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DISSERTATION

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Metamaterial-Based Electronically Tunable Filters: Design and Optimization

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Abstract

This dissertation focuses on addressing the growing need for compact and frequency-agile microwave components through the design, optimization, and implementation of innovative metamaterial unit cells, leading to the development of an electronically reconfigurable microstrip stopband filter. The proposed approach is based on a novel unit cell inspired by the Split-Ring Resonator (SRR) concept, redesigned to achieve increased electrical length while ensuring high compactness. By incorporating PIN diodes within the structure, dynamic control over the filter's operating frequency and response is achieved, enabling reconfigurability without compromising performance or size. The effectiveness of the proposed filter is validated by its capability to selectively suppress multiple frequency bands, specifically 3.6 GHz, 5.2 GHz, and 5.8 GHz, through different diode switching states. This versatility highlights its suitability for integration into contemporary wireless communication systems, including WLAN, ISM-band, IoT, and short-range communication applications.

Keywords: stopband filter, microstrip, metamaterial, SRR, reconfigurability, PIN diodes.

ملخص

تتناول هذه الأطروحة الحاجة المتزايدة إلى مكونات ميكروويف مدمجة وقابلة لإعادة الضبط من حيث التردد، من خلال تصميم وتحسين وتنفيذ خلايا ميتاماتيريال مبتكرة، مما أدى إلى تطوير مرشح إيقاف نطاق (Stopband) ميكروشرطي قابل لإعادة التشكيل إلكترونياً. يعتمد هذا العمل على خلية جديدة مستوحاة من مفهوم الرنان الحلقي المشقوق (SRR)، حيث تم إعادة تصميمها لتحقيق طول كهربائي أكبر مع الحفاظ على درجة عالية من التصغير. ومن خلال دمج ثنائيات PIN داخل البنية، تم تحقيق تحكم ديناميكي في تردد التشغيل واستجابة المرشح، مما يتيح خاصية إعادة التشكيل دون التأثير على الأداء أو الحجم. وقد تم التحقق من فعالية المرشح المقترح من خلال قدرته على حجب عدة نطاقات ترددية بشكل انتقائي، وهي 3.6 جيجاهرتز و 5.2 جيجاهرتز و 5.8 جيجاهرتز، وذلك عبر حالات تبديل مختلفة للثنائيات. وتبرز هذه المرونة إمكاناته الكبيرة للاستخدام في أنظمة الاتصالات اللاسلكية الحديثة، بما في ذلك شبكات WLAN ونطاقات ISM وتطبيقات إنترنت الأشياء والاتصالات قصيرة المدى.

كلمات مفتاحية: بمرشح إيقاف نطاق، شريط دقيق، المواد الميتا، SRR، إعادة التشكيل، الصمام الثنائي PIN.

Résumé

Cette thèse répond au besoin croissant de composants micro-ondes compacts et accordables en fréquence, à travers la conception, l'optimisation et la mise en œuvre de cellules métamatériaux innovantes, aboutissant au développement d'un filtre coupe-bande micro-ruban reconfigurable électroniquement. L'approche proposée repose sur une cellule unitaire originale inspirée du concept de résonateur à anneau fendu (SRR), repensée afin d'augmenter la longueur électrique tout en conservant une grande compacité. L'intégration de diodes PIN au sein de la structure permet un contrôle dynamique de la fréquence de fonctionnement et de la réponse du filtre, assurant ainsi la reconfigurabilité sans dégradation des performances ni augmentation significative de la taille. L'efficacité du filtre est démontrée par sa capacité à rejeter sélectivement plusieurs bandes de fréquences, notamment 3,6 GHz, 5,2 GHz et 5,8 GHz, grâce à différentes configurations de commutation des diodes. Cette flexibilité met en évidence son fort potentiel d'intégration dans les systèmes de communication sans fil modernes, tels que les réseaux WLAN, les bandes ISM, ainsi que les applications IoT et les communications à courte portée.

Mots clés : filtres coupe-bande, microruban, Métamatériau, SRR, reconfigurabilité, diodes PIN.

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List of Abbreviations

4G: Fourth-Generation

5G: Fifth-Generation

6G: Sixth-Generation

BPF: Band-Pass Filter

BW: Bandwidth

CAD: Computer-Aided Design

CPW: Coplanar Waveguide

CRLH: Composite Right/Left-Handed (Transmission Line)

CSRR: Complementary Split-Ring resonators

CST: Microwave Studio

dB: Decibel

DNG: Double Negative

DPS: Double-Positive

ECM: Electronic Countermeasures

EM: Electromagnetic

ENG: Epsilon-Negative

FH: High cutoff frequency

FL: Low cutoff frequency

GCPW: Grounded Coplanar Waveguide

GHz: Gigahertz

GPS: Global Positioning System

HFSS: High-Frequency Structure Simulator

IoT: Internet of Things

IR: Infrared

LC: Inductance-Capacitance

LH: Left-Handed (Materials/Transmission Line)

LHM: Left-Handed Materials

MEMS: Micro-Electro-Mechanical Systems

MHz: Megahertz

MMIC: Monolithic Microwave Integrated Circuits

MMT: Metamaterials

mmWave: Millimeter-Wave

MNG: Mu-Negative

MRI: Magnetic Resonance Imaging

NRW: Nicolson-Ross-Weir

NWB BPF: Notched-Wideband Bandpass Filter

ON/OFF: Diode Biasing States

PIN: Positive-Intrinsic-Negative (Diode)

QA: Quality Factor (Loaded/Unloaded)

Quasi-TEM: Quasi-Transverse Electromagnetic

RC: Resistor-Capacitor

RF: Radio Frequency

RH: Right-Handed (Materials/Transmission Line)

RHM: Right-Handed materials

RHTL: Right-Handed Transmission Line

RLC: Resistor-Inductor-Capacitor

SBF: Stopband Filter

SI: Système International

SNR: Signal-to-Noise Ratio

S-Parameters: Scattering Parameters

SRR: Split-Ring Resonators

TE: Transverse Electric

TEM: Transverse Electromagnetic

THz: Terahertz

TL: Transmission line

TM: Transverse Magnetic

TR: Transition radiation

TZ: Transmission Zero

UV: Ultraviolet

WIFI: Wireless Fidelity

x-ray: X –radiation

GENERAL INTRODUCTION:

General introduction

The microwave frequency spectrum, spanning from 300 MHz to 300 GHz, underpins a vast array of modern technological infrastructures, including wireless communications, navigation networks, and advanced radar systems. The successful deployment of these technologies relies heavily on components capable of efficiently guiding and manipulating electromagnetic waves. Consequently, planar guiding structures—such as microstrip lines and coplanar waveguides (CPW)—are extensively utilized in contemporary designs due to their low fabrication costs, compact form factors, and seamless integrability into broader circuit architectures [1].

Within the complex architecture of microwave circuits, filters function as indispensable two-port networks. As the operational electromagnetic spectrum becomes increasingly congested, filters play a critical role in signal isolation and management. They achieve this by selectively facilitating the transmission of desired frequency bands (passbands) while rigorously attenuating unwanted interference (stopbands), thereby ensuring high-fidelity signal processing [2].

Simultaneously, the relentless industry drive toward device miniaturization and enhanced performance has catalyzed extensive research into metamaterials. These artificially engineered structures are designed to exhibit unique electromagnetic characteristics rarely found in natural materials, such as negative permittivity and negative permeability [3]. Since their foundational practical demonstrations utilizing sub-wavelength elements like split-ring resonators (SRRs) [4, 5], metamaterials have provided a robust framework for developing radio frequency (RF) components that are substantially smaller than conventional alternatives operating at equivalent frequencies.

Beyond physical miniaturization, the rapid evolution of complex RF systems increasingly dictates a need for operational agility. Reconfigurable filters, which are capable of dynamically modulating their frequency responses, are essential for the realization of multi-functional devices. Such adaptability is crucial for augmenting radar capabilities in electronic countermeasures (ECM) and for facilitating secure communication protocols, including frequency hopping [6]. This dynamic electronic tuning is typically realized by embedding active semiconductor components, such as varactor or PIN diodes, directly into the planar filter topologies. Integrating these active elements with metamaterial-inspired structures has emerged as a highly effective methodology for shifting resonant frequencies or entirely

altering filter functions. This approach successfully merges the advantages of ultra-compact physical footprints with rapid, electronically controlled system adaptability [7].

This dissertation directly responds to these converging demands for extreme compactness and dynamic versatility by presenting the design, theoretical analysis, and implementation of a novel, miniaturized stopband filter endowed with both frequency and mode reconfigurability. The cornerstone of this innovation is a newly synthesized metamaterial unit cell. While rooted in the traditional SRR concept, this proposed cell is explicitly engineered to maximize electrical length, thereby achieving superior compactness compared to conventional designs. This optimized building block forms the foundation of the multi-resonator stopband filter architecture, which integrates strategically positioned PIN diodes. Functioning as high-speed electronic switches, these diodes govern the dynamic control of the filter's operational modes and resonant characteristics.

To systematically present this research, the dissertation is organized into the following chapters:

- **Chapter 1** establishes the theoretical foundation of metamaterials. It reviews their historical progression, fundamental physical concepts, transmission line theories (including Right-Handed, Left-Handed, and Composite Right/Left-Handed[CRLH] media), and highlights prominent applications within the field.
- **Chapter 2** provides the requisite background on microwave engineering principles. It covers the foundational theory of microwave filters, examines the guiding technologies employed in their physical design, and presents a comprehensive survey of contemporary methodologies used to realize reconfigurable filter architectures.
- **Chapter 3** details the core technical contributions of this work. It encompasses the rigorous design process, electromagnetic simulation, analytical modeling, and final performance characterization of both the proposed novel metamaterial unit cell and the resulting PIN-diode-tuned reconfigurable stopband filter.

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CHAPTER I:

History, Fundamental Concepts and Applications of Metamaterials

I.1. Introduction

Metamaterials (MMTs) exhibit exceptional properties that often seem futuristic, enabling innovative applications such as microwave filters, invisibility cloaks, and advanced smart antennas. What distinguishes these materials is their engineered structure, which allows them to challenge and extend conventional electromagnetic principles. As their name suggests, metamaterials are artificially designed to display physical properties not typically found in natural materials. They go beyond the inherent limitations of conventional substances, providing new possibilities in electromagnetics and optics. Among their most notable characteristics are the negative refractive index and the reversed Doppler effect, both of which have attracted significant interest in the research community.

This chapter presents a comprehensive overview of metamaterials, establishing the theoretical foundation necessary for understanding and applying the structures discussed in later chapters. It begins with an introduction to the fundamental concepts of metamaterials, including their defining features such as negative refractive index and unusual electromagnetic responses. The discussion then moves to metamaterial transmission line theory, emphasizing the differences between right-handed (RH), left-handed (LH), and composite right/left-handed (CRLH) transmission lines. The chapter concludes with a review of key applications, illustrating the wide-ranging impact and potential of metamaterials across various technological domains.

I.2. Origins of Metamaterials

The notion of artificial materials possessing customized electromagnetic characteristics extends back to the late 19th century. Among the earliest recorded initiatives in this domain was undertaken by J.C. Bose in 1898, who executed groundbreaking experiments investigating microwave polarization. Bose illustrated that electromagnetic wave polarization could be rotated utilizing twisted synthetic fibers submerged in a liquid medium, specifically sugar syrup [I.1]. This pioneering work represented the initial advancement toward comprehending and controlling wave propagation within artificial structures.

During the early 20th century, additional theoretical progress was accomplished by H. Lamb and H.C. Pocklington (1904-1905), who investigated the characteristics of phase velocity and group velocity within mechanical systems. They established that, under particular conditions,

these velocities could propagate in opposing directions, contingent upon the arrangement of charged chains [I.2, I.3]. This discovery established the foundation for comprehending wave dynamics within structured materials.

In 1914, Lindman built upon these concepts by examining chiral artificial media, comprising randomly oriented spiral threads incorporated within a host medium [I.4]. His research yielded early understanding into the design of materials exhibiting distinctive electromagnetic responses, especially regarding wave polarization and propagation.

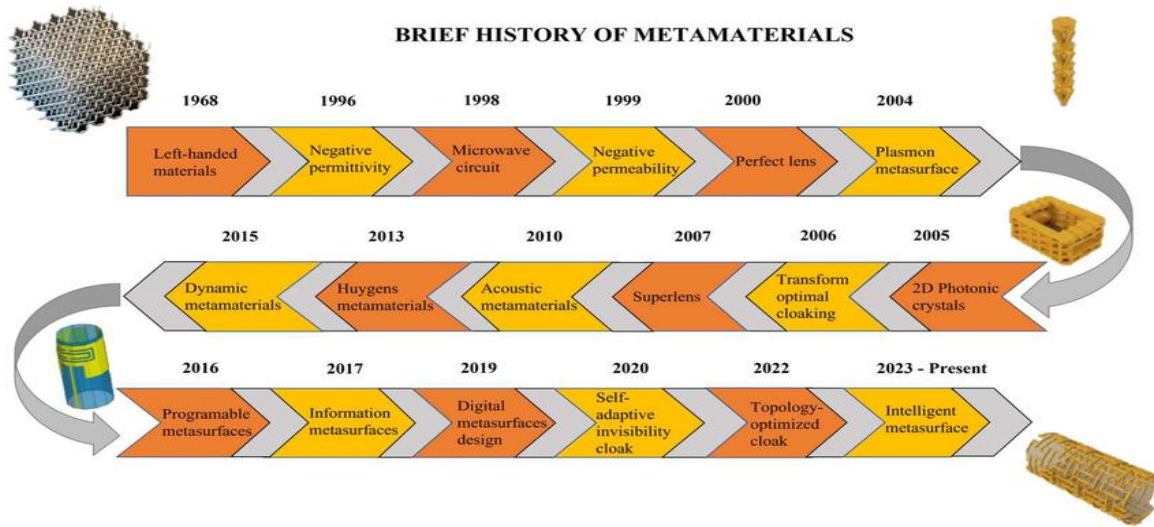


Figure I-1: Evolution of metamaterials and metasurfaces: from fundamental discoveries to intelligent 6G applications.

This figure (I-1) illustrates the progression of metamaterials and metasurfaces throughout the years. The development initiated with the identification of unconventional materials possessing unusual properties such as negative permittivity, which created new opportunities in physics [I.5]. With time, these theoretical concepts materialized into practical applications, including perfect lenses and optical cloaking technologies [I.6]. Subsequently, researchers created increasingly sophisticated metasurfaces capable of accurately controlling electromagnetic waves. Currently, attention has transitioned toward integrating these surfaces with artificial intelligence to satisfy the requirements of 6G networks, signaling a genuine revolution in intelligent communications.

A substantial achievement in this field occurred in 1944 when L.I. Mandelshtam performed the initial theoretical investigation on negative refraction. His research presented the concept of negative refractive index and featured the first diagram depicting this phenomenon, as

displayed in Figure (I-2) [I.7]. Mandelshtam's contributions proved crucial in establishing the contemporary understanding of metamaterials and their prospective applications.

These preliminary investigations collectively established the groundwork for the advancement of metamaterials, which have subsequently transformed domains including optics, electromagnetics, and material science.

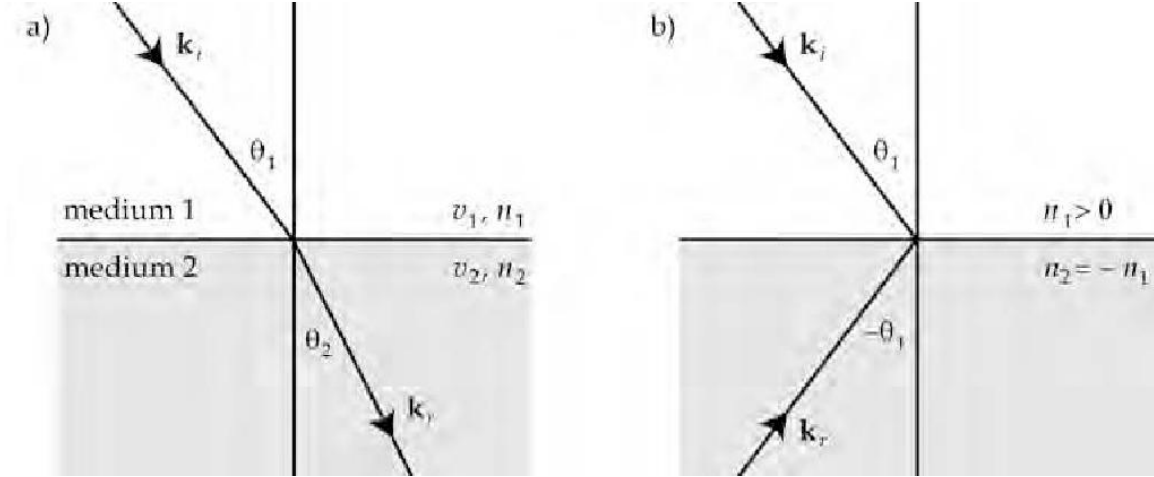


Figure I-2: L.I. Mandelshtam presentation of the first schematic diagram of negative refraction in 1950.

