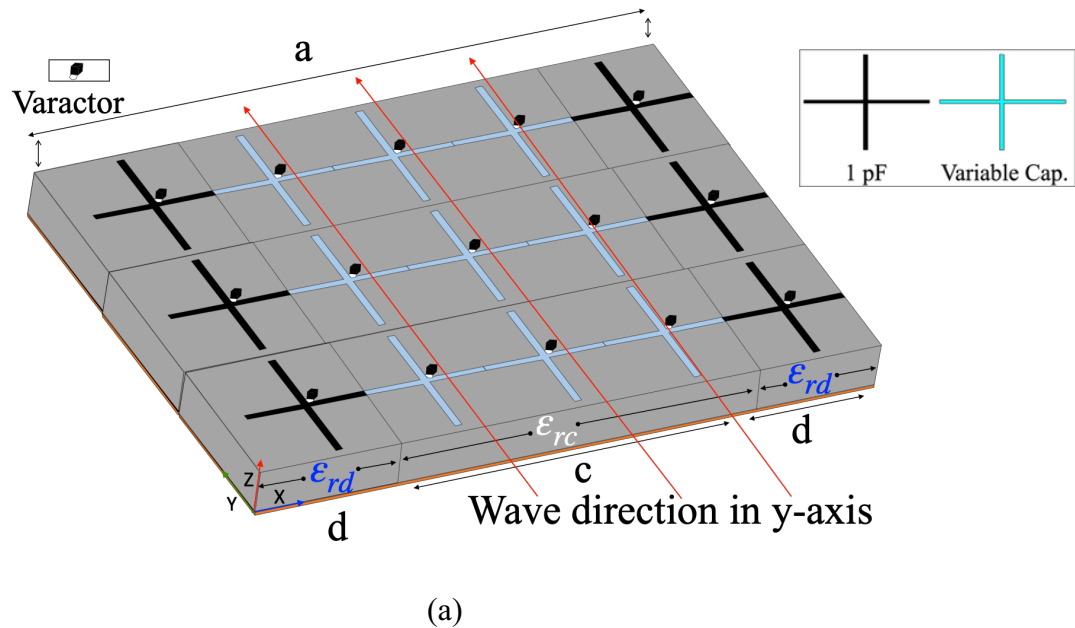
4.4 Analytical Modeling of Wave Propagation

Building upon the physical configuration introduced in the previous section, the objective here is to develop a rigorous analytical framework that captures how the guided wave propagates through the programmable dielectric region. The analytical treatment is divided into two closely connected parts. First, the central propagation region is modeled as a homogeneous dielectric slab in which all intermediate columns are assigned the same capacitance. This simplified representation provides fundamental insight into the relationship between effective permittivity and propagation constant. Next, the formulation is extended to non-homogeneous configurations, where the capacitance of each column can be independently controlled and the dielectric properties vary across the width of the structure. By operating the structure both in homogeneous and non-homogeneous modes shows the versatility of the FPMS concept. It is because of this dynamic

nature that FPMS can provide efficient reconfigurability which is generally absent from other techniques and methodologies.

4.4.1 Homogeneous Dielectric Model

Initially, the 3 middle columns of the structure are all deemed to have same applied bias resulting in a homogeneous dielectric permittivity for the incoming wave, Fig. 4.5 (a). The 2 columns at the extreme form the boundaries by providing negative permittivity to the propagating wave. Since the EM properties are uniform across the entire propagation region, the structure can be represented by a single dielectric layer bounded by two identical side regions that emulate the synthesized waveguide walls. This simplified representation captures the essential mechanism governing phase tuning of the propagating wave. Although the actual structure is composed of discrete programmable cells, it is simpler to model the guided structure as a single unit for ease of developing the theoretical model, Fig. 4.5 (b).



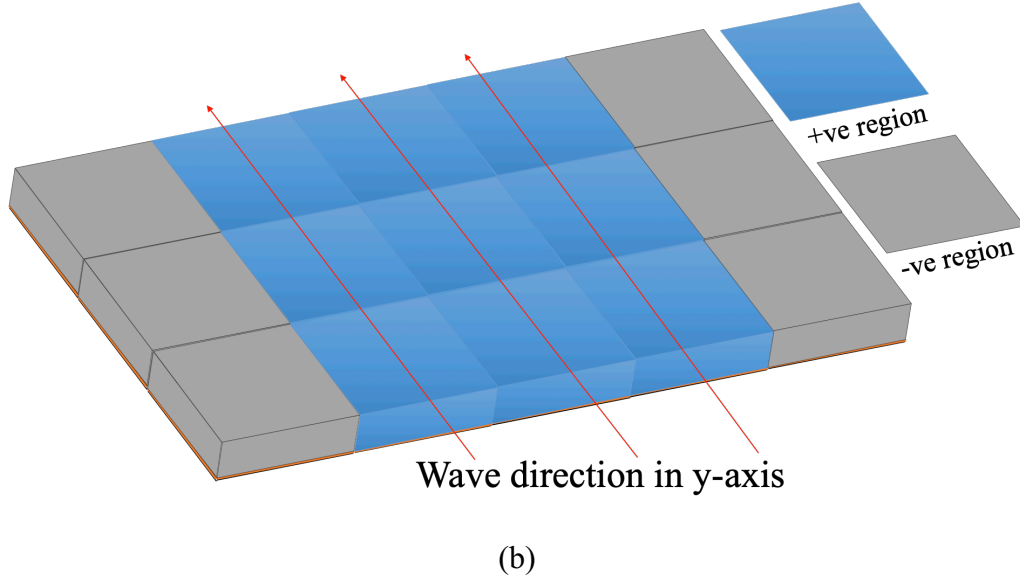


Fig. 4.5(a) Homogeneous setup with FPMS unit cells (b) Analogous of homogeneous setup

The central dielectric region with uniform permittivity is defined with a c and effective permittivity ϵ_{rc} , whereas the two outer boundaries have identical width d with an effective permittivity ϵ_{rd} . At the operating frequency of 2.5 GHz, the outer regions are biased to approximately 1 pF to model the evanescence of wave in these regions. In contrast, the central region supports the dominant guided mode and determines the overall propagation constant of the structure.

Assuming propagation along the y -direction, the transverse electric field in each region satisfies the one-dimensional Helmholtz equation [83],

$$\frac{d^2 E_z}{dx^2} + (k_0^2 \epsilon_r - \beta^2) E_z = 0 \quad (4.6)$$

where k_0 is the free-space wavenumber and β is the unknown propagation constant. By enforcing continuity of the tangential electric and magnetic fields at the interfaces between the dielectric regions, the propagation constant of the dominant TE_{10} mode is obtained from the transcendental equation (4.7) originally presented in [83]:

$$\sqrt{k_c^2 - \beta_{10}^2} \cdot \tan \left[\left(\frac{a - c}{2} \right) \sqrt{k_d^2 - \beta_{10}^2} \right] = \sqrt{k_d^2 - \beta_{10}^2} \cdot \cot \left[\frac{c}{2} \sqrt{k_c^2 - \beta_{10}^2} \right] \quad (4.7)$$

Here, β_{10} is the propagation constant of TE₁₀ mode, $k_d = \omega \sqrt{\epsilon_{rd}} / c_0$, $k_c = \omega \sqrt{\epsilon_{rc}} / c_0$ are the wave numbers in the two regions, ω is the angular frequency, c_0 is the speed of light. The values of a , c and d can be extracted from full-wave simulations. Using the transcendental equation of 4.7 one can plot the curves for the phase constant, β , for the varying values of capacitance, Fig. 4.6. This equation establishes the fundamental relationship between the effective permittivity of the central region and the phase constant of the supported mode. The transcendental equation is solved numerically using MATLAB to determine the corresponding value of β . It can be observed that the value of β increases systematically as the capacitance is increased. Once the phase constant is obtained, the accumulated phase delay is calculated using the relationship introduced in Section 4.3, namely $\phi = \beta L$. Therefore, the homogeneous dielectric model provides a computationally efficient and physically transparent method for predicting the phase response when all intermediate columns are tuned identically. The phase difference between two bias states over a structure of length L is given by $\Delta\phi = (\beta_1 - \beta_2)L$. Thus, one can evaluate the phase shift or deference generated by the structure as the bias is varied. Although this model assumes a uniform dielectric core, it captures the dominant behavior of the phase shifter and forms the basis for understanding the full-wave simulation results. Moreover, it serves as a reference solution against which more general non-homogeneous configurations can be evaluated. Consequently, the homogeneous dielectric model establishes the theoretical foundation for the broader analytical framework developed in the following subsection.