According to L.I. Mandelshtam, negative refraction constitutes a phenomenon emerging from the opposition between phase velocity and group velocity, as depicted in Figure I-2. Mandelshtam contended that, despite its apparently counterintuitive character, this phenomenon remains physically valid since the phase itself does not transport energy [I.7]. This understanding established the foundation for comprehending the distinctive behavior of waves within materials possessing unconventional electromagnetic properties.

The concept of artificial dielectrics was initially proposed by W.E. Kock in 1948. Kock introduced the notion of fabricating lightweight dielectric lenses by incorporating conductive spheres, disks, and ribbons within a regular matrix, consequently controlling the effective refractive index of the artificial medium [I.8]. This methodology represented a considerable advancement toward engineering materials possessing customized electromagnetic responses.

In 1951, Malyuzhinets furnished experimental verification for one-dimensional (1D) infinite periodic structures supporting "backward" waves. These structures comprised series

capacitors linked in parallel with inductors, demonstrating the practicability of designing materials exhibiting unconventional wave propagation characteristics [I.9]. Shortly thereafter, in 1957, Sivukhin investigated the correlation between negative permittivity (ϵ) and permeability (μ) and their function in facilitating negative refraction, further progressing the theoretical comprehension of these phenomena [I.10].

In 1959, Pafomov released a comprehensive study examining the Doppler effect and Vavilov-Cherenkov radiation within media possessing negative ϵ and μ [I.11]. His research provided enhanced understanding into the behavior of electromagnetic waves in such unconventional materials. Expanding upon these foundations, Russian physicist Victor G. Veselago performed an extensive theoretical investigation in 1968 examining the electromagnetic properties of materials exhibiting simultaneously negative ϵ and μ . Veselago's research established the essential principles of wave propagation within these hypothetical media, as illustrated in Figure I-3, and created the theoretical foundation for the field of metamaterials [I.5].

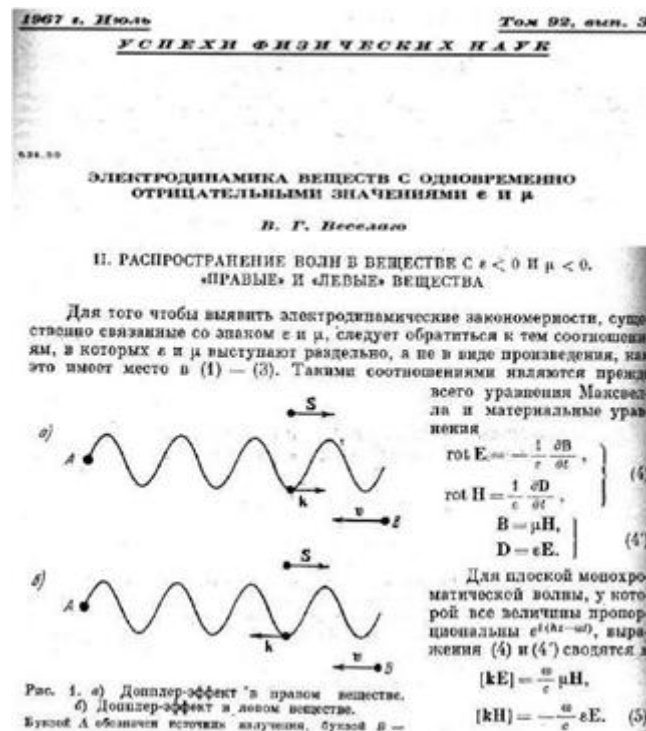


Figure I-3: Professor V.G.Veselago's theory of left-hand propagation.

I.3. Early Theoretical Foundations and the First Realization of Metamaterials

Until the 1960s, research on metamaterials remained entirely theoretical. Although their underlying principles were well established, no practical implementation had yet been realized. This situation evolved when J. Brown and W. Rotman conducted studies on wire networks during the 1960s, leading to the development of RF lenses based on artificial dielectric materials [I.12].

At roughly the same period, S. Schelkunoff investigated resonant magnetic loops, laying the groundwork for structures that were later refined and widely applied throughout the 1980s and 1990s [I.13].

I.3.1. The First Realization of Metamaterials

It was established that electromagnetic wave propagation occurs only in media that are either double-positive (both permittivity and permeability are positive) or double-negative (both are negative). In contrast, materials with only one negative parameter exhibit evanescent wave behavior, preventing propagation. This finding motivated extensive research aimed at developing double-negative metamaterials capable of operating across a wide portion of the electromagnetic spectrum, from microwave to optical frequencies.

In 1999, J. Pendry proposed the concept of a double resonant ring structure, which is widely regarded as the basis for the first practical realization of metamaterials [I.14]. Shortly afterward, R. A. Shelby experimentally demonstrated the first material exhibiting negative electromagnetic properties [I.15]. Although the individual components used in this realization were already known, their combination into a single functional structure had not previously been attempted.

The concept of the superlens, introduced by J. B. Pendry in 2000, became a major driving force in metamaterial research [I.6]. This concept aimed to achieve imaging beyond the diffraction limit, enabling resolution finer than the wavelength of light. However, practical implementation faced significant challenges, particularly due to losses and the difficulty of achieving isotropic behavior in double-negative media. These limitations led researchers to explore alternative approaches based on simpler negative-index structures.

In the same year, D. R. Smith demonstrated that a composite structure combining split-ring resonators (SRRs) with continuous metallic wires could simultaneously exhibit negative permittivity and permeability, resulting in a negative refractive index [I.16]. This configuration effectively realized a Veselago-type medium through a periodic arrangement of these elements, with negative refraction occurring near the resonant frequency of the SRRs. Efforts were also made to extend these concepts to higher frequencies, including the infrared and visible ranges, which required nanometer-scale fabrication. For example, in the visible spectrum (around 500 nm wavelength), structural periodicity needed to be on the order of 50 nm, with metallic features as small as 10 nm.

More recently, advances in fabrication and synthesis techniques have enabled the creation of complex structures and composite materials that either mimic known electromagnetic responses or exhibit entirely new properties not found in nature. In 2002, C. Caloz, T. Itoh, and G. V. Eleftheriades introduced a new class of non-resonant metamaterials, offering reduced losses compared to resonant designs (see Figure I-4) [I.17].

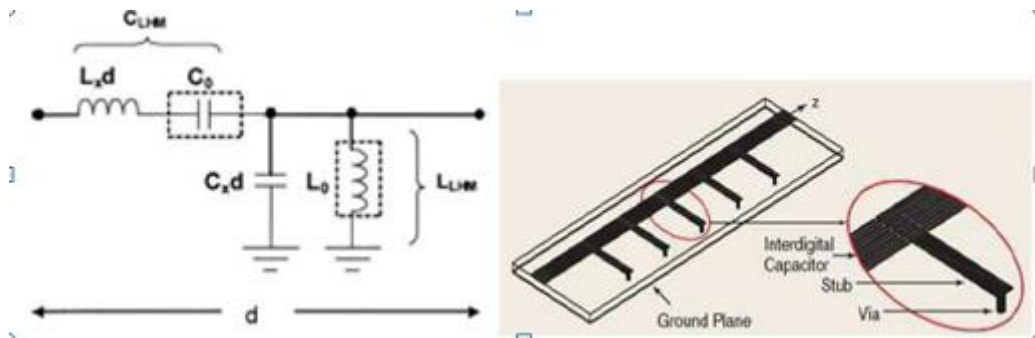


Figure I-4: The non-resonant material structure in the left-hand developed by of C. Caloz .

I.4. Categorization of Materials

Within the microwave domain, materials are classified into four categories based on the sign of their permittivity and permeability, as illustrated in Figure I-5 [I.18].

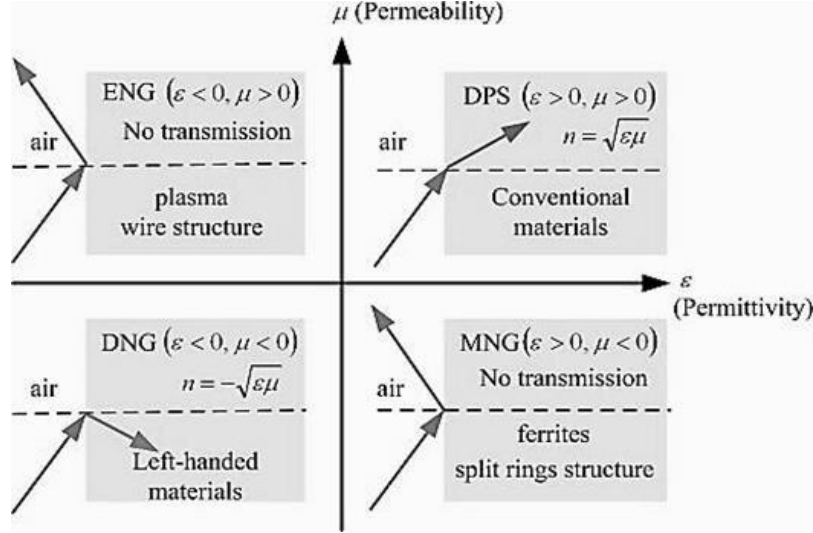


Figure I-5: Permittivity and permeability signs are used to classify materials.

The initial three materials are not completely original. Specifically, right-handed materials (RHM), such as isotropic dielectric materials, correspond to the zone (+, +) where the media are doubly positive (DPS) [I.5]. Negative permittivity (ENG, (-, +) and permeability (MNG, (+, -) media have also been documented [I.16, I.19].

I.5. Foundations and Classification of Metamaterials

The fundamental concepts behind metamaterials have long been rooted in classical electromagnetism. The Drude–Lorentz model, which describes the response of most natural materials, indicates that permittivity becomes negative when operating below the plasma frequency [I.20]. In a similar manner, certain ferrimagnetic materials exhibit regions of negative permeability due to strong magnetic interactions [I.21].

Based on their electromagnetic properties, natural materials can generally be classified into three categories: double-positive (DPS), epsilon-negative (ENG), and mu-negative (MNG) media. However, materials exhibiting both negative permittivity and negative permeability simultaneously—known as doubly negative (DNG) media—do not occur naturally. These materials, often referred to as left-handed materials (LHM), form the fundamental basis of metamaterial research [I.5].

In the literature, multiple terms are used interchangeably to describe left-handed materials, each emphasizing a specific feature or behavior of these engineered structures [I.22, I.19]:

- **Left-handed materials (LHM):** This term emphasizes the distinctive property of metamaterials where the phase velocity and group velocity propagate in opposing directions.
- **Negative refractive index materials:** These materials display a negative effective refractive index for electromagnetic waves within a particular frequency range.
- **Double negative materials:** This designation refers to the simultaneous negativity of permittivity and permeability within the material.
- **Veselago medium:** Named after Victor Veselago, who initially theorized the concept of materials with negative refractive index, this term recognizes his foundational work.
- **Backward wave materials:** This term emphasizes the backward propagation of waves in such materials.

I.6. Effective Homogeneous Case

Metamaterials are frequently characterized as pseudo-homogeneous artificial materials possessing electromagnetic properties absent in nature. Nevertheless, this definition has generated debate, particularly concerning the terms "artificial" and "not available in nature," since numerous modern materials are artificially engineered, and their properties may not occur naturally. A. Shivola appropriately summarizes the challenges in defining metamaterials, emphasizing the necessity for a clear and precise description [I.23].

The term "pseudo-homogeneous" indicates that the unit cell size (P) of the material must be substantially smaller than the guided wavelength λ_g at the operating frequency. This guarantees that the material appears homogeneous to the propagating wave, minimizing diffraction effects. A widely accepted criterion for homogeneity is $P < \lambda_g/4$; this limit ensures that diffraction phenomena can be disregarded. In other words, the dimension of the basic unit cell " P " must be smaller than the guided wavelength λ_g in pseudo-homogeneous material over a frequency band (see Figure I-6) [I.18].

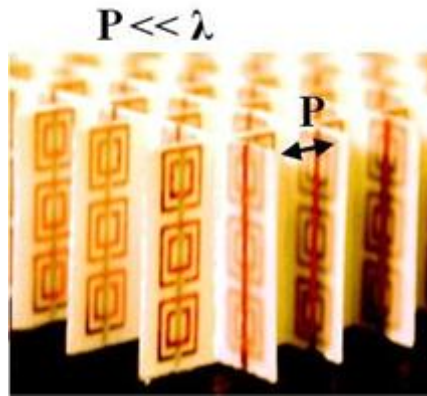


Figure I-6: From article [16], a pseudo homogeneous structure of a periodic network resonating rings and wires.

When a pencil is partially inserted into a glass of water, the positive refractive index generates a "folded" image [I.21]. Light or an electromagnetic wave at the appropriate frequency will be compelled to deflect in the opposite direction by a "metaliqid" [I.15]. While anticipating its experimental synthesis, one can visualize a pencil submerged in such a liquid (Figure I-7).



Figure I-7: medium with a normal refractive index vs a liquid metamaterial with a negative refractive index.

The assertion "electromagnetic properties not found in nature" can be somewhat misleading [I.5]. It is feasible to envision numerous novel structures being classified as "metamaterials" based exclusively on this criterion. Since metamaterials are pseudo-homogeneous, it is sufficient to characterize their magnetic permeability (μ) and electrical permittivity (ϵ). The signs of these two parameters, which constitute the refractive index (n), can subsequently be utilized to classify conventional materials. A fundamental characteristic that has garnered considerable attention for metamaterials is the simultaneous presence of negative permeability and permittivity. This property is defined by the following relationship for the refractive index [I.5]:

$$n = \sqrt{\epsilon_r \mu_r} \quad (I.1)$$

When both μ and ϵ are negative, the refractive index becomes negative, a phenomenon not observed in naturally occurring materials and a defining characteristic of metamaterials.

I.7. Negative Index Materials

The refractive index, designated as " n ", constitutes a fundamental property that characterizes materials [I.21]. According to Snell-Descartes' laws, the refractive index determines the angles of reflection and refraction of an incident light ray [I.19]. In conventional materials, the refractive index is typically greater than or equal to 1 [I.18]. However, in metamaterials, achieving a refractive index less than 1, or even 0, is possible [I.21]. This distinctive property implies that the transmitted ray may appear on the same side of the normal as the incident ray, a phenomenon not observed in natural materials [I.19].

The refractive index (n) is intrinsically connected to two key material parameters: the magnetic permeability (μ) and the electric permittivity (ϵ) [I.18]. These parameters are related to the refractive index through the following equation:

$$n = \sqrt{(\mu \cdot \epsilon)} \quad (I.2)$$

In most natural materials, the relative permittivity ϵ_r and relative permeability μ_r are complex quantities due to inherent losses, with their real parts being positive. Consequently, the sign of the refractive index is typically unambiguous. However, in metamaterials, where both the real parts of ϵ and μ can be negative simultaneously, the situation becomes more

nuanced. Specifically, when $(\epsilon_r) < 0$ and $(\mu_r) < 0$, the refractive index (n) can assume a negative value, meaning ($n < 0$).

To better comprehend this, we can represent ϵ and μ in the complex plane, as observed in Figure I-8. Let:

$$\epsilon_r = |\epsilon_r| \cdot e^{i\varphi_\epsilon} \quad \text{and} \quad \mu_r = |\mu_r| \cdot e^{i\varphi_\mu} \quad (\text{I.3})$$

Here, φ_ϵ and φ_μ represent the phase angles of ϵ_r and μ_r , respectively. The refractive index can then be expressed as:

$$n = \sqrt{(|\epsilon_r| \cdot |\mu_r|) \cdot e^{\frac{1}{2}i(\varphi_\mu + \varphi_\epsilon)}} \quad (\text{I.4})$$

For passive media, the imaginary parts of ϵ_r and μ_r must be positive, which constrains the phase angles φ_ϵ and φ_μ to the range $[0, \pi]$. Consequently, the phase angle of the refractive index " n " also lies within the same range, $[0, \pi]$.