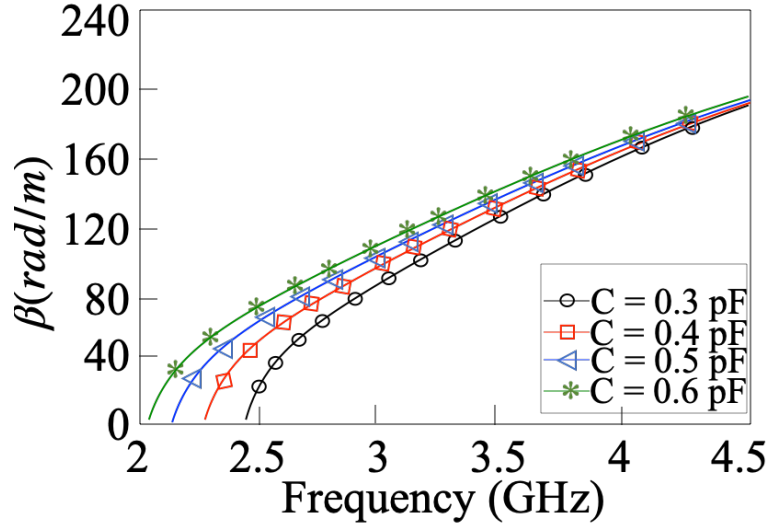


Fig. 4.6 Dispersion graph of Homogeneous setup

4.4.2 Non-Homogeneous Dielectric Model

The FPMS concept is versatile enough that it should be studied for a range of operation conditions. Therefore, it is pertinent to study this design in non-homogeneous bias scenario also. Since each programmable section can be assigned an independent capacitance value, the effective permittivity may vary across the transverse direction of the structure. Under this condition, the dielectric region becomes non-homogeneous, and the propagation constant can no longer be obtained directly from the three-region transcendental equation used for the homogeneous case, (4.7). Therefore, a more reliable analytical formulation is required for this scenario. To model this condition, the transverse cross section of the phase shifter is divided into several dielectric regions, as shown in Fig. 4.7. Each region is assumed to be locally homogeneous, while the effective permittivity may vary from one section to another. In this formulation, the unit-cell sections are arranged along the x -direction, whereas the guided wave propagates along the y -direction. For the i -th dielectric region, the effective permittivity is denoted by ϵ_{ri} , and the width of the region along the x -direction is denoted by w_i . For the dominant TE_{10} -like mode, the out-of-

plane magnetic field component is used to describe the transverse field variation. Thus, the magnetic field in the i -th region can be written as Eq. (4.8):

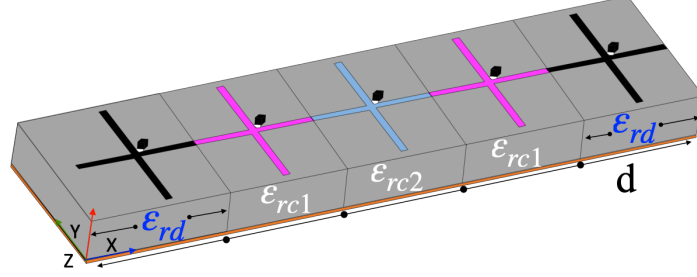


Fig. 4.7 Non-homogeneous setup with FPMS unit cells

$$H_z^{(i)}(x, y) = H_z^{(i)}(x)e^{-j\beta z} \quad (4.8)$$

where β is the phase constant along the y -direction. Substituting this field dependence into the scalar wave equation gives the one-dimensional Helmholtz equation (4.9) in the i -th dielectric region:

$$\frac{d^2 H_z^{(i)}}{dx^2} + (k_0^2 \epsilon_{ri} - \beta^2) H_z^{(i)} = 0 \quad (4.9)$$

where k_0 is the free-space wavenumber. The transverse wavenumber in the i -th region is therefore defined as Eq. (4.10):

$$k_{xi} = \sqrt{k_0^2 \epsilon_{ri} - \beta^2} \quad (4.10)$$

The general solution for the magnetic field inside the i -th region can be expressed as Eq. (4.11):

$$H_z^{(i)}(x_i) = A_i \cos(k_{xi} x_i) + B_i \sin(k_{xi} x_i) \quad (4.11)$$

where A_i and B_i are unknown coefficients, and x_i is the local coordinate measured from the left boundary of the i -th region. The corresponding tangential electric field is obtained from Maxwell's equation (4.12):

$$\nabla \times \mathbf{H} = j\omega\epsilon_0\epsilon_{ri}\mathbf{E} \quad (4.12)$$

The y -directed electric field is related to the derivative of H_z with respect to x . Therefore,

$$E_y^{(i)}(x_i) = -\frac{1}{j\omega\epsilon_0\epsilon_{ri}} \frac{dH_z^{(i)}(x_i)}{dx_i} \quad (4.13)$$

Substituting the magnetic-field solution gives Eq. (4.14):

$$E_y^{(i)}(x_i) = -\frac{k_{xi}}{j\omega\epsilon_0\epsilon_{ri}} [-A_i \sin(k_{xi}x_i) + B_i \cos(k_{xi}x_i)] \quad (4.14)$$

This can be written as Eq. (4.15):

$$E_y^{(i)}(x_i) = \frac{k_{xi}}{j\omega\epsilon_0\epsilon_{ri}} [A_i \sin(k_{xi}x_i) - B_i \cos(k_{xi}x_i)] \quad (4.15)$$

For compact notation, the modal admittance of the i -th region is defined as Eq. (4.16):

$$Y_i = \frac{k_{xi}}{j\omega\epsilon_0\epsilon_{ri}} \quad (4.16)$$

Thus, the field expressions become (4.17 and 4.18):

$$H_z^{(i)}(x_i) = A_i \cos(k_{xi}x_i) + B_i \sin(k_{xi}x_i), \quad (4.17)$$

$$E_y^{(i)}(x_i) = Y_i [A_i \sin(k_{xi}x_i) - B_i \cos(k_{xi}x_i)] \quad (4.18)$$

The transfer matrix method relates the tangential fields at the two sides of each dielectric region [70]. At the left boundary of the i -th region, where $x_i = 0$, the fields are described in Eq. 4.19 and Eq.4.20:

$$H_z^{(i)}(0) = A_i, \quad (4.19)$$

$$E_y^{(i)}(0) = -Y_i B_i. \quad (4.20)$$

Therefore,

$$A_i = H_z^{(i)}(0), \quad (4.21)$$

$$B_i = -\frac{E_y^{(i)}(0)}{Y_i} \quad (4.22)$$

At the right boundary of the same region, where $x_i = w_i$, the fields are defined in Eq. 4.23 and Eq. 4.24:

$$H_z^{(i)}(w_i) = A_i \cos(k_{xi} w_i) + B_i \sin(k_{xi} w_i), \quad (4.23)$$

$$E_y^{(i)}(w_i) = Y_i [A_i \sin(k_{xi} w_i) - B_i \cos(k_{xi} w_i)] \quad (4.24)$$

Substituting $A_i = H_z^{(i)}(0)$ and $B_i = -E_y^{(i)}(0)/Y_i$, the fields at $x_i = w_i$ become

$$H_z^{(i)}(w_i) = H_z^{(i)}(0) \cos(k_{xi} w_i) - \frac{1}{Y_i} E_y^{(i)}(0) \sin(k_{xi} w_i), \quad (4.25)$$

$$E_y^{(i)}(w_i) = Y_i H_z^{(i)}(0) \sin(k_{xi} w_i) + E_y^{(i)}(0) \cos(k_{xi} w_i). \quad (4.26)$$

Hence, the fields at the right side of the region can be written in matrix form as Eq. 4.27:

$$\begin{bmatrix} H_z^{(i)}(w_i) \\ E_y^{(i)}(w_i) \end{bmatrix} = \begin{bmatrix} \cos(k_{xi} w_i) & -\frac{1}{Y_i} \sin(k_{xi} w_i) \\ Y_i \sin(k_{xi} w_i) & \cos(k_{xi} w_i) \end{bmatrix} \begin{bmatrix} H_z^{(i)}(0) \\ E_y^{(i)}(0) \end{bmatrix} \quad (4.27)$$

Equivalently, the fields at the left side can be related to the fields at the right side as Eq. (4.28):

$$\begin{bmatrix} H_z^{(i)}(0) \\ E_y^{(i)}(0) \end{bmatrix} = \begin{bmatrix} \cos(k_{xi}w_i) & \frac{1}{Y_i} \sin(k_{xi}w_i) \\ -Y_i \sin(k_{xi}w_i) & \cos(k_{xi}w_i) \end{bmatrix} \begin{bmatrix} H_z^{(i)}(w_i) \\ E_y^{(i)}(w_i) \end{bmatrix} \quad (4.28)$$

This matrix is the characteristic transfer matrix of the i -th dielectric section. For a non-homogeneous dielectric medium consisting of multiple adjacent regions along the x -direction, the total transfer matrix is obtained by cascading the individual matrices [72]. If the structure is divided into five regions, the total transfer matrix is written as Eq. 4.29:

$$\mathbf{T}_{\text{total}} = \mathbf{T}_1 \mathbf{T}_2 \mathbf{T}_3 \mathbf{T}_4 \mathbf{T}_5 \quad (4.29)$$

where

$$\mathbf{T}_i = \begin{bmatrix} \cos(k_{xi}w_i) & \frac{1}{Y_i} \sin(k_{xi}w_i) \\ -Y_i \sin(k_{xi}w_i) & \cos(k_{xi}w_i) \end{bmatrix} \quad (4.30)$$

Therefore, the fields at the left and right boundaries of the complete non-homogeneous structure are related as Eq. (4.31):

$$\begin{bmatrix} H_L \\ E_L \end{bmatrix} = \mathbf{T}_{\text{total}} \begin{bmatrix} H_R \\ E_R \end{bmatrix} \quad (4.31)$$

where H_L and E_L are the tangential fields at the left boundary, while H_R and E_R are the corresponding fields at the right boundary. The phase constant is obtained by applying the boundary conditions at the outer walls of the structure. Since the side boundaries are treated as effective evanescent regions (negative permittivity), the tangential electric field must vanish at both boundaries. Therefore,

$$\begin{aligned} E_L &= 0, \\ E_R &= 0 \end{aligned} \quad (4.32)$$

Using the total matrix relation,

$$\begin{bmatrix} H_L \\ 0 \end{bmatrix} = \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} \begin{bmatrix} H_R \\ 0 \end{bmatrix} \quad (4.33)$$

Expanding this equation gives Eq. (4.34):

$$\begin{aligned} H_L &= T_{11} H_R, \\ 0 &= T_{21} H_R \end{aligned} \quad (4.34)$$

For a non-trivial guided mode, $H_R \neq 0$. Therefore, the characteristic equation becomes Eq. (4.35)

$$T_{21} = 0 \quad (4.35)$$

This equation is the transcendental condition for the allowed guided modes of the non-homogeneous dielectric medium. Since T_{21} depends on β through k_{xi} , the equation is implicit in the phase constant and must be solved numerically at each frequency. Since the phase shifter consists of 5 columns, each column can be assigned an independent capacitance value. However, to keep the present study concise, only one representative non-uniform configuration is considered which is shown in Fig. 4.8.