In the case of a doubly negative medium, where both $(\epsilon) < 0$ and $(\mu) < 0$, the phase angles φ_ϵ and φ_μ are restricted to the range $[\pi/2, \pi]$. This results in a negative real part for the refractive index " n ", as the combined phase angle $(\varphi_\epsilon + \varphi_\mu)/2$ falls within $[\pi/2, \pi]$.

To illustrate this, consider the following example where both ϵ and μ are negative:

$$\epsilon_r(\omega) = |\epsilon_r| \cdot e^{-i\pi} \quad (\text{I.5})$$

$$\mu_r(\omega) = |\mu_r| \cdot e^{-i\pi} \quad (\text{I.6})$$

Substituting these into the expression for the refractive index yields:

$$n = \sqrt{|\epsilon_r| \cdot |\mu_r|} e^{-i\pi} = -\sqrt{(|\epsilon_r| \cdot |\mu_r|)} \quad (\text{I.7})$$

This clearly demonstrates that when both ϵ and μ are negative, the refractive index " n " becomes negative. This distinctive property of negative index materials is a hallmark of metamaterials and has substantial implications for their applications in optics and electromagnetics.

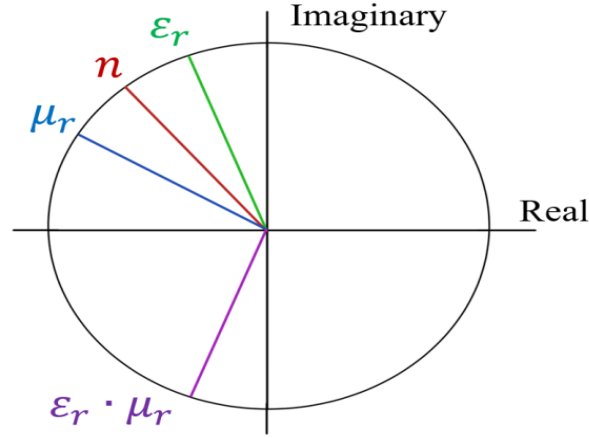


Figure I-8: The arguments of ϵ_r , and μ_r , as well as their product, are represented in the complex plane.

I.7.1. Negative Refractive Index in Left-Handed Materials (LHM)

It is apparent from the preceding discussion that the refractive index of Left-Handed Materials (LHM) is negative relative to that of a vacuum [I.5, I.6]. This distinctive property results in unconventional electromagnetic wave behavior, as depicted in Figure I-9 (a). When an electromagnetic (EM) wave interacts with a negative index medium, it refracts at a negative angle relative to the normal, contrary to the positive refraction observed in conventional materials [I.5, I.15].

I.7.2. Reversal of Lens Functionality

In metamaterials with a negative refractive index, the conventional behavior of optical lenses is reversed [I.6]. A convex lens, which normally focuses light rays in standard materials, instead causes them to diverge. In contrast, a concave lens, typically associated with divergence, produces a converging effect in such media. This inversion of optical behavior is illustrated in Figures I-9 (b) and I-9 (c) [I.5, I.6].

I.7.3. Applications in Superlenses

This phenomenon has substantial implications for overcoming the limitations of traditional optical systems. By leveraging the properties of negative index materials, researchers have developed advanced structures known as superlenses. These superlenses are capable of achieving resolutions beyond the diffraction limit, creating new possibilities in fields such as nanophotonics, microscopy, and medical imaging [I.22, I.24, I.25].

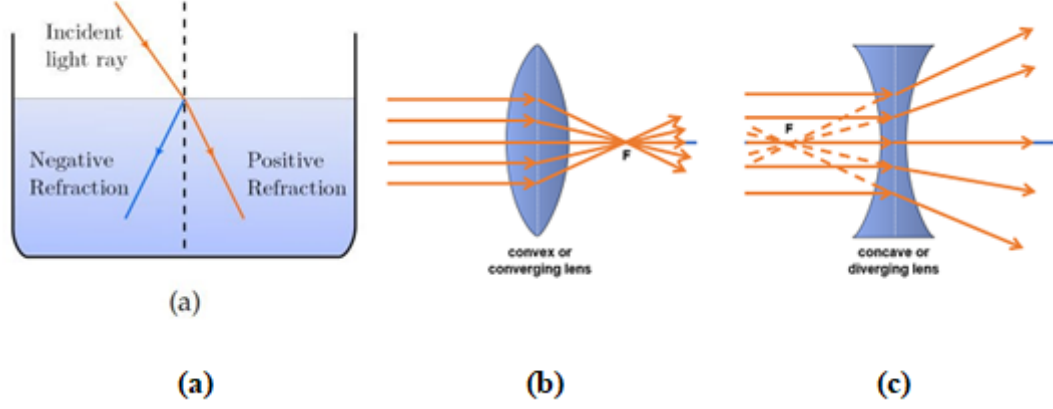


Figure I-9: (a) incident EM wave in a medium based on metamaterial. (b) convex lens has a diverging effect. (c) concave lens has a converging effect.

I.8. Electromagnetic Propagation in a Homogeneous Medium

Maxwell's equations can be expressed in the following form for a monochromatic electromagnetic wave propagating in a homogeneous medium [I.21, I.26]:

$$\vec{\nabla} \times \vec{E} = -\frac{\partial \vec{B}}{\partial t} \quad (I.8)$$

$$\vec{\nabla} \times \vec{H} = \frac{\partial \vec{D}}{\partial t} \quad (I.9)$$

Where:

- E: represents the electric field.
- H: represents the magnetic field.
- B: represents the magnetic induction.
- D: represents the electric induction.
- μ : is the magnetic permeability.
- ϵ : is the electrical permittivity.

The relationships between these quantities are given by:

$$\vec{B} = \mu \vec{H} \text{ and } \vec{D} = \epsilon \vec{E} \quad (I.10)$$

The relative permittivity and permeability of the propagation medium are denoted by ϵ_r and μ_r , respectively, while the permittivity and permeability of free space (vacuum) are represented by ϵ_0 and μ_0 . These are related as follows:

$$\mu = \mu_r \times \mu_0 \quad \text{and} \quad \epsilon = \epsilon_r \times \epsilon_0 \quad (\text{I.11})$$

For a plane wave propagating in the z-direction, the electromagnetic fields (E) and (H) exhibit both temporal ($e^{j\omega t}$) and spatial ($e^{-j\gamma z}$) dependencies. The fields can be expressed as:

$$\vec{E} = \vec{E}_0 e^{j\omega t} e^{-j\gamma z} \quad \text{and} \quad \vec{H} = \vec{H}_0 e^{j\omega t} e^{-j\gamma z} \quad (\text{I.12})$$

Here, E_0 is the amplitude of the electric field, and ω is the angular frequency, defined as:

$$\omega = 2\pi f \quad (\text{I.13})$$

where f is the frequency of the wave [I.27].

The constant γ is known as the propagation constant and is defined as:

$$\Gamma = \alpha + j\beta \quad (\text{I.14})$$

Where α represents the attenuation constant, and β represents the phase constant [I.28]. Based on the form of γ , three distinct cases can be identified:

- Purely Real Propagation Constant $\gamma = \alpha$:

The electromagnetic wave is evanescent, and its amplitude decays exponentially during propagation.

- Purely Imaginary Propagation Constant $\gamma = j\beta$:

The electromagnetic wave propagates without attenuation.

- Complex Propagation Constant $\gamma = \alpha + j\beta$:

The electromagnetic wave propagates with a loss term, characterized by both attenuation and phase shift [I.29].

In the case of a lossless medium, where $\alpha = 0$, the propagation equations simplify to:

$$\vec{E} = \vec{E}_0 e^{j\omega t} e^{-jkz} \quad \text{and} \quad \vec{H} = \vec{H}_0 e^{j\omega t} e^{-jkz} \quad (\text{I.15})$$

These equations describe the propagation of electromagnetic waves in a homogeneous medium, highlighting the interplay between the electric and magnetic fields as governed by Maxwell's equations.

With k the wave vector, knowing that in the hypothesis of a TEM propagation:

$$k = \beta \quad (I.16)$$

We can rewrite Maxwell's equations in this form using equations (I.8), (I.9) and (I.10):

$$\vec{K} \times \vec{E} = \omega\mu \vec{H} \quad \text{and} \quad \vec{K} \times \vec{H} = -\omega\epsilon \vec{E} \quad (I.17)$$

I.9. Electromagnetic Wave Propagation in Double-Negative Medium (Metamaterials)

From equations (I.17), it can be observed that when both the electric permittivity (ϵ) and magnetic permeability (μ) are positive, the vectors E , H , and K form a direct, or right-handed, system, as illustrated in Figure I-10. In this configuration, the vector S denotes the Poynting vector, which represents the power flow per unit area of the electromagnetic wave and is defined by the following relation:

$$\vec{S} = \frac{1}{2} \vec{E} \times \vec{H}^* \quad (I.18)$$

As a result, the vectors S and K propagate in the same direction.

In contrast, another solution to this relation arises when both ϵ and μ are simultaneously negative. Under these conditions, the vectors E , H , and K form an indirect, or left-handed, system, as illustrated in Figure I-10. This behavior is described by Veselago's theory, first introduced in 1964 [I.5]. Materials exhibiting such properties are referred to as metamaterials, as they possess electromagnetic characteristics that are not found in naturally occurring substances

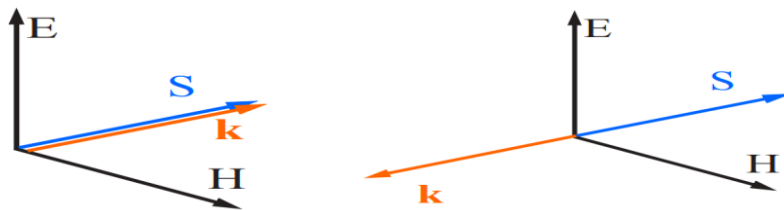


Figure I-10: Vector representation of a plane wave.

Equation (I.19) defines the phase velocity, while equation (I.20) defines the group velocity [I.30].

Since the three vectors E , H , and K form an indirect triplet, the wave's phase and group velocity are in opposite directions (see Figure I-10):

$$v_{\phi} = \frac{\omega}{k} \quad (I.19)$$

$$v_g = \frac{\partial \omega}{\partial k} \quad (I.20)$$

Since the three vectors form an indirect triplet, the wave's phase and group velocity are in opposite directions (see Figure I-11):

$$\begin{cases} \text{Right-handed medium: } v_{\phi} > 0 \text{ (} k > 0 \text{) and } v_g > 0 \\ \text{Left-handed medium: } v_{\phi} < 0 \text{ (} k < 0 \text{) and } v_g > 0 \end{cases} \quad (I.21)$$

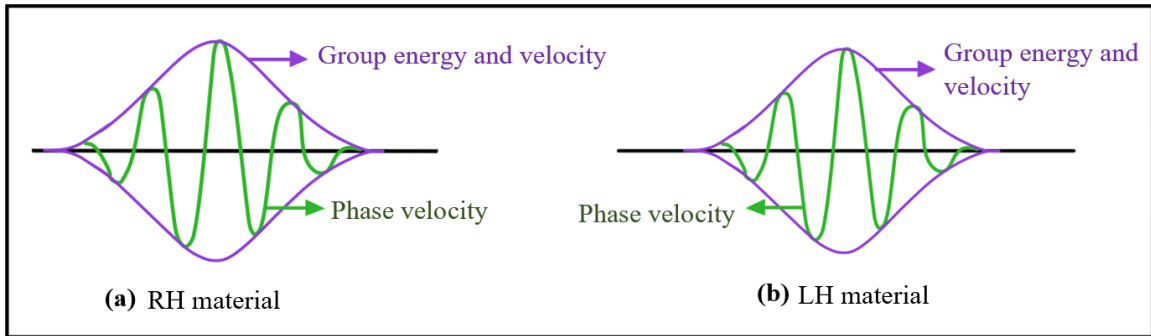


Figure I-11: Phase and energy propagation in a traditional material and a left-handed material.

Veselago predicted that media with simultaneously negative permittivity (ϵ) and permeability (μ) would exhibit several unique and counterintuitive properties, including:

- **Reversed Doppler Effect:** In contrast to conventional media, the observed frequency shift occurs in the opposite direction relative to the motion of the source.
- **Reversed Snell–Descartes law (negative refractive index):** When light enters such a medium, it refracts on the same side of the normal, a behavior associated with a negative refractive index.
- **Reversed Cerenkov radiation:** The Cerenkov effect, which describes electromagnetic radiation emitted when a charged particle moves through a medium at

a speed greater than the phase velocity of light in that medium, is also inverted in left-handed materials [I.5].

Cerenkov radiation is a visible electromagnetic phenomenon that occurs when charged particles travel through a dielectric medium at speeds exceeding the speed of light within that medium [I.21]. In conventional materials, where the refractive index is positive ($n > 0$), the emitted wavefronts trail behind the moving particle, forming a radiation cone with an angle θ less than 90 degrees relative to the direction of motion. This angle is defined between the Poynting vector of the emitted radiation and the particle velocity.

In contrast, in left-handed media, this behavior is reversed, as illustrated in Figure I-11 (a) and described in [I.31], where the radiation is emitted in the opposite direction.

The angle θ associated with Cerenkov radiation is given by:

$$\cos \theta = \frac{c}{n(\omega)v} \quad (I.22)$$

Here, c represents the speed of light in a vacuum, and v denotes the speed of the charged particle.

In a conventional medium with a positive refractive index ($n > 0$), the Cerenkov radiation wavefront propagates in a direction opposite to the motion of the charged particle, forming an angle θ less than 90 degrees with respect to the particle velocity.

In contrast, when the refractive index is negative ($n < 0$), the Cerenkov effect is inverted [I.5]. The emitted wavefront propagates in the same general direction as the particle, with a velocity equal to $c/|n|$. As a result, the angle θ between the radiation and the particle velocity becomes greater than 90 degrees, as illustrated in Figure I-12 (b) [I.32].

It should be emphasized that materials with a negative refractive index operate only over limited frequency bands [I.31]. Since Cerenkov radiation spans a wide range of frequencies, including regions where the refractive index may be positive, the emitted wavefronts can propagate either forward or backward relative to the particle motion, depending on the frequency range considered.

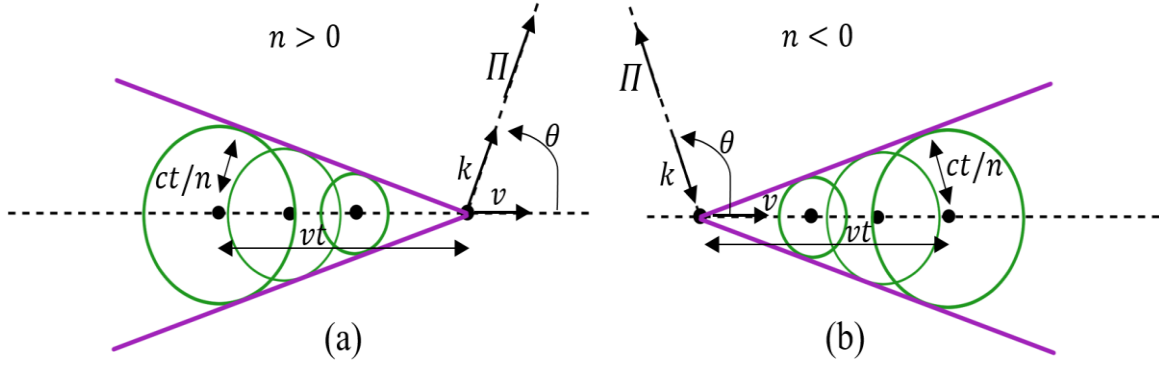


Figure I-12: Illustration of the Cerenkov effect. (a) Conventional medium. (b) Left-handed medium.

I.9.1. The Reversed Doppler Effect

The reversal of the Doppler effect is another notable phenomenon observed in electromagnetic wave propagation within left-handed media, as originally predicted by Veselago [I.5]. To illustrate this behavior, consider a source S moving along the z -axis while emitting electromagnetic radiation.

waves with angular frequency ω_0 as depicted in Figure I-13. If the source moves at velocity $v_s = \frac{z}{t}$ in the positive z direction, observed Doppler frequency $\omega_{Doppler}$ can be expressed as:

$$\omega_{Doppler} = \omega_0 - \Delta\omega \quad \text{with} \quad \Delta\omega = \omega_0 \frac{n v_s}{c} \quad (\text{I.23})$$

Here, c represents the speed of light in vacuum, while n denotes the refractive index of the medium.

In conventional materials ($n > 0$), the Doppler effect leads to an increase in observed frequency when the source approaches the observer, and a decrease when it moves away [I.5]. However, in left-handed media ($n < 0$), this behavior is reversed [I.6]. In this case, observers located in the direction of motion detect a decrease in frequency, whereas those positioned opposite to the motion observe an increase [I.15]. This inversion results directly from the negative refractive index, which fundamentally modifies wave propagation characteristics [I.5].

This effect highlights the unusual and non-intuitive behavior of left-handed media, providing further support for Veselago's theoretical framework [I.6].

$$\omega_{\text{observed}} = \omega_0 - \Delta\omega \quad (\text{I.24})$$

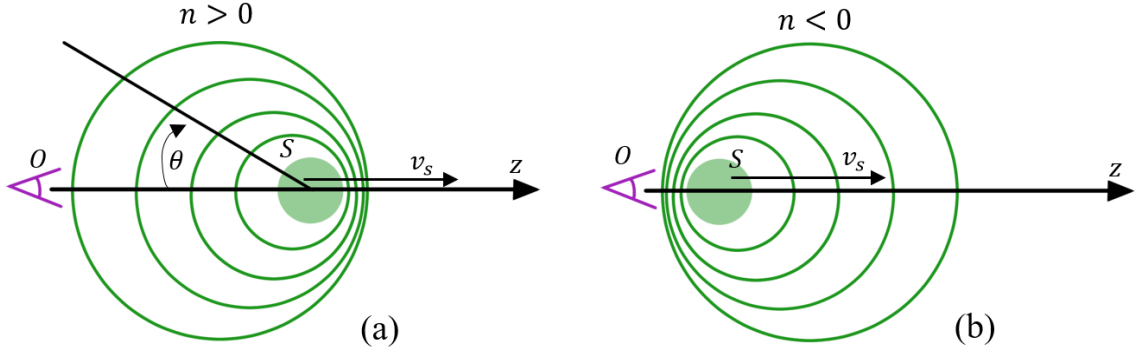


Figure I-13: The doppler effect for: (a) RH media. (b) LH media.

I.9.2. The Reversed Snell's Law with a Negative Refractive Index

The refractive index of a material is calculated using Equation (I.1). The propagation constant is determined as follows:

$$k = nk_0 = n \frac{\omega}{c} \quad (\text{I.25})$$

Given that $k < 0$, it can be concluded that the refractive index (n) of a double-negative (DNG) material is negative.

As a result of this finding, Snell's law can be generalized to account for the possibility of an interface between a DNG material and a double-positive (DPS) material [I.5]:

$$n_i \sin(\theta_i) = n_t \sin(\theta_t) \quad (\text{I.26})$$

Here:

- n_i and n_t are the refractive indices of the materials supporting the incident and transmitted waves, respectively.
- θ_i and θ_t are the angles of incidence and transmission, measured with respect to the normal at the interface between the two materials [I.15].

A distinction can be made between two cases:

Positive refraction: This occurs when both materials have the same sign of refractive index (i.e., both positive or both negative). In this case, the angles θ_i and θ_t are both positive (see Figure I-14 (a)).

Negative refraction: If the signs of the two materials vary, the refraction is negative since one of the two angles, θ_i or θ_t , would be negative [I.32], as displayed in Figure I-14 (b).

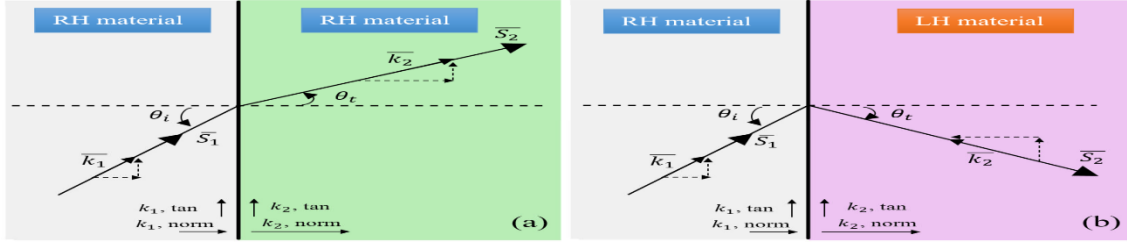


Figure I-14: Electromagnetic wave refraction at the interface between two different materials. (a) positive refraction. (b) negative refraction.

I.10. Theory of the Metamaterial Transmission Line

Most naturally occurring materials are characterized by positive permittivity (ϵ) and permeability (μ). Conventional right-handed materials (RHMs), including standard transmission line structures, naturally exhibit these positive electromagnetic properties. In contrast, metamaterial transmission lines form a class of one-dimensional engineered structures designed to realize unconventional electromagnetic responses, such as negative refractive index or left-handed (LH) behavior, which do not exist in naturally occurring media [I.18].

I.10.1. Mathematical Modeling of Transmission Lines

A transmission line supporting transverse electromagnetic (TEM) wave propagation can be modeled as a distributed parameter system [I.5]. For a differential line segment Δz :

- Series impedance: $\frac{Z}{\Delta z}$ includes inductance L and resistance R.
- Shunt admittance: $\frac{Y}{\Delta y}$ includes capacitance C and conductance G.

In an ideal lossless system ($R = 0$, $G = 0$), the governing telegrapher's equations simplify to:

$$\frac{\partial v}{\partial z} = -L \frac{\partial I}{\partial t} \quad \text{and} \quad \frac{\partial I}{\partial z} = -C \frac{\partial v}{\partial t} \quad (\text{I.27})$$

where V and I represent the voltage and current along the line, respectively [I.30].

I.10.2. Significance of Metamaterial Transmission Lines

- **Negative refractive index:** Achieved when both ϵ and μ are negative, enabling phenomena like backward wave propagation [I.5].
- **Miniaturization:** LH transmission lines allow compact device designs by exploiting phase-advance properties [I.33].
- **Frequency selectivity:** Tunable resonant frequencies enable applications in filters, sensors, and antennas [I.34].

Veselago demonstrated in his work that the Cerenkov effect is also reversed in left-handed media [I.5]. Furthermore, subsequent studies have supported and expanded the concept of negative refraction through both theoretical and experimental contributions, notably those of Pendry [I.6] and Shelby [I.15].

This framework establishes a connection between conventional transmission line theory and advanced metamaterial concepts, enabling the development of novel electromagnetic devices.

I.10.3. Right-Handed Transmission Line

To gain a clearer understanding of metamaterial properties and behavior in the microwave regime, C. Caloz, T. Itoh [I.35, I.18], and G. Eleftheriades [I.36] introduced, in 2002, a transmission line–based modeling approach. This technique relies on a periodic arrangement of unit cells to represent either a conventional transverse electromagnetic (TEM) propagation line or a lossy right-handed transmission line (RHTL). For clarity, the idealized lossless case is considered in the following discussion (see Figure I-15).

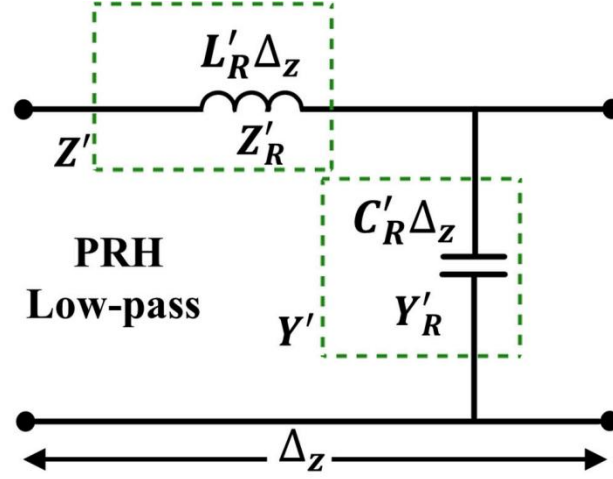


Figure I-15: Pure right-handed transmission line model [I.37].

A unit cell of a typical right-handed transmission line consists of:

- A series inductance (L)
- A shunt capacitance (C)

In practical scenarios (with losses), additional elements are introduced:

- A series resistance (R) to account for metallic losses.
- A shunt conductance (G) to represent dielectric losses.

From the equivalent circuit, the following relationships can be derived:

$$\gamma = \alpha + j\beta = \pm\sqrt{ZY} \quad (\text{I.28})$$

For a lossless line, Equation (I.28) simplifies to:

$$\gamma = j\beta = \pm\sqrt{ZY} \quad (\text{I.29})$$

Where:

- Z is the impedance of the series elements:

$$Z = j\omega L \quad (\text{I.30})$$

- Y is the admittance of the shunted elements:

$$Y = j\omega C \quad (\text{I.31})$$

- $\beta = \omega\sqrt{LC}$ is the phase constant.

- The phase velocity v_{ph} and the group velocity v_g are given by:

$$v_{ph} = \sqrt{\frac{1}{LC}} > 0 \quad (I.32)$$

$$v_g = \sqrt{\frac{1}{LC}} > 0 \quad (I.33)$$

In this case, the phase velocity and group velocity are equal and share the same sign, indicating a right-handed medium [I.30]. The phase constant β of the pure right-handed transmission line, as a function of frequency, is always positive and follows the relationship $\beta = \omega\sqrt{LC}$ [I.30].

I.10.4. Left-Hand Transmission Line

In 2002, a research team from the University of Toronto introduced the transmission line technique for creating left-handed materials (LHMs) [I.38, I.18]. This approach is based on mapping electromagnetic field components (E and H) to their equivalent voltage and current distributions in a distributed LC network. As established in the literature [I.18], dielectric properties such as permittivity and permeability can be effectively represented using distributed LC networks. Figure I-16 illustrates a two-dimensional (2D) distributed LC network, demonstrating how these material properties correspond to the network's series impedances and distributed shunt admittances. The network consists of a series impedance X' per unit length in the X and Z' directions, as well as a shunt admittance Y' [I.18] per unit length in the Y direction. This configuration provides a practical framework for realizing negative refractive index materials through engineered circuit elements [I.15].

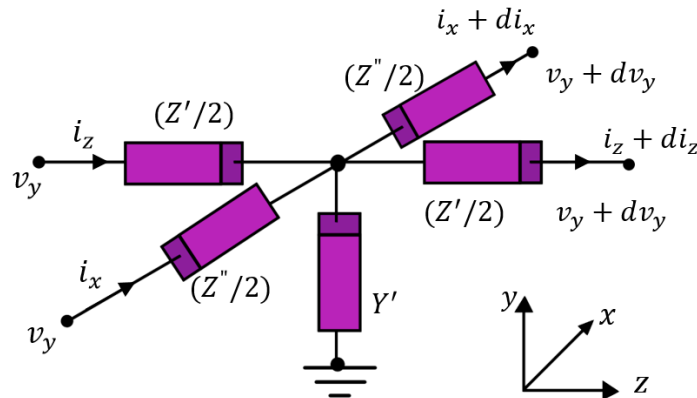


Figure I-16: Unit cell for a distributed L-C 2D network.

Since left-handed materials (LHMs) exhibit effective homogeneity, they can be represented using one-dimensional transmission line models. Although wave propagation may occur in any spatial direction, the key distinguishing feature of LHMs is the reversal of the propagation characteristics compared to conventional media. Based on fundamental transmission line theory, the core properties of left-handed transmission lines can be derived in a straightforward manner.

In the ideal case, the left-handed transmission line circuit shown in Figure I-17 can be viewed as the dual counterpart of the conventional right-handed transmission line. This duality results from swapping inductive and capacitive elements in the equivalent circuit model, which leads to effective negative permittivity and permeability—defining characteristics of LHMs [I.18, I.39].

The unusual electromagnetic behavior of LHMs, including backward-wave propagation and negative refraction, makes them particularly useful for applications such as subwavelength imaging, cloaking technologies, and advanced antenna systems [I.22]. In addition, the one-dimensional transmission line representation provides a simplified yet robust analytical tool for studying and designing these materials, allowing for more precise exploration of their non-conventional electromagnetic responses [I.40].

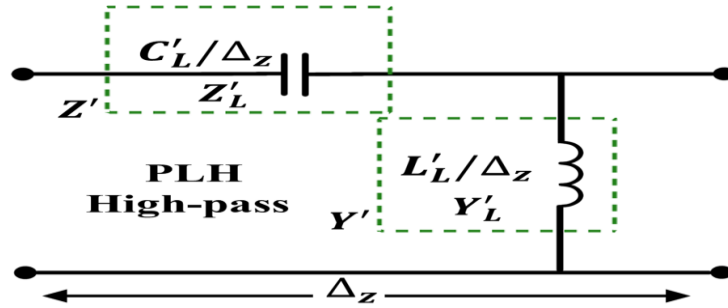


Figure I-17: Pure left-hand transmission line model [I.37].

The unique properties of left-handed metamaterials (LHMs), such as negative permittivity (ϵ) and permeability (μ), naturally raise an important question: whether the impedance (Z') and admittance (Y') parameters in a network representation can also become negative. This can be understood by examining the behavior of the equivalent circuit elements.

By exchanging the roles of inductance (L) and capacitance (C), negative effective values can be realized. In practice, this corresponds to replacing the series inductor with a series

capacitor, and the shunt capacitor with a shunt inductor. From an impedance standpoint, this transformation leads to the following relations [I.33]:

$$Z' = \frac{1}{j\omega C} \rightarrow \mu = \frac{Z'}{j\omega} = -\frac{1}{\omega^2 C} \quad (I.34)$$

$$Y' = \frac{1}{j\omega L} \rightarrow \varepsilon = \frac{Y'}{j\omega} = -\frac{1}{\omega^2 L} \quad (I.35)$$

Here, Z' and Y' represent the series reactance and shunt susceptance per unit length, respectively. The negative values of L' and C' are a direct consequence of the unconventional electromagnetic properties of LHMs.

The propagation constant (β) of the left-handed transmission line can be derived as:

$$\beta = -\sqrt{-Z'Y'} = \frac{1}{\omega\sqrt{LC}} \quad (I.36)$$

From the propagation constant, the phase velocity v_{ph} and group velocity v_g can be expressed as:

$$V_{ph} = \frac{\omega}{\beta} = -\omega^2\sqrt{LC} \quad (I.37)$$

$$V_g = \frac{\partial\omega}{\partial\beta} = \frac{1}{\frac{\partial\beta}{\partial\omega}} = \omega^2\sqrt{LC} \quad (I.38)$$

The opposite signs of the phase velocity and group velocity indicate the presence of backward wave propagation in the left-handed transmission line network. This phenomenon is a hallmark of LHMs and is directly linked to their negative refractive index (n), which is given by:

$$n = \frac{C_0}{V_{ph}} = -\frac{C_0}{\omega^2\sqrt{LC}} < 0 \quad (I.39)$$

where C_0 is the speed of light in a vacuum.

Here, C_0 represents the speed of light in vacuum. The negative refractive index is a key defining feature of LHMs and enables unusual applications such as subwavelength imaging and superlens systems [I.5, I.6].

The possibility of achieving opposite phase and group velocities, together with a negative refractive index, highlights the strong potential of LHMs in areas such as telecommunications, optics, and microwave engineering. These materials fundamentally challenge classical electromagnetic intuition and open new pathways for controlling and manipulating wave propagation [I.16].

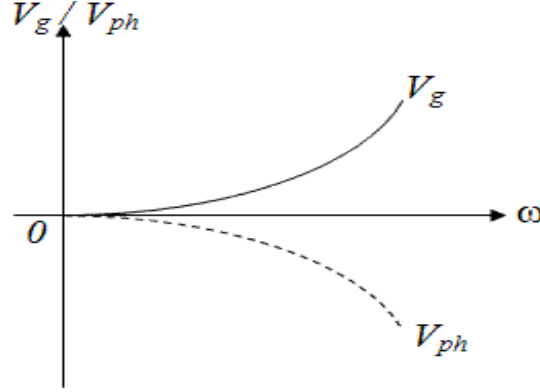


Figure I-18: group and phase velocity diagram as a function of the frequency of a metamaterial transmission line.

I.10.5. Phase Constant and Frequency Dependence in Left-Handed Transmission Line

The phase constant (β) of a purely left-handed transmission line exhibits a negative slope across the entire frequency range, as illustrated in Figure I-18. This behavior is derived from the equation $\beta = \frac{1}{\omega\sqrt{LC}}$, where ω is the angular frequency, and L' and C' represent the equivalent inductance and capacitance per unit length, respectively. This negative phase constant is a direct consequence of the unique electromagnetic properties of left-handed metamaterials (LHMs), which support backward wave propagation and negative refractive index [I.5].