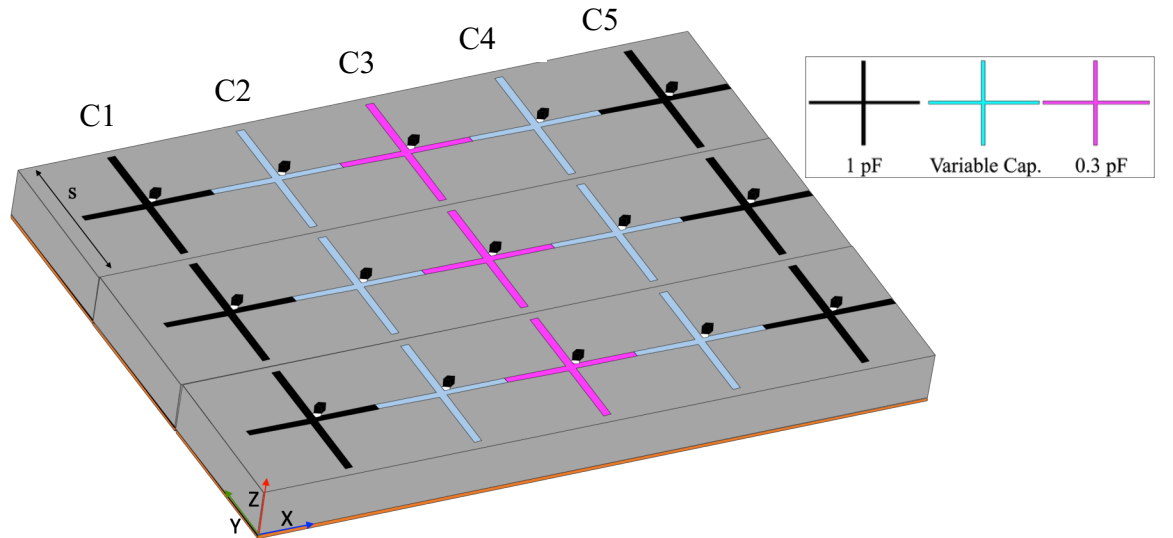


Fig. 4.8 Non-homogeneous example case

1st and 5th columns are fixed at 1 pF to emulate the side boundaries, while Columns *C2 and *C4 are maintained at 0.3 pF. The capacitance of the center column, *C3, is then tuned independently. This arrangement allows the dielectric profile across the structure to become non-uniform, enabling the effect of localized permittivity variation on the propagation constant and the resulting phase shift to be examined. The corresponding dispersion curves are included in Fig. 4.9 and are represented by dashed lines. It can be observed that the non-uniform configurations produce intermediate phase constants that are not obtained under uniform tuning. This provides additional flexibility in tailoring the phase response without changing the physical dimensions of the phase shifter. Although the non-uniform dielectric profile may introduce slight field distortion and a small increase in insertion loss, these effects are expected to have only a minor impact on the overall performance. Other non-uniform capacitance distributions can also be realized by independently controlling Columns *C2 to *C4. To avoid unnecessary redundancy, only one representative configuration is presented in this work. Based on the dispersion analysis presented above, the theoretical phase response can be calculated and compared with full-wave simulations.

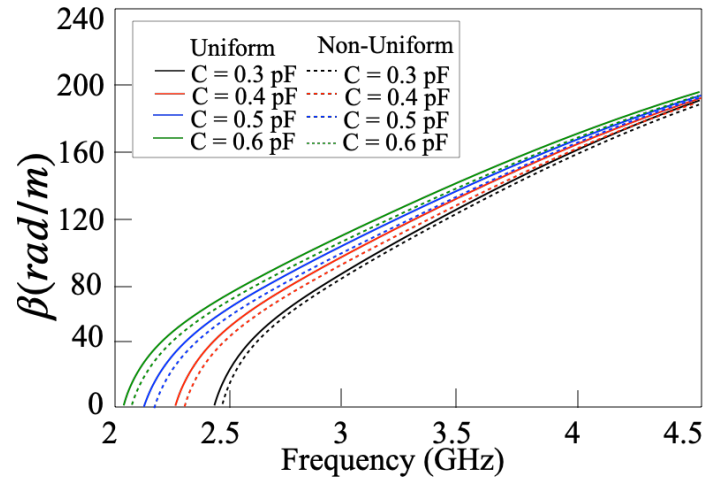


Fig. 4.9 Dispersion graph of non-homogeneous case compared with homogeneous case

4.5 Phase Shifter Simulations

The phase shifter consists of three major sections: 1) the interconnected FPMS unit cells, 2) the tapered microstrip feed line, and 3) the biasing network. The complete schematic of the proposed phase shifter is illustrated again in Fig. 4.10 for ease of the reader. Fig. 4.10(a) shows that all unit cells are physically connected together, forming a mesh-like structure. These unit cells are integrated with lumped elements including varactor diodes, direct current block capacitors, and resistors. Once the FPMS unit cells are designed and optimized, they are integrated with tapered microstrip feed lines which is shown in Fig. 4.10(b). The tapered feed line is incorporated to improve the impedance matching and reflection coefficient characteristics of the overall structure. The final section of the proposed design is the biasing network, which is responsible for controlling the capacitance of the varactor diodes which is shown in Fig. 4.10(c).

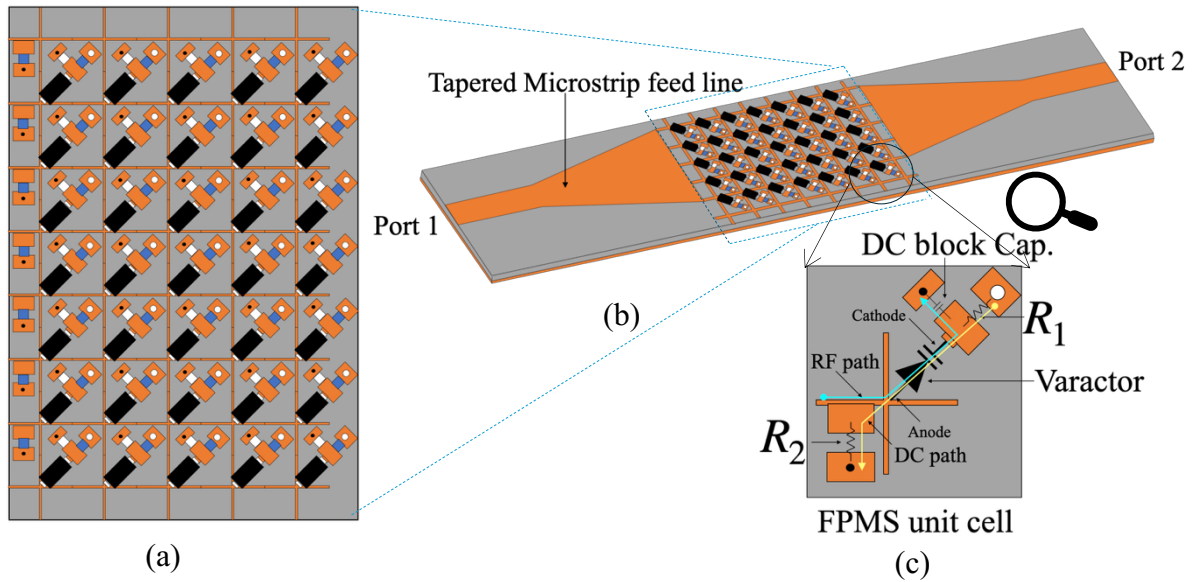


Fig. 4.10 (a) Mesh structure of FPMS unit cells (b) 3D perspective view of phase shifter (c) Zoom in image of FPMS unit cell with biasing strategy

In the proposed configuration, the anode terminal of the varactor diode is directly connected to the metallic unit cell, whereas the cathode terminal is connected to resistor R_1 and the DC block capacitor. To operate the varactor diode as a variable capacitor, it must remain under reverse bias condition. Therefore, the cathode terminal is maintained at a higher positive voltage compared to the anode terminal. This direct current voltage is supplied through resistor R_1 . The DC path is highlighted with yellow color line. Since RF waves propagate through the same structure, careful consideration is required for DC biasing to ensure proper isolation between the RF signal and the DC. The RF path is highlighted with a blue ice color. In the proposed configuration, resistor R_1 acts as a radio frequency choke, which suppresses RF leakage into the DC source. To simulate the proposed phase shifter in Ansys HFSS, lumped ports are utilized for excitation. One lumped port is assigned as the input excitation source, whereas the second port is terminated with a $50\ \Omega$ load to emulate standard microwave measurement conditions.