In [I.41], a two-dimensional (2D) network of left-handed transmission lines was shown to support negative refraction and subwavelength focusing. The work included fabrication of the structure, full-wave electromagnetic simulations confirming negative refraction behavior, and experimental validation of focusing effects using transmission line implementations.

For example, an antenna design reported in [I.42] experimentally demonstrated the existence of backward-wave propagation, which is a key signature of left-handed media. In addition, various microwave components such as leaky-wave antennas [I.42, I.43], power dividers

[I.44, I.45], couplers [I.46], and other devices based on left-handed transmission line principles have been proposed and experimentally realized.

Building upon the physical implementation of the symmetric condensed node in transmission line matrix modeling, the 2D structure illustrated in Figure I-19 has been extended into a three-dimensional (3D) isotropic metamaterial transmission line topology. This development allows for free-space excitation and significantly broadens the range of potential applications of LHMs in practical systems [I.74].

I.10.6. Composite Right/Left-Handed (CRLH) Transmission Line

In 2003, a more general model known as the composite right/left-handed (CRLH) transmission line was introduced [I.48]. This model provides a unified framework for the analysis and design of practical left-handed transmission lines. As illustrated in Figure I-19, the CRLH transmission line is characterized by a series impedance per unit length Z' (Ω/m) and a shunt admittance per unit length Y' (S/m).

The series branch Z' is formed by a combination of a series inductance per unit length L'_R (H/m) and a series capacitance per unit length C'_L (F/m), while the shunt branch Y' consists of a shunt capacitance per unit length C'_R (F/m) in parallel with a shunt inductance per unit length L'_L (H/m).

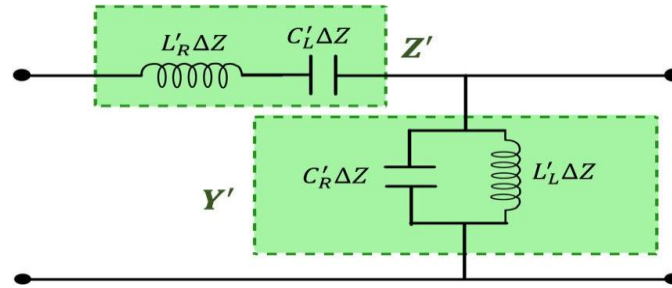


Figure I-19: A unit cell of the equivalent circuit for a composite right-hand / left-hand transmission line.

The propagation constant (γ) of the CRLH transmission line is given by:

$$\gamma = \alpha + j\beta = \sqrt{Z'Y'} \quad (\text{I.40})$$

Here, α represents the attenuation constant, and β is the phase constant. The series impedance Z' and shunt admittance Y' are defined as:

$$Z' = j \left(\omega L'_R - \frac{1}{\omega C'_L} \right) \quad (\text{I.41})$$

$$Y' = j \left(\omega C'_R - \frac{1}{\omega L'_L} \right) \quad (\text{I.42})$$

For simplicity, the following resonant frequencies are introduced:

$$\omega'_R = \frac{1}{\sqrt{L'_R C'_R}} \quad \text{rad} \cdot \text{m/s} \quad (\text{I.43})$$

$$\omega'_L = \frac{1}{\sqrt{L'_L C'_L}} \quad \text{rad} \cdot \text{m/s} \quad (\text{I.44})$$

These frequencies define the transition regions between right-handed and left-handed propagation regimes in the CRLH transmission line. The CRLH model offers a flexible framework for the design of metamaterial-based devices with engineered electromagnetic responses, supporting applications such as broadband antennas, filters, and phase shifters [I.49, I.34].

I.10.7. Series and Shunt Resonance Frequencies in CRLH Transmission Lines

The behavior of a composite right/left-handed (CRLH) transmission line is governed by its series and shunt resonance frequencies, which are defined as follows:

$$\omega_{se} = \frac{1}{\sqrt{L'_R C'_L}} \quad \text{rad/s} \quad (\text{I.45})$$

$$\omega_{sh} = \frac{1}{\sqrt{L'_L C'_R}} \quad \text{rad/s} \quad (\text{I.46})$$

Here, ω_{se} and ω_{sh} denote the series and shunt resonance frequencies, respectively. These parameters play a key role in defining the transition between right-handed (RH) and left-handed (LH) propagation regimes in the CRLH transmission line.

By substituting Equations (I.41) to (I.46) into Equation (I.40), and taking into account the sign reversal of the phase velocity in left-handed media, the complex propagation constant γ can be written as:

$$\gamma = js(\omega)\sqrt{\left(\frac{\omega}{\omega'_R}\right)^2 + \left(\frac{\omega'_L}{\omega'_R}\right)^2 - \left(\frac{C'_R}{C'_L} + \frac{L'_R}{L'_L}\right)} \quad (I.47)$$

The function $s(\omega)$ determines the sign of the propagation constant and is defined as:

$$s(\omega) = \begin{cases} +1 & \text{if } \omega < \max(\omega_{se}, \omega_{sh}) \text{ the right hand band} \\ -1 & \text{if } \omega < \min(\omega_{se}, \omega_{sh}) \text{ the left hand band} \end{cases} \quad (I.48)$$

This sign function ensures the correct phase velocity direction for both RH and LH modes.

I.10.8. Stopband and Dispersion Characteristics

Even in a lossless CRLH transmission line, when the angular frequency ω satisfies $\min(\omega_{se}, \omega_{sh}) < \max(\omega_{se}, \omega_{sh})$, the propagation constant γ becomes purely real ($\gamma = \alpha$), indicating a stopband region. In this frequency range, wave propagation does not occur and the structure behaves as an attenuating medium rather than a propagating one [I.34].

Figure I-20 illustrates the dispersion and attenuation characteristics of the CRLH transmission line [I.16]. The curves shown represent the phase constants corresponding to the right-handed β_{RH} and left-handed β_{LH} bands, as well as the ideal phase constants for the right-handed β_{PRH} and left-handed β_{PLH} regimes.

.

These parameters are defined as follows:

- β_{RH} : Phase constant of the right-hand band of the CRLH transmission line.
- β_{LH} : Phase constant of the left-hand band of the CRLH transmission line.
- β_{PRH} : Pure phase constant of the right-hand transmission line.

➤ β_{PLH} : Pure phase constant of the left-hand transmission line.

The dispersion diagram highlights the unique dual-band behavior of CRLH transmission lines, where the transition between RH and LH modes occurs at the resonance frequencies ω_{se} and ω_{sh} [I.18]. This behavior was experimentally verified in [I.52] through detailed S-parameter measurements.

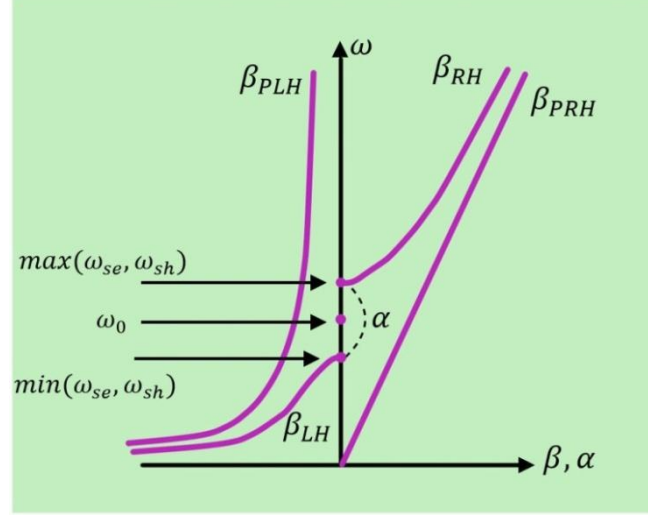


Figure I-20: The CRLH line dispersion / attenuation chart, when the energy propagation is along the z-axis.

I.10.9. Low-Frequency and High-Frequency Behavior of CRLH Transmission Lines

At low frequencies, the impedance of the series inductance L'_R is significantly lower than that of the series capacitance C'_L , and the admittance of the shunt capacitance C'_R is much less than that of the shunt inductance L'_L . Under these conditions, the series impedance Z' and shunt admittance Y' can be approximated as: $Z' \approx -\frac{1}{j\omega C'_L}$ and $Y' \approx -\frac{1}{j\omega L'_L}$. This implies that the right-handed (RH) contribution is negligible, and the transmission line behaves almost entirely as a left-handed (LH) line, exhibiting a hyperbolic dispersion pattern.

At high frequencies, the behavior of the CRLH transmission line shifts dramatically. Here, the series impedance Z' and shunt admittance Y' are dominated by the inductive and capacitive components, respectively: $Z' \approx j\omega L'_R$ and $Y' \approx j\omega C'_R$.

In this regime, the CRLH transmission line behaves like a pure right-handed transmission line, displaying a linear dispersion diagram.

I.10.10. Balanced and Unbalanced CRLH Transmission Lines

The performance of the CRLH transmission line in the intermediate frequency region is governed by the combined effects of its left-handed (LH) and right-handed (RH) components. The structure is defined as “balanced” when the series and shunt resonance frequencies are equal ($\omega_{se} = \omega_{sh}$) [I.50]. In this condition, the zeros of Z' and Y' coincide at the same frequency, eliminating the stopband region. As a result, the characteristic impedance of the balanced CRLH transmission line becomes independent of frequency, enabling ideal impedance matching over an effectively infinite bandwidth [I.53]. The characteristic impedance Z_C of the CRLH transmission line is given by:

$$Z_C = \frac{Z'}{Y} = \sqrt{\frac{Z'}{Y'}} = \sqrt{\frac{L'_L}{C'_L}} \cdot \sqrt{\frac{\left(\frac{\omega}{\omega_{se}}\right)^2 - 1}{\left(\frac{\omega}{\omega_{sh}}\right)^2 - 1}} = Z_R = Z_L \quad (I.49)$$

When ($\omega_{se} = \omega_{sh}$), the characteristic impedances for the pure right-handed Z_R and left-handed Z_L transmission lines simplify to:

$$Z_R = \sqrt{\frac{Z'}{Y'}} = \sqrt{\frac{L'_R}{C'_R}} \quad (I.50)$$

$$Z_L = \sqrt{\frac{Z'}{Y'}} = \sqrt{\frac{L'_L}{C'_L}} \quad (I.51)$$

I.10.11. Applications of CRLH Transmission Lines

The distinctive phase behavior of CRLH transmission lines, including dual-band operation, enhanced bandwidth, nonlinear frequency dependence, and the presence of a critical frequency at which the phase velocity becomes zero, has enabled a wide range of innovative applications [I.54]. These can be grouped as follows:

- **Guided-wave applications:** CRLH transmission lines are employed in components such as phase shifters, filters, and couplers, offering improved performance alongside reduced physical size [I.55].

- **Radiated-wave applications:** These include frequency-scanning leaky-wave antennas, electronically steerable antenna systems, and zero-order resonant antennas [I.43].

I.11. Progression in Realizing Metamaterials

The realization of metamaterials has followed a three-stage development process:

- **Engineering a medium with negative permittivity:** achieved through structures that exhibit plasma-like electromagnetic behavior at specific frequency ranges [I.54].
- **Achieving a medium with negative permeability:** realized by designing magnetic resonators that generate an effective negative magnetic response [I.55].
- **Combining both responses:** the integration of these two artificial media leads to a doubly negative medium, characterized by simultaneous negative permittivity and permeability [I.43].

I.11.1. Metamaterials with Negative Permittivity

Negative permittivity is a plasma-related electromagnetic property that has been widely investigated in optics, infrared physics, and electromagnetics. Pendry and co-workers showed that a periodic array of parallel metallic rods, with radius r and lattice spacing α , can reproduce an effective response similar to that of a dilute electron plasma [I.56]. This effect results from the collective motion of conduction electrons within the metallic structure under high-frequency excitation.

When the operating frequency approaches the plasma frequency f_p , the applied electromagnetic field induces a restoring force on the charge carriers, leading to an oscillatory response that can be described by:

$$f_p^2 = \frac{nq^2}{4\pi^2\epsilon_0 m_{eff}} \quad (I.52)$$

where q is the elementary charge, n is the electron density, m_{eff} is the effective mass of the electrons, and ϵ_0 is the permittivity of free space. The plasma permittivity is only observable for frequencies below the plasma frequency.

The permittivity dispersion of the plasma is given by:

$$\varepsilon(\omega) = 1 - \frac{\omega_p^2}{\omega^2} \quad (I.53)$$

The plasma frequency is calculated as:

$$\omega_p^2 = \frac{nq^2}{\varepsilon_0 m_{\text{eff}}} \quad (I.54)$$

With q indicating the charge of the electron and m_{eff} denoting the effective mass of the electron.

The approach consisted in making a network of heavier metal inclusions with a lower electron density. The chosen configuration consists of a network of metal rods of radius r , with a pitch of the network a (Figure I-21).

For this setting, the plasma frequency is:

$$\omega_p^2 = \frac{2\pi C_0^2}{a^2 \ln\left(\frac{a}{r}\right)} \quad (I.55)$$

Here, C_0 is the speed of light, a is the lattice constant, and r is the radius of the metal rods.

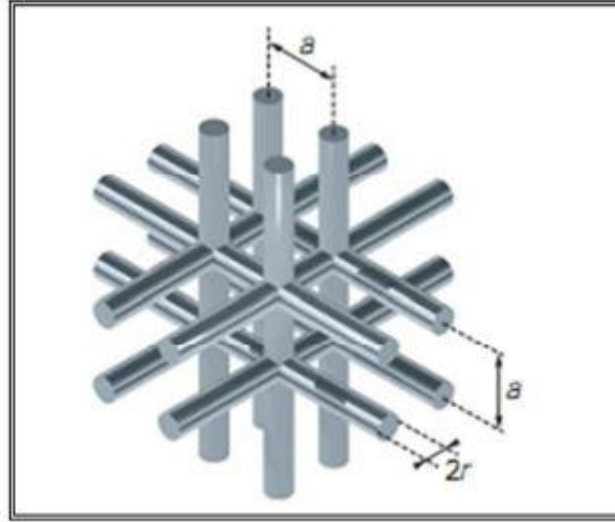


Figure I-21: Network of cylindrical metal wires with negative permittivity.

I.11.2. Metamaterials with Negative Permeability

In the microwave regime, certain materials such as ferromagnetic and antiferromagnetic composites like MgF_2 and FeF_2 can exhibit negative permeability [I.51]. However, these materials present major limitations, including high density and significant magnetic losses, which restrict their practical use. Consequently, considerable research has focused on

developing non-magnetic structures capable of producing effective negative permeability without relying on conventional magnetic media [I.15].

In 1999, J. Pendry proposed a key solution to this problem by introducing a metallic structure known as the “Swiss roll” configuration (Figure I-22) [I.51]. This design generates an artificial magnetic resonance without requiring any intrinsic magnetic material [I.51, I.16]. The Swiss roll consists of tightly wound conductive spirals arranged around a cylindrical core of radius R .

The spiral turns are separated by a dielectric layer N , while adjacent windings are spaced by a distance d_C [I.51]. When an external magnetic field is applied along the axis of the cylinder, it induces circulating currents within the conductive spiral structure [I.16]. These currents pass through the capacitive gaps between turns, forming a resonant LC-type behavior. The resulting resonance produces strong circulating currents that emulate a magnetic dipole response [I.51, I.16]. Consequently, the structure exhibits an effective negative permeability in the frequency range close to resonance [I.51].

The Swiss roll structure can be treated as an effectively homogeneous medium because its largest physical dimension remains much smaller than the operating wavelength [I.16]. This allows it to be modeled as a bulk material with effective electromagnetic properties, making it a useful building block in metamaterial design [I.51, I.16].

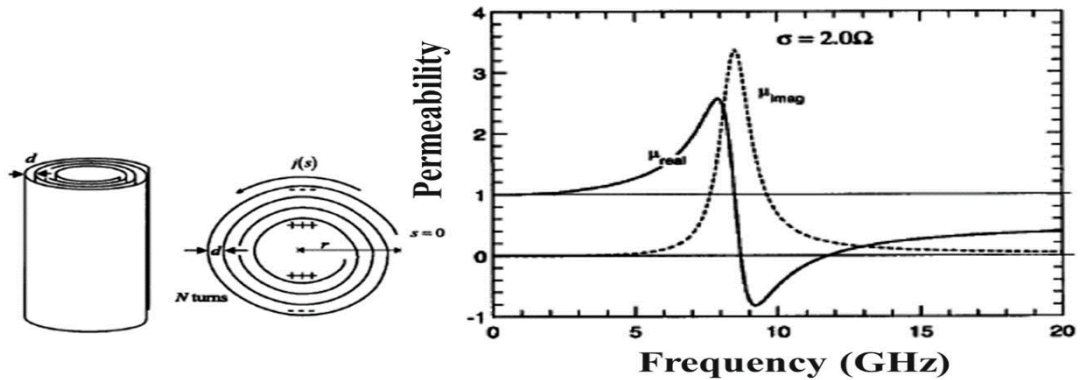


Figure I-22: The Swiss-Roll structure introduced by pendry, (a) front view and profile view
(b) the calculated and measured Permeability for a Swiss 35-turn roller.

I.12. Effective Permeability of Swiss Roll Structures

Based on the pioneering work of Pendry et al. [I.51], the effective permeability μ_{eff} of a Swiss roll structure can be determined using the following relationship:

$$\mu_{\text{eff}} = 1 - \frac{F}{1 + \frac{2\sigma_i}{\omega r \mu_0 (N-1)} - \frac{dC_0^2}{2\pi^2 \omega^2 r^3 (N-1)}} \quad (\text{I.56})$$

Here, C_0 represents the speed of light in a vacuum, and ω is the angular frequency. The conductivity of the spiral is denoted by σ , while ϵ represents the permittivity of the insulating material between the conductive layers. The filling rate of the active material is given by F .

The effective permeability can also be expressed in a simplified form as:

$$\mu_{\text{eff}} = 1 - \frac{F\omega^2}{\omega^2 - \omega_0^2 + i\Gamma\omega} \quad (\text{I.57})$$

Here, ω_0 is the resonance frequency of the structure, and γ accounts for losses in the system. The resonance frequency ω_0 is given by:

$$\omega_0 = \sqrt{\frac{dC_0^2}{2\pi^2 r^3 (N-1)}} \quad (\text{I.58})$$

In this equation, r is the radius of the Swiss roll, and N is the number of turns in the spiral.

I.12.1. Resonance Damping in Swiss Roll Structures

The resonance damping γ in a Swiss roll structure is a critical parameter that accounts for energy losses within the system. It is given by the following relation:

$$\omega_0 = \sqrt{\frac{dC_0^2}{2\pi^2 r^3 (N-1)}} \quad (\text{I.59})$$

Here, σ is the conductivity of the spiral, ω is the angular frequency, ϵ_0 is the permittivity of free space, and N is the number of turns in the spiral. This damping factor influences the sharpness of the resonance and the overall performance of the structure [I.15].

Figure I-22 illustrates the measured and estimated effective permeability of the Swiss roll structure. The structure exhibits resonance only when the magnetic field is applied along the axis of the roll [I.15]. In other directions, the structure does not resonate, leading to anisotropic behavior [I.16]. Additionally, for transverse waves incident on a network of these structures, where the electric field is parallel to the cylinders, significant absorption is

observed [I.18]. This anisotropic behavior can be undesirable in some applications, as it limits the versatility of the structure [I.19].

I.13. Metamaterials with Negative Permittivity ϵ and Permeability μ

To achieve a material with a negative refractive index, both the relative permittivity ϵ and permeability μ must be negative simultaneously. While natural materials can exhibit negative permittivity (e.g., plasmas and metals in the infrared range) or negative permeability (e.g., ferromagnetic materials at gyromagnetic resonance), no natural material exhibits both properties at the same time [I.5].

The first practical example of a doubly negative composite medium was introduced by D. Smith in 2000 [I.16]. Inspired by the work of J. Pendry [I.57], Smith combined a network of metal rods (to achieve negative permittivity) with split-ring resonators (SRRs) (to achieve negative permeability) (Figure I-23). This composite structure demonstrated transmission in a frequency band where both ϵ and μ were negative, effectively realizing a left-handed medium [I.16].

This groundbreaking experiment provided the first experimental validation of Veselago's theory [I.6], which predicted the propagation of electromagnetic waves through a medium with simultaneously negative permittivity and permeability. Smith's work marked the birth of the first practical metamaterial, opening the door to a wide range of applications in electromagnetics and optics [I.16].

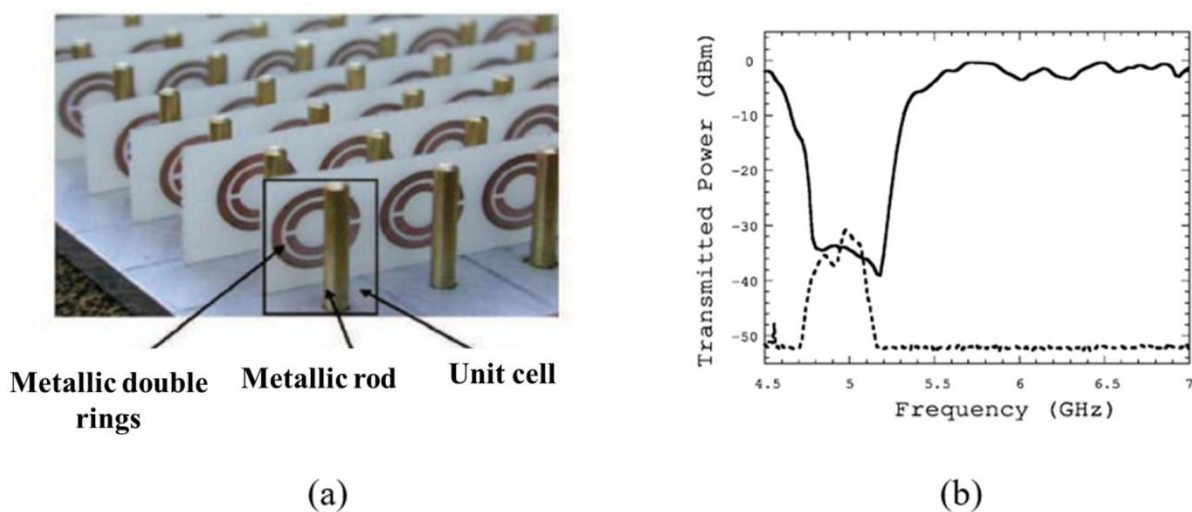


Figure I-23: Association of a network of rods with periodic SRRs. (b) the measured transmission power of the SRR array.

I.14. Evolution of Resonator Designs: Ω -Shaped and S-Shaped Resonators

The development of resonator designs has played a crucial role in advancing metamaterial technology. The first Ω -shaped resonator was introduced by Saadoun [I.58], while the S-shaped resonator was originally proposed by Prosvirnin for bi-anisotropic structures (BIP) [I.59] (Figure I-24). These resonator designs are unique because they incorporate both a magnetic dipole in the ring and an electric dipole in the arms, enabling them to interact with both electric and magnetic fields simultaneously.

According to the authors of [I.60], the appeal of these resonator designs lies in their simplicity and versatility. They can be easily arranged in a periodic structure, allowing for polarization in two orthogonal directions. This property makes them ideal for creating two-dimensional (2D) metamaterials by stacking multiple layers of dielectric substrates. The ability to control both electric and magnetic responses in a single structure opens up new possibilities for designing compact and efficient metamaterials.

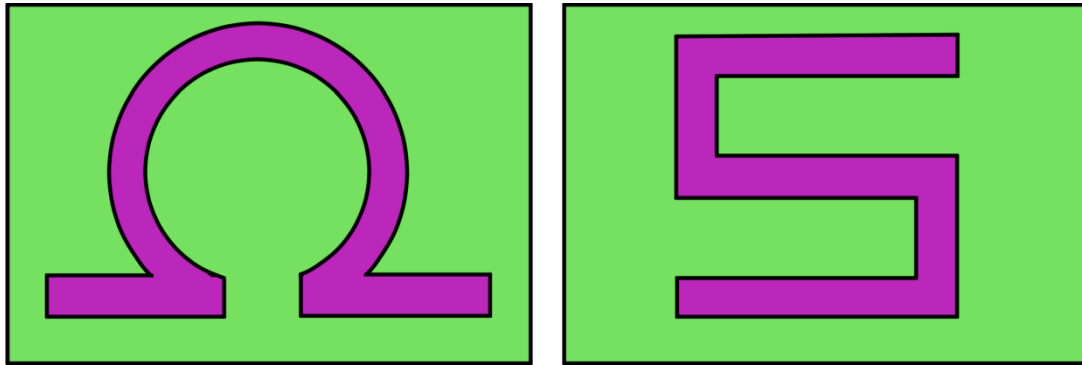


Figure I-24: Structure of resonators with negative refractive index. (a) "S" resonator. (b) " Ω " resonator.

I.15. Areas of Application

I.15.1. Metamaterial Superlens

Metamaterials have revolutionized the concept of lenses by enabling negative refraction, which allows electromagnetic waves to converge at a focal point rather than diverging outward. Unlike traditional convex lenses, which rely on curvature to focus light, a metamaterial lens can achieve focusing using a flat plate with a negative refractive index [I.5]. This groundbreaking idea was first proposed by J. Pendry in 2000, who suggested the realization of Veselago's "flat lens"—a simple plate with a refractive index of (-1) [I.6].

Pendry's work demonstrated the practical realization of Veselago's theoretical optical medium, albeit limited to a specific frequency range [I.5, I.6].

In 2005, a significant breakthrough was achieved by the team of Xiang Zhang at the University of California, Los Angeles [I.61]. They developed a metamaterial superlens composed of a thin film of silver sandwiched between two semiconductor layers. This superlens demonstrated a resolution greater than one-fifth of the wavelength for frequencies in the visible spectrum [I.61]. Figure I-25 illustrates the imaging process using a convex lens (left-handed medium) and a right-handed (RH) plate. While conventional lenses heavily attenuate evanescent waves, preventing them from reaching the image plane, Pendry showed that a left-handed (LH) medium could reconstruct the amplitude of these evanescent waves, enabling subwavelength resolution [I.6].

I.15.2. Overcoming the Diffraction Limit

Superlenses leverage the unique properties of metamaterials to overcome the diffraction limit, a fundamental constraint in conventional optics. Ramakrishna and others have demonstrated that superlenses can achieve resolutions beyond those of ordinary microscopes [I.62]. Traditional optical materials transmit only the propagating components of light, while non-propagating components and evanescent waves are lost. These evanescent waves carry fine details of an object's structure, but they decay rapidly and do not contribute to the image formed by conventional lenses.

One approach to improving resolution is to increase the refractive index of the lens material. However, this is limited by the availability of high-index materials. Metamaterial superlenses address this limitation by significantly enhancing and recovering evanescent waves, which carry information at subwavelength scales [I.24]. This capability makes superlenses invaluable for applications requiring ultra-high resolution, such as nanoscale imaging and microscopy.

Despite these advancements, no superlens has yet been able to fully reconstruct all evanescent waves emitted by an object. The future challenge lies in designing superlenses capable of capturing and reconstructing all evanescent waves to achieve a perfect image [I.63].

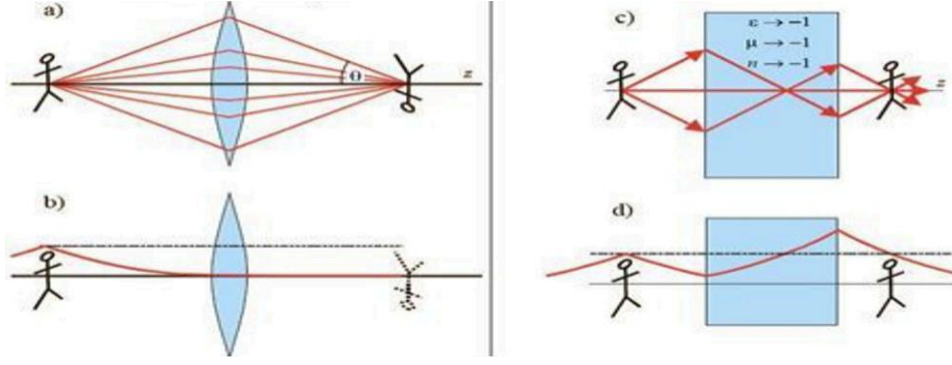


Figure I-25: Case of the classic lens: (a) to focus the waves, it must be convex. (b) The image. Clarity is low, and the evanescent waves are diminishing. Case of the super lens: (c) A planar lens is used to focus the object. (d) The image resolution improves as the evanescent waves amplify in the lens.

Depending on Figure I-25 we can conclude that the focusing happens for the RH lens with incident propagated waves, and the information does not reach the source when focusing the lens for incident evanescent waves. On the other hand, for incident propagated waves on the LH plate, $\epsilon = \mu = -1$, focusing occurs, and the information reaches the source due to a rise in energy within the plate [I.6].

I.15.3. Modeling Microwave Components

In recent years, metamaterials have been widely utilized to design and model advanced microwave components, offering significant advantages over traditional designs. These components include antennas [I.64], absorbers [I.65], sensors [I.66], filters [I.67], and more. One of the key benefits of metamaterial-based devices is their compact size. For example, metamaterial phase shifters are much smaller than their conventional counterparts. Additionally, left-handed (LH) transmission lines provide enhanced coupling at comparable spacing, making them ideal for couplers and other coupling-based devices.

Metamaterial components are characterized by their ability to exhibit simultaneously negative permittivity ϵ and permeability μ . The design of these structures often relies on split-ring resonators (SRRs) and complementary split-ring resonators (CSRRs). These resonators are typically integrated into various transmission lines, such as coplanar waveguides and microstrip lines, to achieve miniaturization and improved electromagnetic properties. In some cases, the reconfigurability of devices is also achieved using these structures.

The focus of this thesis is on microwave filters. Depending on the design, geometry, and parametric architecture of the metamaterial structures, various types of filters can be realized, including: stopband filters, bandpass filters, ultra-wideband filters, and high-selectivity filters [I.68].

I.15.4. Cloaking Metamaterials

One of the most fascinating applications of metamaterials is the concept of invisibility cloaking. This technology aims to render an object invisible by surrounding it with a network of metamaterials that deflect electromagnetic waves, reconstructing the wavefronts downstream of the object. This approach has significant potential, particularly in defense applications for stealth and camouflage.

The first experimental realization of an invisibility cloak was demonstrated by Schurig et al. [I.69], as shown in Figure I-26 (a). The results of this study clearly illustrate how electromagnetic radiation (denoted by black lines) is not reflected by the cloaked object (Figure I-25 (b)). Furthermore, the wave propagation in front of and behind the object remain nearly identical, effectively achieving invisibility.

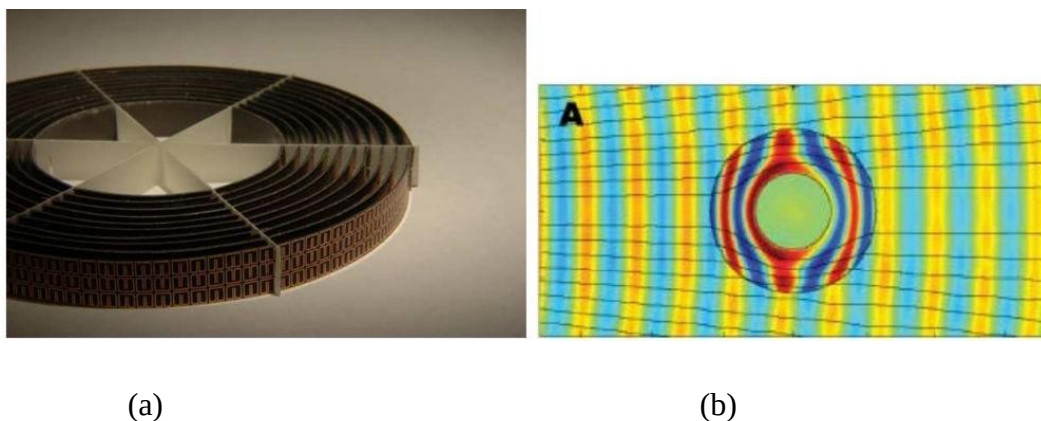


Figure I-26: (a) A photo of the first metamaterial cloaking device. (b) The simulation result of the cloaking device.

I.16. Conclusion

This chapter has provided an in-depth exploration of metamaterials, from their theoretical foundations to their groundbreaking applications. We began by defining metamaterials as artificially engineered structures capable of exhibiting electromagnetic properties not found in nature, particularly negative permittivity (ϵ) and permeability (μ). We then classified materials based on the signs of ϵ and μ , and discussed the implications of these properties on Maxwell's

equations, emphasizing the significance of a negative refractive index as a defining characteristic of metamaterials.

We delved into the fundamental concepts of transmission lines, highlighting the distinctions between right-handed (RH), left-handed (LH), and composite right/left-handed (CRLH) transmission lines. These structures serve as the building blocks for designing advanced microwave components, enabling miniaturization and enhanced performance in devices such as antennas, filters, and couplers.