The simulations are to be carried out in two stages. One for the homogeneous scenario and the other takes into account the non-homogeneous case. For the former case, the capacitors in the middle columns (i.e. *C2 to *C4) are all kept at the same value and are varied from 0.3 pF to 0.6 pF while the corner capacitors *C1 and *C5 are kept constant at 1 pF. This scenario models the homogeneous case. The simulated results of the phase shifter are shown in Fig. 4.9. It can be observed that the increase in the capacitor values causes an increase in the phase shift response. The plot only shows the range from 0.3 pf to 0.6 pF considering the best impedance matching scenario at the two ports in Fig. 4.11. A maximum phase tuning of $\sim 300^\circ$ is achieved at 2.5 GHz which is illustrated in Fig. 4.12. The insertion loss of the design does not deteriorate above 2.5 dB which includes the loss from the feedlines, Fig. 4.13. These results show quite a promising

performance from the proposed FPMS phase shifter design for the homogeneous case of bias and compares well with theory. This comparison will be discussed together with the measurements.

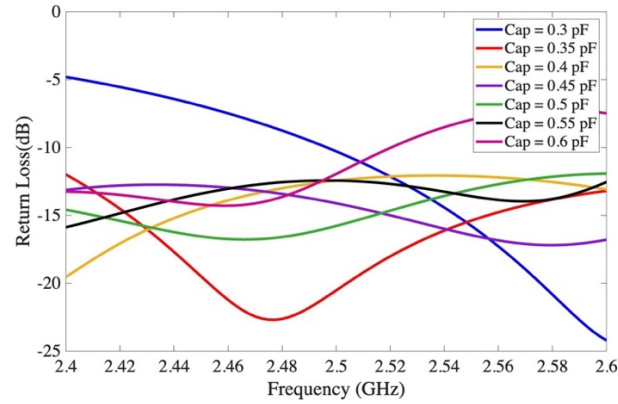


Fig. 4.11. Simulated reflection coefficient of the proposed Phase Shifter for homogeneous case

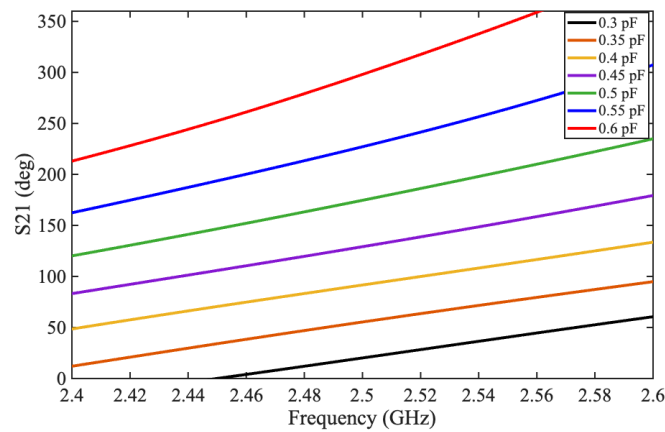


Fig. 4.12. Simulated phase of the proposed Phase Shifter for homogeneous medium

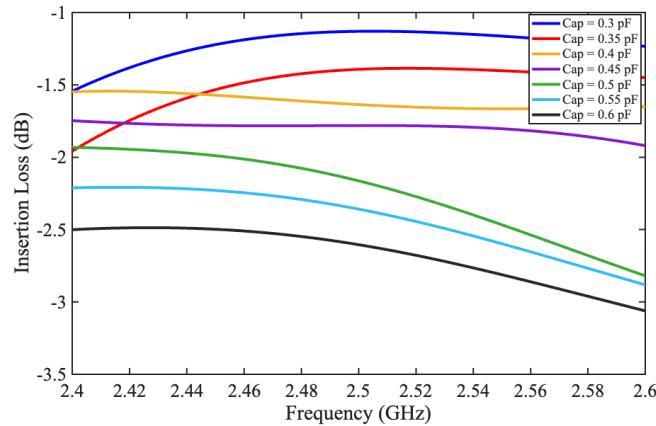


Fig. 4.13. Simulated Insertion loss for homogeneous case

The next step is to study the same design for the non-homogeneous case. For the non-homogeneous case, it would be pertinent to show the analytical results with the simulations at this point as there can be several permutations and a reference would be needed. This comparison is shown in Fig. 4.14. Three different cases of bias scenarios are chosen to study in this part of the thesis. In all cases, Columns *C1 and *C5 are fixed at 1 pF so that they exhibit negative permittivity and act as walls for the propagating waves.

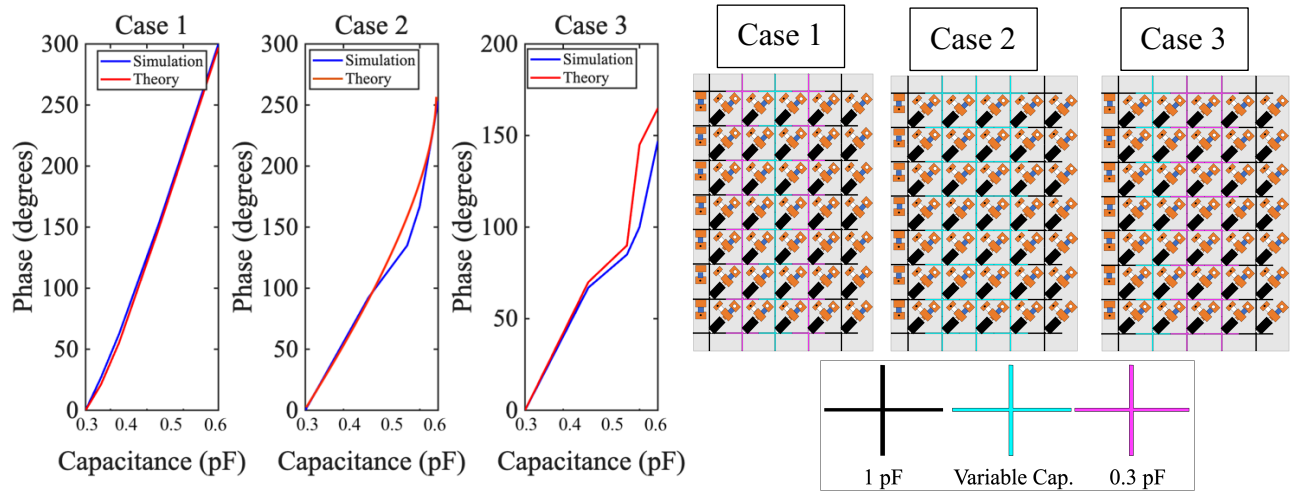


Fig. 4.14. Comparison of simulated phase shift with the proposed theory (non-homogeneous) at 2.5 GHz

Case 1 corresponds to the guided structure, where columns *C2 and *C4 are assigned the same capacitance value of 0.3 pF and are kept constant while *C3 is varied. In the scenario of case 2 Columns *C2 and *C4 are fixed at 0.3 pF, while column *C3 is tuned independently. In other words, case 1 and case 2 are exactly opposite to each other. A third permutation considered here is case 3, where column *C2 is tuned independently and Columns *C3 and *C4 are fixed at 0.3 pF. The three non-homogeneous scenarios are chosen as test case studies. Other combinations are also studied in simulations but the results do not differ or improve much from the ones shown in Fig. 4.9. The theoretical and simulated phase shifts compare well with very slight deviation between

the two, Fig. 4.9. For all three cases, the variable capacitor is changed from 0.3 to 0.6 pF. Here, the relative phase shift at 2.5 GHz is plotted as a function of capacitance, with 0.3 pF being the reference value. The close correlation between the theory and simulations provides a good initial validation of the analytical model. Among the three configurations, Case 1 provides the best phase-tuning range, reaching approximately $\sim 300^\circ$ which is quite comparable to the homogeneous case but still slightly lower than it. Case 2 and Case 3 on the other hand result in 240° and 180° , respectively. These results indicate that the homogeneous configuration offers the best results which is not surprising. Since the homogeneous case provides a larger effective substrate area with tunable permittivity, the wave experiences more change in its impedance for this case. Thus, the resulting effect is demonstrated herewith. These results of the phase shifter design are quite encouraging and provide a sound platform for the component to be used in actual beam steering application.

4.6 Fabrication and Measurements

The fabricated prototype of the proposed phase shifter is shown in Fig. 4.15. The process/steps involved in the fabrication of the phase shifter are quite adjacent to the ones reported in the last chapter. The prototype is implemented on a single Rogers substrate with a thickness 1.575 mm. For the DC biasing of the component, jumper wires are soldered on the bottom side of the board and connected to the varactor diodes mounted on the top layer. Since the phase shifter contains 35 varactor diodes, a total of 35 jumper-wire connections is required. The design could have been improved by adding an extra PCB layer for biasing alone. Since the varactors in one column need to have the same bias value, they all could be integrated to a single DC biasing line. For this purpose, an extra PCB board would have been needed. However, that results in a significant

increase in the overall cost of the prototype fabrication. Therefore, individual connection for each diode is kept for this design which somewhat complicates the biasing but can be overlooked for economical reasons. Different color-coded jumper wires are used to simplify the biasing arrangement during the measurements. The jumper wires with black and blue colors correspond to columns 1 and 5, respectively. As discussed in the previous section, these two outer columns emulate the synthesized sidewalls of the dielectric waveguide. To realize this behavior, both columns are permanently biased at the minimum reverse-bias voltage of 0 V, corresponding to the maximum capacitance state of the SMV2023-011LF-ND varactor diode. Increasing the capacitance shifts the resonant frequency of the unit cell toward lower frequencies, causing the operating frequency of 2.5 GHz to lie within the negative-permittivity region. Consequently, EM waves cannot propagate through these outer columns, allowing them to behave as synthesized electromagnetic walls that confine the guided wave within the central propagation region.

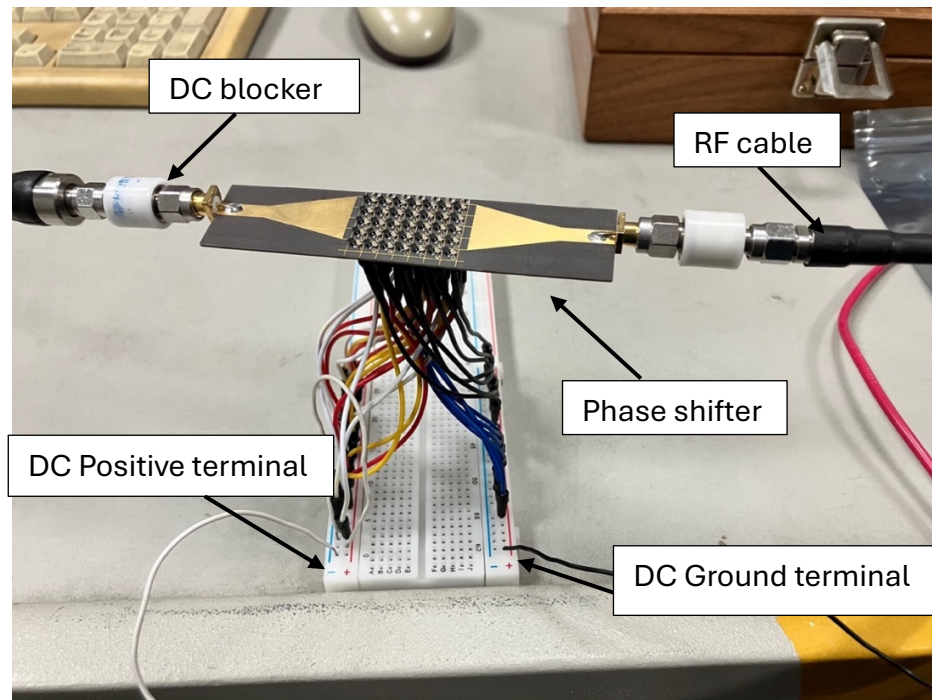


Fig. 4.15. Close view of phase shifter testing setup

The remaining 3 central columns (columns 2–4) form the programmable dielectric region of the phase shifter. At first, the homogeneous dielectric configuration is tested and therefore all 3 columns are biased using the same DC voltage. The reverse-bias voltage is varied from 6 V to 20 V, thereby changing the capacitance of all central unit cells simultaneously and producing different effective permittivity and corresponding phase responses. Another important aspect of the measurement setup is the use of DC blockers. Since the prototype incorporates active varactor diodes, the applied DC bias may leak into the RF transmission path and potentially damage the vector network analyzer (VNA). To prevent this, DC blockers are connected at both the input and output ports of the phase shifter, ensuring complete isolation between the DC bias network and the RF measurement system. The whole setup of phase shifter is shown in Fig. 4.16. During the measurements, the center frequency of the VNA is set to 2.5 GHz, while the frequency span is selected from 2.4 to 2.6 GHz to observe the broadband response of the phase shifter. The voltage of the power supply is varied from 6 V to 20 V. According to the varactor datasheet, a reverse-bias voltage of 6 V corresponds to a capacitance of approximately 0.6 pF, whereas in the case of 20 V the capacitance value reduces to ~0.3 pF. These capacitance values are identical to those used in the simulations, enabling a direct comparison between the simulated and measured phase responses. The measured insertion loss and reflection coefficient characteristics of the phase shifter in the homogeneous case are presented in Fig. 4.17 and Fig. 4.18, respectively. At the operating frequency, the loss varies approximately from 2.5 to 4.8 dB for the different biasing states with acceptable matching conditions. The higher-than-expected loss may be attributed to fabrication and assembly imperfections, including soldering losses, component tolerances, conductor and dielectric losses. Also, this loss includes the input and output feedlines that are used to excite the signal using VNA.

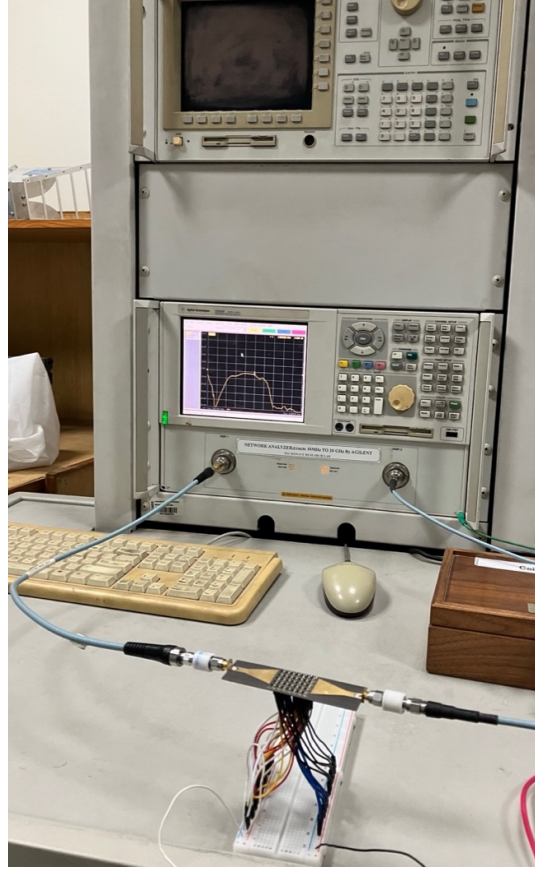


Fig. 4.16. VNA measurement setup. (The DC power supply is positioned to the right of the VNA and is not visible in the photograph)

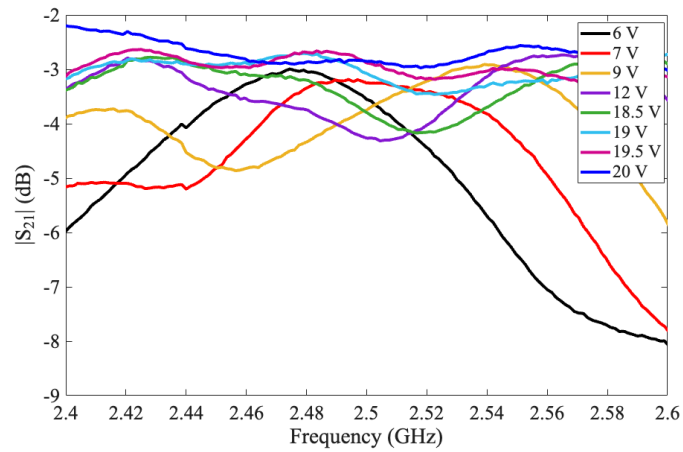


Fig. 4.17. Insertion loss of phase shifter

As the reverse-bias voltage increases, the capacitance of the varactor diode decreases, which shifts the unit-cell resonance and the corresponding effective-permittivity response toward higher

frequencies. Consequently, at the operating frequency of 2.5 GHz, the unit cells move farther into the positive-permittivity region, where the guided wave experiences improved transmission. This behavior is generally reflected in Fig. 4.17, as the higher-voltage states, particularly between 19 and 20 V, exhibit lower insertion loss, with $|S_{21}|$ remaining close to -3 dB. However, the variation is not strictly monotonic for every bias state because the insertion loss is also influenced by impedance matching, fabrication tolerances, varactor losses, and resonant interactions within the complete phase shifter structure. Therefore, the results generally support the observation that lower capacitance improves transmission at 2.5 GHz, although the exact insertion loss depends on the overall response of the fabricated device. The reflection coefficient of the phase shifter for all cases if bias is lower than -10 dB establishing acceptable matching conditions, Fig. 4.18. The corresponding measured phase response of the phase shifter is presented in Fig. 4.19, showing the variation in transmission phase obtained by changing the applied bias voltage. The maximum phase shift is around 298° . The measured phase shift shows good agreement with the simulated results, with a maximum deviation of approximately $\pm 3^\circ$ – 4° . A comparison of the measured results for the homogeneous case with the corresponding simulated and theoretical results is presented in Fig. 4.20. The results show good agreement among all three approaches.