Furthermore, we explored the realization of metamaterials with negative ϵ or μ , detailing the design and functionality of structures like split-ring resonators (SRRs) and Swiss rolls. These innovations paved the way for the development of doubly negative materials, where both ϵ and μ are negative, leading to the creation of the first practical metamaterials.

The chapter also showcased several transformative applications of metamaterials, including:

- **Superlenses:** Overcoming the diffraction limit to achieve subwavelength resolution, revolutionizing fields such as nanoscale imaging and microscopy.
- **Microwave components:** Enhancing the performance of devices like antennas, filters, and phase shifters through compact and efficient designs.
- **Invisibility cloaks:** Demonstrating the potential of metamaterials to manipulate electromagnetic waves, rendering objects invisible and opening new possibilities in stealth technology.

Despite these remarkable advancements, challenges remain, such as minimizing losses, improving efficiency, and expanding the operational bandwidth of metamaterials. However, the rapid progress in this field promises continued innovation, with the potential to redefine the boundaries of science and technology.

In conclusion, metamaterials represent a paradigm shift in electromagnetics and optics, offering unprecedented control over electromagnetic waves and enabling applications that were once considered science fiction. As research and development in this field continue to advance, we can anticipate even more groundbreaking discoveries that will shape the future of technology.

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CHAPTER II:

Microwave Filters: Background and Theory

II.1 Introduction

The continuous development of science and technology has resulted in major improvements in the design and miniaturization of passive microwave components, including filters, antennas, and duplexers. These components play a vital role in modern communication systems, as they improve electromagnetic performance while enabling reductions in overall device dimensions. In recent years, numerous tunable devices with novel geometries and design techniques have been introduced. Such structures provide an effective approach for the realization of compact integrated circuits, allowing several components to be replaced by a single device while preserving, or even enhancing, their electromagnetic performance [II.1].

In this chapter, the evolution of the electromagnetic spectrum and the applications associated with its different frequency bands will first be discussed. Next, an overview of microwave transmission line theory and its corresponding models will be presented [II.2]. Furthermore, filter theory will be analyzed, with emphasis on their functions, classifications, and the various approximations employed in their implementation [II.3]. Finally, a detailed review and comparative analysis of several recently proposed reconfigurable microstrip filters will be provided [II.4].

II.2. The Electromagnetic Spectrum

II.2.1. History

In 1831, the British scientist Faraday conducted a series of fundamental experiments that led to the discovery of electromagnetic induction. Through these experiments, he demonstrated that a varying magnetic field can generate an electric field. Faraday also formulated the law of electromagnetic induction, which quantitatively describes this phenomenon. He observed that when the magnetic field of an electromagnet was expanded and reduced by opening and closing the electrical circuit connected to it, an electric current could be detected in a nearby conductor. Likewise, a current could be induced in a coil of wire when a permanent magnet was moved in and out of it. Furthermore, when a conductor was moved close to a stationary permanent magnet, an electric current flowed through the wire as long as the conductor remained in motion. Later in his life, Faraday introduced the idea that electromagnetic forces are not limited to conductors but also extend into the surrounding space [II.5].

In 1855, James Clerk Maxwell began his studies on electromagnetism. Building upon the work of earlier scientists and applying advanced mathematical analysis and physical insight, he published three influential papers: On Faraday's Lines of Force, On Physical Lines of Force, and A Dynamical Theory of the Electromagnetic Field. These works provided a systematic and comprehensive description of electromagnetic phenomena while synthesizing the discoveries of previous researchers [II.6]. After being revised and compiled, these papers formed the foundation of classical electromagnetic theory.

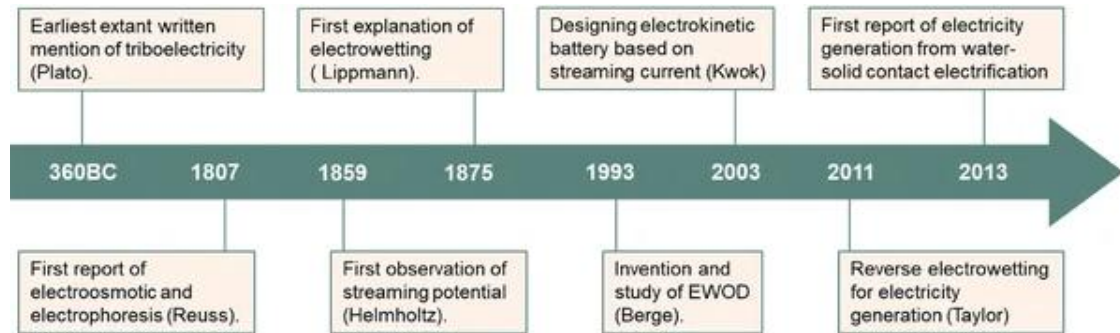


Figure II-1: Chronology of discoveries and developments in electromagnetism.

In 1865, Maxwell predicted the existence of electromagnetic waves and demonstrated theoretically that they propagate as transverse waves at a speed equal to that of light. He also established the fundamental relationship between light and electromagnetic phenomena, proving that light itself is a form of electromagnetic radiation. Later, in 1873, Maxwell published the well-known General Theory of Electromagnetism, which strongly influenced the German physicist Heinrich Rudolf Hertz. Hertz later validated Maxwell's predictions experimentally in 1888 by demonstrating the existence of electromagnetic waves. He further showed that these waves exhibit properties such as reflection, refraction, and polarization, while propagating at the same speed as light [II.7].

From 1888 onward, electromagnetic theory continued to develop and its applications expanded significantly. Electromagnetic waves became an essential natural resource widely utilized in fields such as medicine, daily life, and military technology. In 1895, the Russian scientist Popov developed the first wireless telegraph system, and in the same year Wilhelm Roentgen discovered X-rays. By 1914, voice communication had become possible, and in 1920 commercial radio broadcasting began to emerge. During the 1930s, radar technology was invented, followed by rapid advancements in radar and electromagnetic energy systems

in the 1940s. In the 1950s, the launch of the first artificial satellite marked the beginning of satellite communications, which subsequently experienced rapid development.

II.2.2. Nature of Electromagnetic Waves

Electromagnetic waves can be described as oscillatory disturbances characterized by phase coherence, in which electric and magnetic fields—perpendicular to one another in space—are generated by the emission of oscillating charged particles. These waves propagate as variations in the electromagnetic field and may also exhibit wave–particle duality through the transmission of photons [II.8]. Nevertheless, the complex relationship between the wave and particle characteristics of electromagnetic phenomena remains an important topic of ongoing research, particularly due to the fundamentally different nature of these two aspects.

The work of James Clerk Maxwell on electromagnetism has often been regarded as the “second great unification in physics.” His pioneering research brought together previously separate theories of electricity, magnetism, and optics, representing a major scientific advancement since Isaac Newton’s unification of terrestrial and celestial mechanics. Maxwell’s formulation of classical electrodynamics, particularly in his 1873 publication *A Treatise on Electricity and Magnetism*, established the basis for the modern interpretation of electromagnetic fields. Through the introduction of Maxwell’s equations, he successfully predicted the existence and behavior of electromagnetic waves [II.9].

Electromagnetic waves exhibit several characteristics similar to those of classical waves, including diffraction, interference, reflection, and refraction. However, unlike mechanical waves, they do not require a material medium for propagation. Electromagnetic radiation in the visible spectrum, typically ranging from 380 to 780 nanometers, is emitted by objects with temperatures above absolute zero [II.10]. This radiation, perceived by the human eye as visible light, represents a fundamental physical phenomenon. The emission of such radiation is essential for the conditions that support life, since objects at temperatures equal to or below absolute zero would cease to emit radiation, a state incompatible with the existence of conventional life forms.

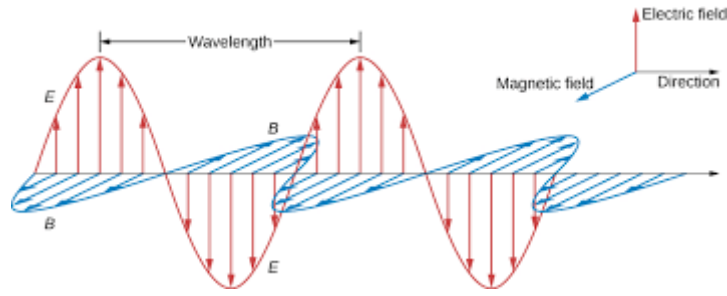


Figure II-2: Example of a plane electromagnetic wave.

II.2.3. Characteristics of Electromagnetic Waves

Electromagnetic waves are defined by three main parameters: intensity, frequency, and waveform. Intensity is associated with the amplitude of the wave and, in the case of visible light, corresponds to the perceived brightness. Frequency is related to the wavelength and determines the perceived color or hue. The waveform, or spectral distribution, is linked to chromaticity. For monochromatic radiation, such as that produced by laser sources, the waveform closely resembles a sinusoidal (cosine) function. The closer the waveform is to an ideal sinusoidal form, the purer the electromagnetic spectrum becomes, resulting in a more monochromatic signal [II.11].

Electromagnetic waves can also be categorized according to their wavelengths, which lead to a wide range of practical applications. These applications extend from low-frequency radio broadcasting and television transmission to high-energy nuclear radiation sources. As a result, electromagnetic waves play an essential role in numerous aspects of modern technology and everyday life.

In classical mechanics, a distinction is typically established between “pure” particles and wave-like fluctuations, often represented by matter and light waves. In contrast, quantum mechanics provides a different perspective. Without explicitly incorporating special relativity, fundamental particles such as photons, electrons, and protons can be described using the Schrödinger equation. The solution of this equation is expressed as a wave function, whose squared magnitude represents the probability density of finding a particle at a particular position. More generally, the wave function can be interpreted as a probability amplitude associated with detecting a particle at a given location. Similar to conventional waves, these probability amplitudes can combine through the principle of superposition, representing

multiple possible paths simultaneously. The electromagnetic spectrum is illustrated in Figure I-3 [II.12].

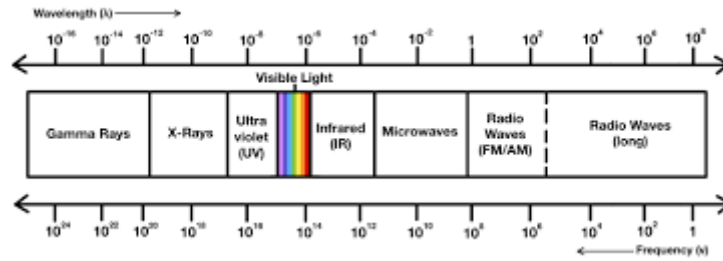


Figure II-3: Types of electromagnetic waves and the electromagnetic spectrum.

Polarization, reflection, and refraction are fundamental properties of electromagnetic waves and have important implications in many scientific and technological domains. Polarization describes the orientation of the electric field oscillations of an electromagnetic wave, which may be linear, circular, or elliptical. This characteristic is widely applied in communication systems, including satellite communications, as well as in optics—for example in polarized lenses used to reduce glare and in astronomy for analyzing radiation emitted by celestial objects [II.13].

Reflection occurs when electromagnetic waves encounter a boundary and return to their original medium. This phenomenon follows the fundamental principle that the angle of incidence is equal to the angle of reflection. Reflection plays a crucial role in various technologies, including radar systems, antenna design, and the development of optical instruments such as telescopes and microscopes.

Refraction refers to the change in direction experienced by electromagnetic waves when they pass from one medium to another due to a variation in propagation velocity. This behavior is described by Snell's law. Refraction is extensively exploited in optical systems, including the design of corrective lenses and telescopes, as well as in communication technologies such as fiber-optic transmission. Its practical importance extends to many applications, ranging from optical component manufacturing to the development of modern fiber-optic communication networks.

II.2.3.1. Field Configuration

The electric field $\vec{E}$ and magnetic field $\vec{H}$ associated with waves involved in the transmission–reflection (TR) process can be expressed in a unified mathematical form suitable for analysis [II.14, II.15]. Consider the propagation of a monochromatic electromagnetic wave with angular frequency ω , where the temporal variation of the electric and magnetic fields is represented by as $\text{Re} [E \exp (-i \omega t)]$, $\text{Re} [H \exp (- i \omega t)]$, respectively.

Assume that the incident plane wave originates from the lower half-space ($z < 0$) (medium 1), which is characterized by dielectric permittivity ϵ_1 magnetic permeability μ_1 and refractive index $n_1 = \sqrt{\mu_1 \epsilon_1}$, The interface separating the two media is located at ($z = 0$), with medium 2 occupying the region ($z > 0$) and defined by parameters ϵ_2 , μ_2 and $n_2 = \sqrt{\mu_2 \epsilon_2}$ (see Figure I-4).

The propagation direction of the wave is described by the wave vector $k_1 = k_1(\sin\theta_1, 0, \cos\theta_1)^T$, θ_1 is the angle of incidence. The wavenumbers in media 1 and 2 are given by $k_1 = n_1(\omega/c)$ and $k_2 = n_2(\omega/c)$ respectively, with (c) representing the speed of light in vacuum. The refractive index is ($n = 1$) in vacuum, approximately 1.00029 in air under normal atmospheric conditions, about 1.33 in water, and typically ranges between 1.5 and 1.7 in glass.

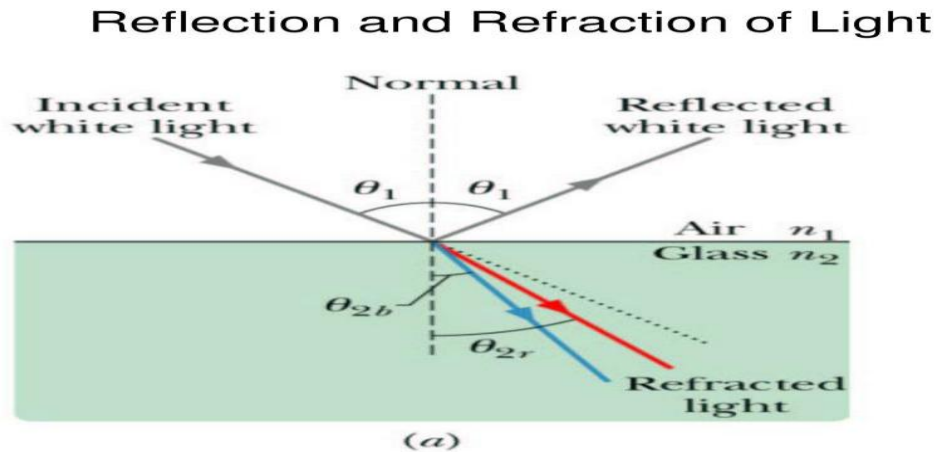


Figure II-4: Refraction of an electromagnetic wave during propagation from medium 1 to medium 2.

The behavior of refraction is governed by Descartes' laws of refraction:

- **First law:** The refracted ray lies in the plane of incidence.
- **Second law:** The relationship between the angle of incidence θ_1 and the angle of refraction θ_2 is given by:

$$n_1 \sin(\theta_1) = n_2 \sin(\theta_2) \quad (\text{II.1})$$

II.2.4. Field of Application of Electromagnetic Waves

The electromagnetic (EM) spectrum encompasses the entire range of electromagnetic radiation, which can be described either in a particle form as photons or in a wave form as electromagnetic waves. It is generally represented using frequency, wavelength, or energy scales. The spectrum theoretically extends continuously from zero to infinity in terms of frequency (or wavelength), without any discontinuity. It includes various types of electromagnetic radiation, each with specific designations, frequency bands, and practical applications. Examples include radio waves (also known as Hertzian waves), microwaves, terahertz radiation, infrared radiation, visible light, ultraviolet radiation, X-rays, and gamma rays [II.16].

Starting from the highest-energy radiation and moving toward lower energies, the electromagnetic spectrum can be categorized as follows:

- **Gamma rays (γ):** These are extremely high-energy radiations produced primarily by the radioactive decay of unstable atomic nuclei. They possess the shortest wavelengths in the electromagnetic spectrum, typically ranging from 10^{-14} m to 10^{-12} m.
- **X-rays:** Widely used in the medical field for diagnostic imaging, X-rays also have applications in airport baggage inspection and various scientific studies related to the structure of matter. Their wavelengths generally range from 10^{-12} m and 10^{-8} m.
- **Ultraviolet (UV) radiation:** Most ultraviolet radiation originates from natural sources such as the Sun and other stars. These rays carry relatively high energy and have wavelengths approximately between 10^{-8} m to $4 \cdot 10^{-7}$ m.
- **Visible light:** This portion represents the small segment of the electromagnetic spectrum that can be detected by the human eye. Its wavelength range typically extends from about 400 nm to 800 nm.