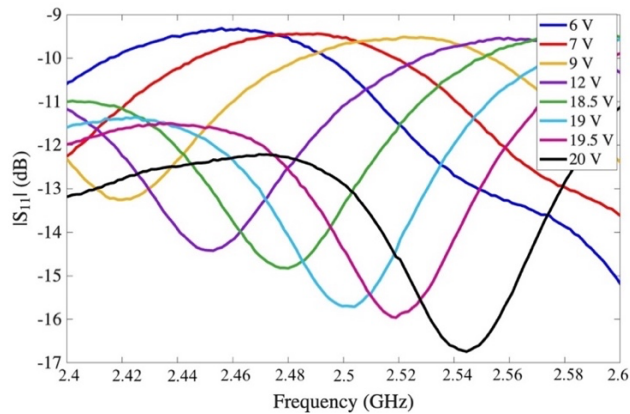


Fig. 4.18. Reflection coefficient of phase shifter

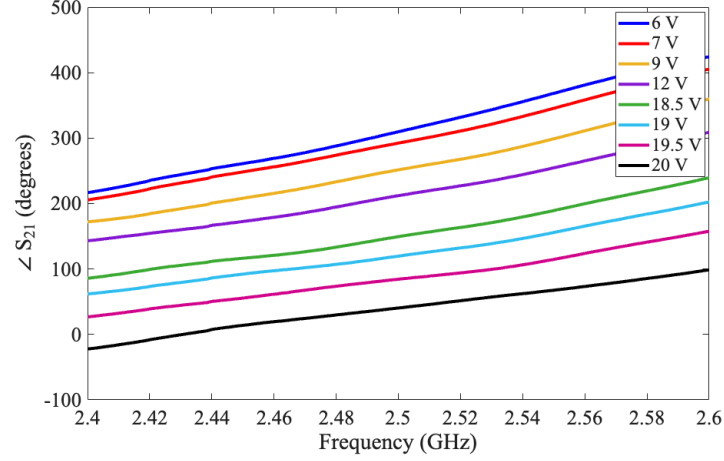


Fig. 4.19. Phase shift of phase shifter at different voltages

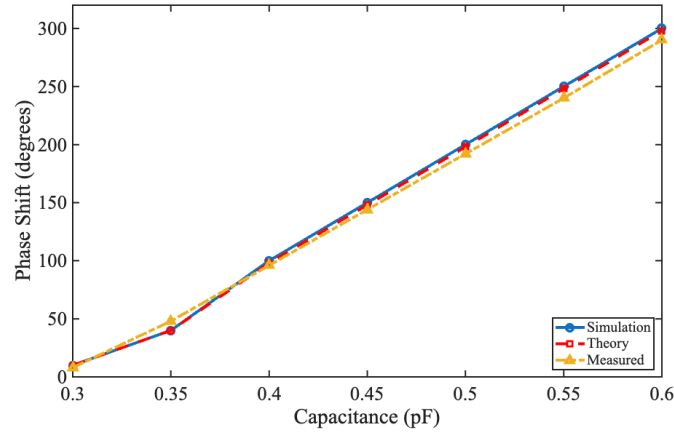


Fig. 4.20. Comparison of phase shift at 2.5 GHz for homogeneous case

In the non-homogeneous configuration, different capacitance values are assigned to each column of the phase shifter. For the sake of conciseness, only case I from Section 4.5 is considered, where Columns *C1 and *C5 are assigned a capacitance of 1 pF, while the center column *C1 is assigned with variable capacitance. Also, it is worth remembering that case I provides the best phase shifter results among the ones used in this thesis. The capacitance values of columns *C2 and *C4 are fixed at 0.3 pF. Compared with the homogeneous configuration, the non-homogeneous setup exhibits higher insertion loss, which is consistent with the theoretical analysis and simulations presented in the previous section. This increase in insertion loss is expected because the guided wave propagates through multiple dielectric regions having different effective permittivities,

resulting in additional impedance discontinuities and scattering at the interfaces between adjacent sections. The measured insertion loss responses for different capacitance values are shown in Fig. 4.21. The corresponding phase responses are presented in Fig. 4.22, where the phase shift varies from 0° to $\sim 284^\circ$, demonstrating that the proposed phase shifter is capable of providing a wide phase tuning range under non-homogeneous biasing conditions.

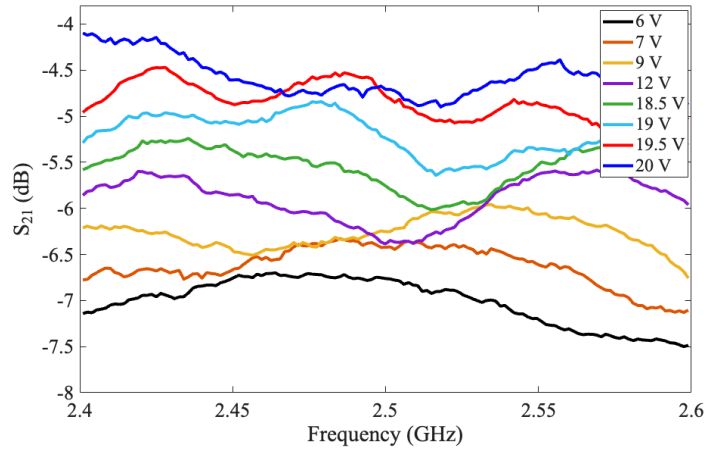


Fig. 4.21. Insertion loss of non-homogeneous

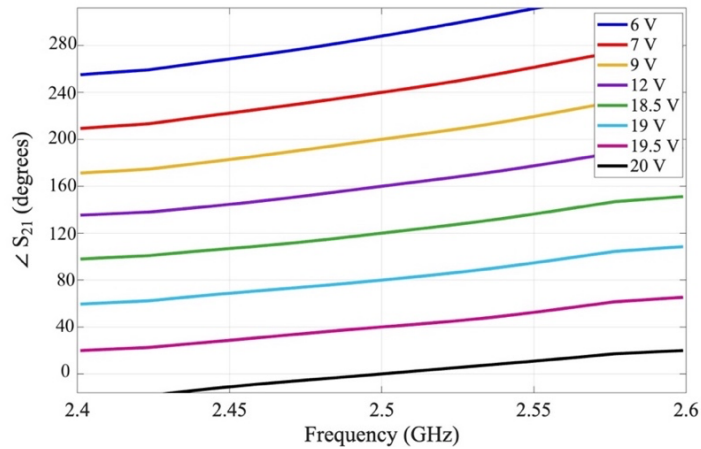


Fig. 4.22. Phase response in case of non-homogeneous medium

The maximum achievable phase shift is approximately 284° for the non-homogeneous case. A comparison among the theory, simulations and measurements is presented in Fig. 4.23. The

measured phase response differs from the simulated and theoretical results by approximately 10° which is quite acceptable.

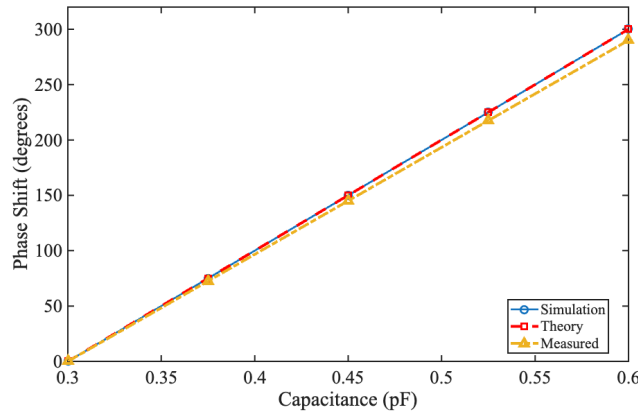


Fig. 4.23. Comparison of phase shift at 2.5 GHz

In the nonhomogeneous configuration, the theoretical model assumes that each section has uniform effective electromagnetic properties and ideal interfaces between adjacent regions. However, the fabricated structure realizes the required nonhomogeneous distribution using discrete varactor-loaded sections. Therefore, the actual capacitance distribution and local electric-field distribution are not perfectly uniform. Moreover, fringing fields and electromagnetic coupling at the interfaces between sections modify the effective propagation constant. Since the total phase shift is accumulated through all nonhomogeneous sections, small phase deviations in individual sections combine and produce the observed difference of approximately 10° . Finally, the FoM for the non-homogeneous case comes out to be $38.7^\circ/\text{dB}$ as compared to the homogeneous value of $66.7^\circ/\text{dB}$.

The figure of merit (FoM) of the phase shifter is defined as the ratio of the maximum phase-shift range to the worst-case insertion loss at the operating frequency. It represents the phase-shifting capability achieved per decibel of transmission loss, where a higher FoM indicates better performance. In some cases, this value is also calculated per unit length. However, the more

prevalent of the two is the former one. This final results clearly shows that the homogeneous case should be preferred over the non-homogeneous not only because it is easy to implement but also for its superior performance.

4.7 Conclusion

This chapter presented the design, analytical modeling, simulation, fabrication, and experimental validation of a reconfigurable phase shifter based on the FPMS concept. Unlike conventional phase shifters that rely on switching transmission-line lengths or impedance loading networks, the proposed design achieves phase tuning by dynamically controlling the effective permittivity of a programmable dielectric waveguide through varactor-loaded unit cells. An analytical model for both homogeneous and non-homogeneous dielectric configurations was developed using guided-wave theory and the transfer matrix method. The analytical predictions showed excellent agreement with full-wave simulations, validating the proposed mathematical framework.

A prototype of the phase shifter was fabricated and experimentally characterized at 2.5 GHz. The measured results demonstrated a maximum continuous phase shift of approximately 298° for the homogeneous configuration and about 284° for the non-homogeneous configuration. The measured phase responses closely agreed with the simulated and theoretical results, exhibiting deviations of approximately $\pm 3^\circ$ – 4° for the homogeneous case and around 10° for the non-homogeneous case. Although the fabricated prototype exhibited higher insertion loss than predicted in simulation due to fabrication tolerances, component losses, and practical implementation constraints, the measured performance successfully verified the proposed operating principle.

A comparison between the homogeneous and non-homogeneous configurations further showed that the homogeneous dielectric configuration provides superior phase-tuning capability and a higher figure of merit while maintaining a simpler biasing strategy. These results demonstrate that the proposed FPMS-based phase shifter provides a compact, continuously tunable, and analytically predictable solution for microwave phase control. The validated analytical model and experimental results establish a strong foundation for integrating the proposed phase shifter with antenna arrays to realize electronically steerable beamforming systems.

Chapter 5

5.1 Conclusions and Future Work

This chapter presents the conclusions drawn from the research conducted in this thesis and also future work. The research focused on investigating FPMS as a versatile reconfigurable EM platform capable of controlling EM wave propagation through manipulation of its effective material properties. By electronically tuning the embedded varactor diodes, the electromagnetic response of FPMS can be dynamically modified, enabling the realization of adaptive microwave components without requiring mechanical movement or complex conventional reconfigurable architectures.

The work presented in this thesis addresses two important areas of reconfigurable microwave engineering: electronically steerable antennas and continuously tunable dielectric phase shifters. Although these applications differ in their operating principles, both exploit the same underlying concept of controlling EM wave propagation through programmable effective permittivity. This demonstrates the versatility of FPMS as a unified platform capable of manipulating both reflected and guided EM waves for adaptive microwave applications. To investigate these capabilities, two complementary microwave structures were developed. The first employed FPMS as a programmable AMC to realize electronic beam steering of a single radiating element through spatial control of the reflection phase. The second utilized FPMS as a programmable dielectric medium to realize a continuously tunable phase shifter by controlling the propagation characteristics of guided waves. For both investigations, analytical models were developed to establish the relationship between the programmable electromagnetic properties of FPMS and the resulting device performance. The proposed analytical formulations were subsequently validated

through full-wave EM simulations and experimental measurements, demonstrating excellent agreement between theoretical predictions and practical results. Overall, the research presented in this thesis demonstrates that FPMS extends beyond the role of a programmable dielectric substrate and can serve as a multifunctional EM platform for the realization of reconfigurable microwave components. The developed analytical models, design methodologies, and experimentally validated prototypes provide a foundation for the future development of compact, adaptive, and electronically reconfigurable RF and microwave systems.

Before presenting the overall summary of the research, this chapter discusses the extension of the proposed FPMS concepts to array configurations and their potential for realizing practical beam steering systems. Following this discussion, the chapter summarizes the research outcomes, evaluates the achievement of the research objectives and finally presents recommendations for future research.

5.2 Summary of Research

The research presented in this thesis was carried out through a systematic investigation of FPMS, beginning with the characterization of its electromagnetic properties and culminating in the development and experimental validation of two reconfigurable microwave components. The overall research methodology combined theoretical analysis, full-wave EM simulations, fabrication, and experimental measurements to establish reliable design procedures for FPMS-based microwave devices. The investigation commenced with the characterization of the FPMS unit cell to understand the influence of varactor capacitance on its effective electromagnetic properties. Effective constitutive parameters were extracted using the NRW technique, providing the material parameters required for analytical modelling and subsequent device design. This

characterization established the relationship between the programmable capacitance and the effective dielectric response of the substrate, forming the basis for all subsequent developments.

Building upon the extracted material characteristics, the first stage of the research investigated the use of FPMS as an AMC for beam steering applications. An analytical model was developed to describe the relationship between the programmable reflection phase and the resulting steering angle, providing an efficient design methodology prior to numerical optimization. The analytical predictions were verified using HFSS simulations, after which the optimized FPMS surface was integrated with a printed antenna. Finally, a prototype was fabricated and experimentally characterized to validate the proposed beam steering concept.

The second stage of the research investigated the guided-wave applications of FPMS through the development of a continuously tunable dielectric phase shifter. Analytical formulations were developed for both homogeneous and non-homogeneous dielectric configurations to predict wave propagation and phase response under different capacitance distributions. These models were subsequently validated through full-wave simulations before fabrication and experimental measurements were performed to verify the predicted performance. Throughout the research, analytical predictions were consistently compared with numerical simulations and experimental measurements to ensure the accuracy of the developed models and the practicality of the proposed designs. The close agreement among theoretical, simulated, and measured results demonstrates the reliability of the proposed methodologies for the design and implementation of FPMS-based microwave components.

Overall, the research establishes a complete design framework encompassing material characterization, analytical modelling, numerical verification, prototype fabrication, and experimental validation. This systematic methodology demonstrates the applicability of FPMS to

both radiating and guided-wave microwave structures and provides a foundation for the development of future reconfigurable microwave devices.

5.3 Achievement of Research Objective

This research successfully achieved all the objectives established in Chapter 1 through a systematic combination of theoretical analysis, analytical modelling, full-wave EM simulations, prototype fabrication, and experimental validation. The accomplishment of each research objective is summarized in Table 5.1, demonstrating how the proposed methodologies collectively fulfilled the overall aim of establishing FPMS as a versatile platform for reconfigurable microwave applications.

Table 5.1 Achievement of the research objectives

Research objective	Achievement
Investigate the electromagnetic behaviour of FPMS under periodic excitation and develop an FPMS-based AMC.	The electromagnetic response of the FPMS unit cell was investigated under periodic excitation. By controlling the embedded varactor capacitance, the reflection phase of the unit cell was varied, enabling the FPMS layer to operate as a programmable AMC. Suitable capacitance states were identified to synthesize the spatial reflection-phase distributions required for beam steering.
Design a gain-enhanced beam-steerable antenna integrated with the FPMS-based AMC.	A printed antenna was integrated with a 5×5 FPMS surface to realize electronic beam steering and gain enhancement. The programmable reflection phase of the FPMS cells was used to redirect the reflected fields, resulting in elevation steering from approximately $+15^\circ$ to $+60^\circ$ and azimuth steering from approximately -25° to $+25^\circ$. The integrated antenna maintained a gain of approximately 7.2 dBi and satisfactory impedance matching over the investigated tuning states.
Fabricate and experimentally validate the FPMS-AMC beam-steering antenna.	The FPMS-AMC antenna was fabricated and experimentally characterized. The measured reflection characteristics, impedance response, gain, and radiation patterns confirmed the programmable AMC behaviour and the resulting beam steering. The measured results showed good agreement with the simulated predictions and validated the practical feasibility of the proposed antenna configuration.

Research objective	Achievement
Design and investigate an FPMS-based dielectric phase shifter.	A continuously tunable dielectric phase shifter was developed by controlling the effective permittivity of the FPMS propagation regions. Both homogeneous and non-homogeneous capacitance configurations were investigated. The proposed structure demonstrated approximately 300° of continuous phase tuning at 2.5 GHz while maintaining good impedance matching. Its applicability was further demonstrated through integration with a printed antenna.
Fabricate and experimentally validate the FPMS-based dielectric phase shifter.	The proposed dielectric phase shifter was fabricated and experimentally measured. Its transmission phase and S-parameters were evaluated under different tuning conditions, and the measured response showed good agreement with the simulated results.
Develop analytical models for the FPMS-AMC beam-steering antenna and the FPMS-based dielectric phase shifter.	Analytical models were developed for both major contributions. For the FPMS-AMC antenna, the model established the relationship between the spatial reflection-phase distribution and the resulting steering angle. For the phase shifter, homogeneous and non-homogeneous models were formulated to predict the propagation constant and transmission phase under different effective-permittivity distributions. The analytical results showed close agreement with HFSS and experimental measurements.

5.4 Limitation of the Present Research

Although the proposed FPMS-based microwave structures demonstrated promising performance through analytical modelling, full-wave simulations, fabrication, and experimental validation, the present research was conducted within a defined scope. Consequently, several aspects remain beyond the objectives of this study and provide opportunities for further investigation.

1). The proposed FPMS-based AMC was investigated using a single radiating element integrated with a programmable surface. Although this configuration successfully demonstrated electronic beam steering and gain enhancement, the behaviour of FPMS in large antenna arrays, where mutual coupling, scan loss, and array-level beamforming become significant, was not investigated.

2). The beam-steering capability of the proposed AMC was achieved using commercially available varactor diodes with a limited capacitance tuning range. Consequently, the achievable steering range and reflection-phase resolution were constrained by the electrical characteristics of the tuning components. The investigation of alternative tuning technologies, wider tuning ranges, and improved biasing techniques was beyond the scope of this work.

3). The proposed dielectric phase shifter was designed and experimentally validated at 2.5 GHz. Although the developed analytical models are generally applicable, the practical implementation and performance of the proposed structure at higher microwave and millimetre-wave frequencies were not investigated in this thesis.

4). Furthermore, the analytical models developed for both the FPMS-AMC beam steering antenna and the dielectric phase shifter were derived under several simplifying assumptions to obtain closed-form design equations. While the proposed models showed excellent agreement with full-wave simulations and experimental measurements, additional effects such as fabrication tolerances, parasitic elements, and complex electromagnetic interactions may require further consideration for highly integrated or large-scale implementations.

Despite these limitations, the research successfully achieved its intended objectives by establishing FPMS as a versatile platform for reconfigurable microwave applications and by experimentally validating both proposed structures. The identified limitations do not diminish the significance of the presented work; rather, they define a clear direction for future research aimed at extending the capabilities and practical implementation of FPMS-based microwave systems.

5.5 Future Research Direction

5.5.1 Application of the FPMS Phase Shifter Design

As a future extension of the proposed FPMS-based phase shifter, the design can be integrated with a two-element antenna array to investigate beam steering which is shown in Fig. 5.1. As illustrated in Fig. 5.2, each antenna element can be excited through an independent FPMS phase shifter. The capacitance of one phase shifter can be maintained at a fixed value, while the capacitance of the other can be varied from 0.25 pF to 0.6 pF. This variation will introduce a controllable phase difference between the two antenna elements and consequently steer the main radiation beam. The beam can be directed toward either the positive or negative angular region by interchanging the fixed and variable capacitance states between the two-phase shifters.

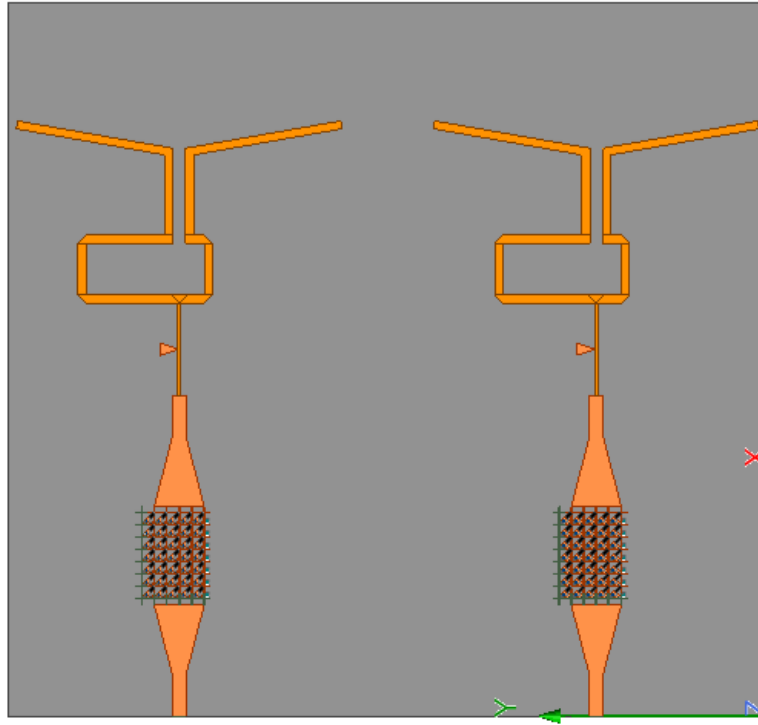
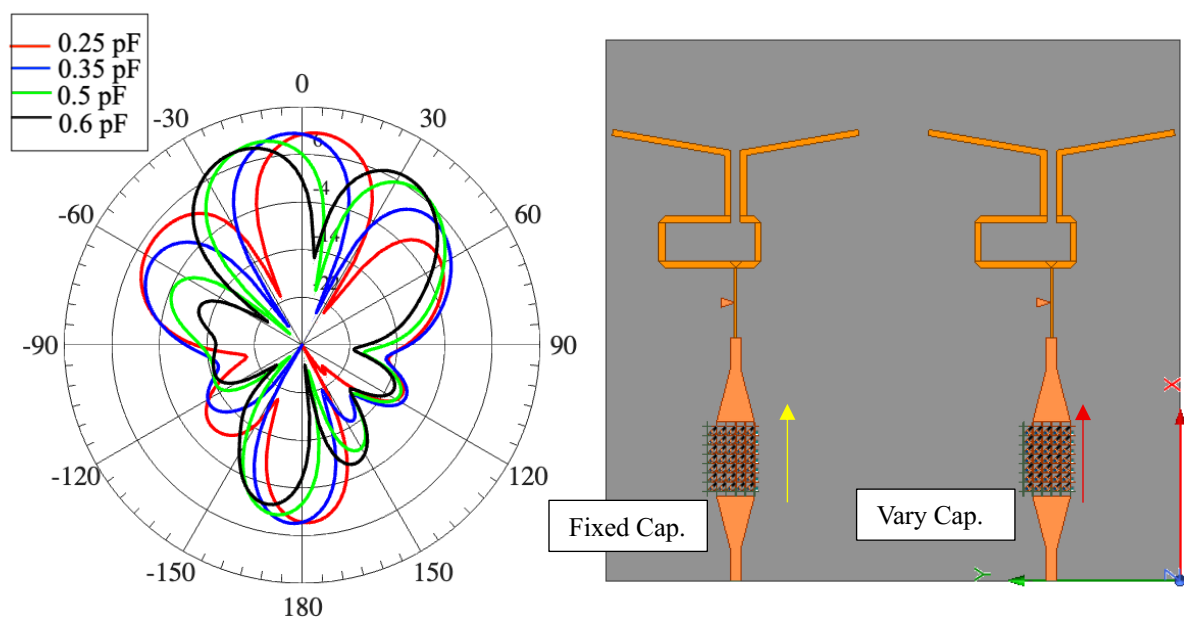
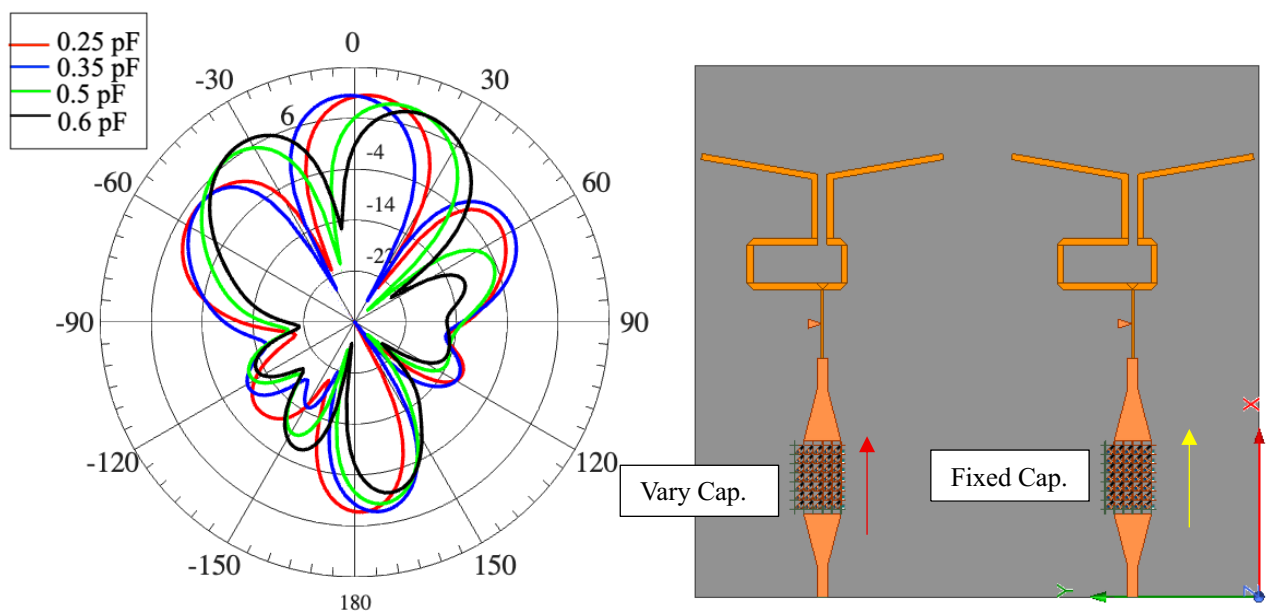


Fig. 5.1 Array setup of antennas



(a)



(b)

Fig. 5.2 (a) Beam steering in negative direction (b) Beam steering in positive direction

The research presented in this thesis establishes FPMS as a promising platform for reconfigurable microwave applications by demonstrating its capability in both radiating and guided-wave structures. Although the proposed concepts have been successfully validated through analytical modelling, full-wave simulations, fabrication, and experimental measurements, several opportunities remain to further enhance the performance, scalability, and practical implementation of FPMS-based microwave devices. Some potential directions for future research are discussed below. One important direction for future work is the extension of the proposed FPMS-based AMC from a single radiating element to larger antenna arrays. While the present study successfully demonstrated electronic beam steering using a single antenna integrated with a programmable AMC surface, practical communication and radar systems generally require planar antenna arrays to achieve higher gain, improved beamforming capability, and wider scanning coverage. Future investigations should therefore examine the behaviour of FPMS in array environments, including mutual coupling effects, scan loss, and the optimization of phase distributions for two-dimensional beam steering.

Another promising research direction is the extension of the proposed FPMS structures to higher microwave and millimetre-wave frequency bands. Although the dielectric phase shifter and AMC were validated at their respective operating frequencies, future wireless communication and sensing systems are increasingly operating at Ku-, Ka-, and millimetre-wave frequencies. Investigating the EM behaviour, fabrication challenges, and performance limitations of FPMS at higher frequencies would further broaden its applicability to emerging technologies such as 6G communications, automotive radar, and high-capacity satellite systems. The tuning mechanism employed in this research was based on commercially available varactor diodes. While these devices provided continuous electronic reconfigurability, future research may investigate

alternative tuning technologies, including RF MEMS switches, ferroelectric materials, liquid crystals, phase-change materials, or graphene-based components. These technologies may provide improved tuning range, lower insertion loss, higher linearity, faster switching speed, and enhanced power-handling capability, thereby expanding the operational performance of FPMS-based microwave structures.

The analytical models developed in this thesis provide efficient design methodologies for both the FPMS-based AMC beam-steering antenna and the dielectric phase shifter. Future work may further extend these models by incorporating additional practical factors such as mutual coupling, fabrication tolerances, parasitic effects, conductor losses, and more complex electromagnetic interactions. Such improvements would increase the accuracy of the analytical predictions and facilitate the design of larger and more sophisticated FPMS-based microwave systems. Another interesting direction involves the integration of FPMS with intelligent control algorithms. Future programmable microwave systems may combine FPMS with FPGA- or microcontroller-based control platforms and artificial intelligence techniques to automatically determine the optimum capacitance distributions for dynamic beam steering or adaptive phase control. Such intelligent reconfigurable systems could continuously optimize their performance in response to changing operating conditions and electromagnetic environments, enabling autonomous and adaptive microwave front ends.

Finally, the concepts developed in this thesis may be extended to a wide range of practical microwave applications. The demonstrated capability of FPMS to realize electronically reconfigurable beam steering and continuous phase control makes it an attractive candidate for phased-array radar systems, CubeSat and satellite communication payloads, adaptive wireless communication systems, electronically steerable antennas, and future reconfigurable RF front

ends. Further development of these technologies may contribute to compact, low-cost, and highly adaptive microwave systems capable of meeting the increasing demands of next-generation communication and sensing applications.

5.6 Published Work

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- [7] S. Ali and F.A Ghaffar 'A Versatile FPMS Based Phase Shifter Design' is submitted for IEEE, MWTL.
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