- **Infrared radiation:** All objects with a temperature above absolute zero emit infrared radiation, even if it is not visible to the human eye. Infrared waves are commonly used in applications such as measuring the temperature of land surfaces, oceans, and clouds. Their wavelengths generally range from approximately $8 \cdot 10^{-7}$ m to a millimeter (10^{-3} m) to one meter
- **Radio waves:** This region constitutes the broadest portion of the electromagnetic spectrum and corresponds to the longest wavelengths, which can extend from a few meters to several kilometers. Radio waves are widely used for communication purposes because they can be easily generated and received, making them suitable for technologies such as radio broadcasting, television transmission, and telecommunication systems.

All electromagnetic devices and communication systems operate within specific frequency bands, many of which lie within the microwave region of the spectrum.

II.3. Transmission Line Theory and Its Modeling

Transmission line (TL) theory can be viewed as a simplified yet coherent framework derived from general electromagnetic theory. In this approach, the governing equations and their solutions can be interpreted as more manageable representations of the corresponding electromagnetic formulations [II.17]. This simplification is particularly attractive for undergraduate students in electrical engineering, as the vector nature of electromagnetic fields is eliminated. Instead, the propagation of signals along the line can be described using forward and backward scalar voltage waves. The introduction of the Smith chart [II.18], [II.19] further enhances this approach, since it replaces complex analytical calculations involved in transmission line analysis with intuitive geometrical constructions.

However, any simplification tends to obscure the underlying theoretical concepts. This situation often arises when students attempt to relate the theoretical expressions of transmission line theory to the practical graphical solutions provided by the Smith chart [I.20]. In many cases, the relationship between these two aspects is not clearly perceived, leading students to neglect the theoretical framework because the graphical method appears more powerful and convenient. After years of teaching this subject, simple concepts from linear algebra have been introduced to bridge this gap, allowing both the desire for simplicity and

the objective of conveying the deeper physical meaning of the subject to be achieved simultaneously.

Regardless of the specific physical structure, a uniform transmission line—defined as a line with a constant cross-section along its length—can be modeled by an equivalent circuit representation. As illustrated in Figure II-5(a), a small segment of the line of length Δz can be represented by the lumped-element circuit shown in Figure II-5(b), consisting of the following parameters:

- **Resistance (R):** Represents power dissipation due to the finite resistance of the conductors. It is expressed in ohms per unit length.
- **Inductance (L):** Represents the magnetic field generated by the current flowing through the conductors. It is expressed in henries per unit length.
- **Capacitance (C):** Represents the capability of the transmission line to store electric energy between the conductors. It is expressed in farads per unit length.
- **Conductance (G):** Represents energy losses associated with the dielectric material between the conductors. It is expressed in siemens per unit length.

Thus, R, L, G, and C are commonly referred to as the resistance, inductance, conductance, and capacitance per unit length (often abbreviated as p.u.l.). In the metric system, their units are respectively expressed as ohms per meter (Ω/m), henries per meter (H/m), siemens per meter (S/m), and farads per meter (F/m).

The values of these parameters depend on both the geometry of the transmission line and the electrical characteristics of the materials used. Capacitance (C) mainly reflects the dielectric's ability to store electrical energy, while conductance (G) represents dielectric losses caused by conduction and dielectric relaxation phenomena. In most microwave substrates, conductivity is very low, so dielectric relaxation losses tend to dominate. These losses originate from the displacement of charge centers within the dielectric material, which leads to distortions in the dielectric lattice or molecular structure. As the electric field varies periodically, part of the electromagnetic energy is converted into mechanical vibrations within the material.

Resistance (R) mainly arises from ohmic losses in the metallic conductors. Inductance (L), on the other hand, represents the capacity to store magnetic energy and is largely determined by

the geometry of the transmission line, since most materials used in transmission lines have a relative permeability close to unity ($\mu \approx 1$).

For most transmission lines, the effects of inductance (L) and capacitance (C) dominate because the series resistance and shunt conductance are relatively small. Consequently, the propagation characteristics of a line are often analyzed using an idealized lossless model, even though practical calculations of power losses still require knowledge of R and G.

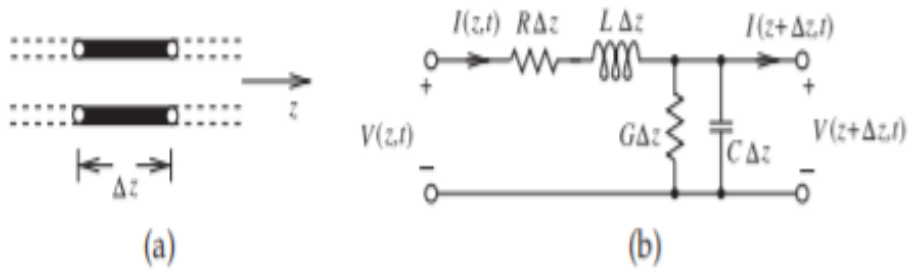


Figure II-5: Transmission line segment: (a) of length Δz ; and (b) lumped-element model.

Several types of transmission lines have been developed to satisfy different practical requirements, as illustrated in Figure II-6:

- **Stripline:** This structure is relatively bulky because it requires two ground planes.
- **Coplanar Waveguide (CPW)** [II.22]: Consists of three metallic strips separated by two slots, with a central signal conductor isolated from two ground conductors by air gaps. A variation of this configuration is the grounded coplanar waveguide (GCPW), which includes an additional ground plane beneath the substrate [II.23].
- **Slotline** [II.24]: Formed by two conductors deposited on the same side of a dielectric substrate, creating a slot that guides the electromagnetic wave.
- **Microstrip line** [II.25]: Comprises a dielectric substrate with a metallized ground plane on the backside and a conducting strip forming the circuit on the front side.

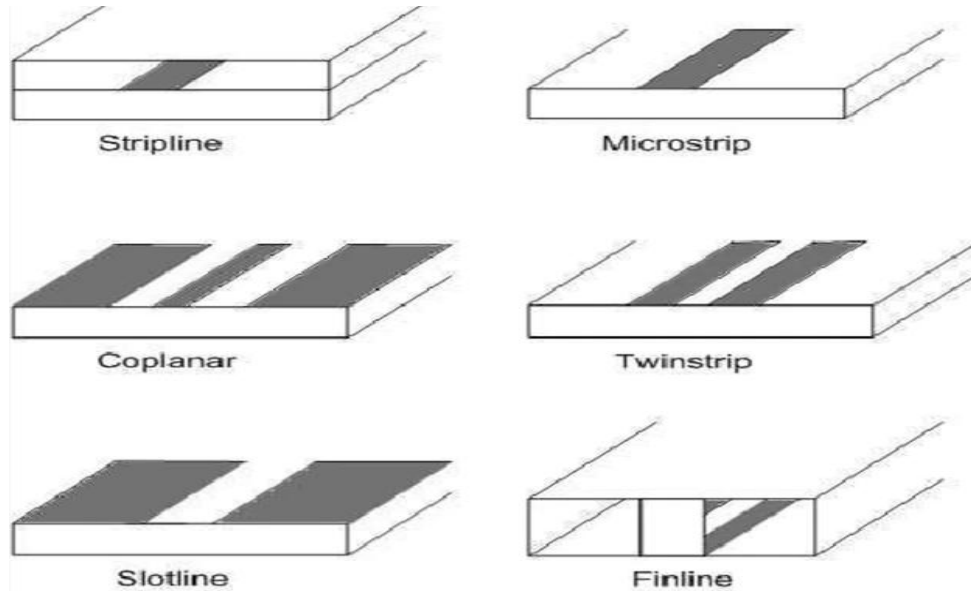


Figure II-6: The different types of planar lines.

In this work, particular attention will be given to microstrip lines and coplanar waveguides, as they will be used to excite the proposed systems. At high frequencies, these transmission lines require specific design considerations to avoid the excitation of parasitic modes. Compared with waveguides, their main limitations are the maximum power they can handle and the insertion losses introduced during signal propagation.

II.3.1. Derivation of Transmission Line Properties

In this section, the differential equations that describe signal propagation along a transmission line are derived. These equations consist of two coupled first-order differential equations and represent the one-dimensional equivalent of Maxwell's equations. Solving these equations makes it possible to understand how electromagnetic signals propagate along the line and allows the extraction of key parameters that characterize transmission line behavior.

By applying Kirchhoff's voltage and current laws to the equivalent circuit model shown in Figure I-7 and considering the limit as $\Delta z \rightarrow 0$, the transmission line equations can be obtained

$$\frac{\partial V(z, t)}{\partial z} + L \frac{\partial I(z, t)}{\partial z} + RI(z, t) = 0 \quad (\text{II.2})$$

$$\frac{\partial I(z, t)}{\partial z} + C \frac{\partial V(z, t)}{\partial z} + GV(z, t) = 0 \quad (\text{II.3})$$

Here, $V(z)$ is a phasor and $V(z, t) = \Re\{V(z)e^{j\omega t}\}$, where $\Re\{\omega\}$ denotes the real part of a complex number.

In sinusoidal steady-state using cosine-based phasors, these become:

$$\frac{\partial V(z)}{\partial z} = -(R + j\omega L)I(z) \quad (\text{II.4})$$

And

$$\frac{\partial I(z)}{\partial z} = -(G + j\omega C)V(z) \quad (\text{II.5})$$

By eliminating the current term from the above expressions, the wave equation for the voltage along the transmission line can be derived:

$$\frac{\partial^2 V(z)}{\partial z^2} - \gamma^2 V(z) = 0 \quad (\text{II.6})$$

Similarly

$$\frac{\partial^2 I(z)}{\partial z^2} - \gamma^2 I(z) = 0 \quad (\text{II.7})$$

Where the propagation constant is:

$$\gamma = \alpha + j\beta \quad (\text{II.8})$$

where α is the attenuation constant, measured in nepers per meter (Np/m), and β is the phase constant, measured in radians per meter (rad/m). Although nepers and radians are dimensionless quantities, they are used to indicate attenuation and phase variation along the transmission line, respectively [I.21] Figure II-7: Equivalent circuit of a transmission line.

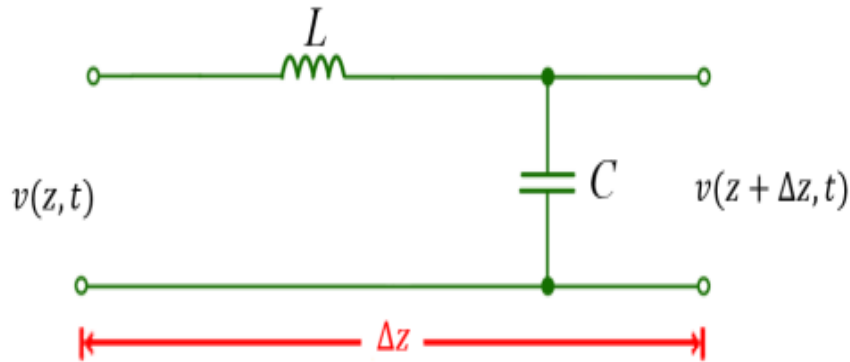


Figure II-7: Equivalent circuit of a transmission line.

II.3.2. Coplanar Lines

The coplanar waveguide (CPW) consists of a central conductor flanked by two ground planes, all positioned on the same plane atop a dielectric substrate, with a defined separation between the conductor and the grounds [II.26].

II.3.2.1. Structure

- **Central conducting strip:** The main conductor responsible for carrying the signal.
- **Ground planes:** Two conductive regions placed on either side of the central strip, providing a return path for current.
- **Dielectric substrate:** The insulating layer supporting the central strip and ground planes [II.1].

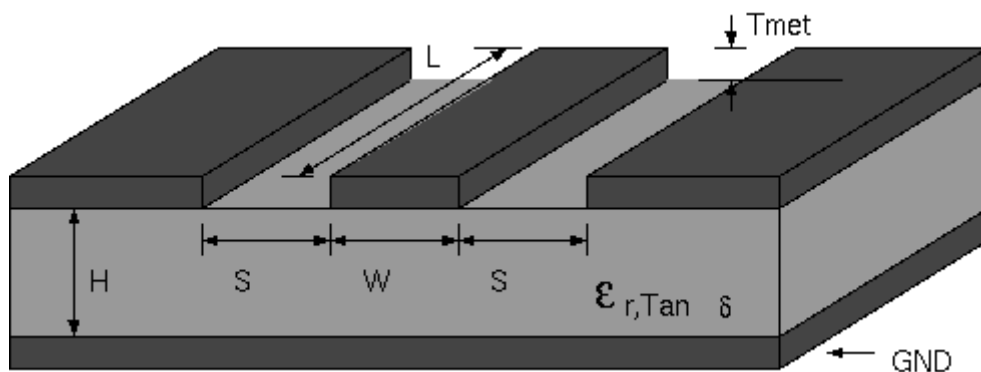


Figure II-8: A cross section of a coplanar line

Coplanar lines are extensively employed as transmission lines in monolithic microwave integrated circuits (MMICs) due to their favorable electromagnetic and fabrication characteristics. With the trend toward circuit miniaturization and higher operating frequencies, MMIC-based devices have become increasingly relevant. In particular, coplanar lines are of interest in transmission line-based characterization techniques, as they can serve directly as measurement cells. This configuration enables straightforward electromagnetic characterization of thin-film materials, since the sample can be integrated directly with the coplanar structure.

To analyze the electromagnetic behavior of a CPW, its S-parameters—reflection parameters (S_{11}), (S_{22}) and transmission parameters (S_{12}), (S_{21})—are measured across the desired frequency range, up to 110 GHz. These measurements, performed using a vector network analyzer, allow the extraction of the propagation constant and characteristic impedance of the coplanar line. From these quantities, the electromagnetic properties of the substrate or thin-film material situated between the line and the substrate can be deduced.

The coplanar line offers several advantages: it supports broadband characterization, is suitable for thin-film measurements, is relatively simple to fabricate, and has low manufacturing cost. Moreover, it operates in a quasi-TEM mode with minimal dispersion, despite the inhomogeneity of the medium (air + substrate). In this mode, the electric and magnetic fields are approximately perpendicular to the propagation axis. The line dimensions can be optimized to ensure propagation of the dominant quasi-TEM mode, improving the accuracy of the measured S-parameters [II.27].

To validate the analytical model used for extracting electromagnetic properties, coplanar lines were fabricated on an Alumina substrate, with specific material characteristics as reported in the literature [II.28].

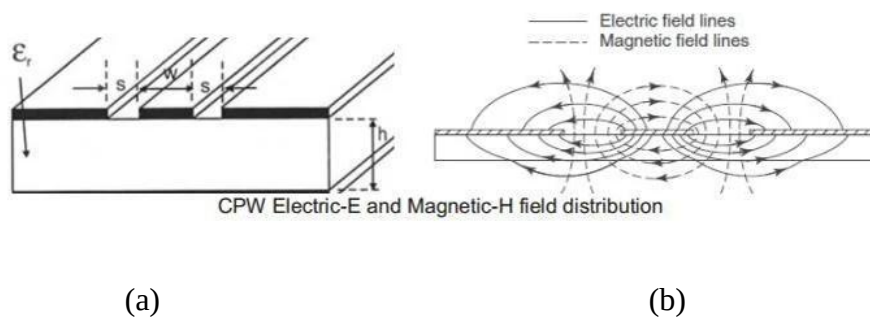


Figure II-9: Coplanar line (a) and distribution of electric and magnetic field lines (b).

II.3.2.2. Applications

Coplanar lines are used in several microwave and communication applications. They can be employed in the design of filters [II.29], impedance matching networks, power dividers, and directional couplers. In addition, coplanar lines are utilized in choke circuits, where they help induce polarization effects in active components. Another important application is their use as measurement cells, since the structure can directly integrate the sample material, enabling simple and non-destructive electromagnetic characterization of thin-film materials [II.28].

II.3.2.3. Advantages

Coplanar lines offer several significant advantages, including the ability to perform broadband characterization in a non-destructive manner and their suitability for thin-film structures. Furthermore, coplanar waveguides are relatively easy to fabricate and involve low technological complexity and cost. They typically operate in a quasi-transverse electromagnetic (quasi-TEM) mode, which exhibits limited dispersion despite the inhomogeneous propagation medium composed of air and substrate materials. In this propagation mode, the electric and magnetic fields are approximately perpendicular to the axis of signal transmission.

Figure II-10 illustrates a coplanar line structure along with its equivalent lumped electrical circuit model for an infinitesimal line segment. The physical dimensions of the line can be optimized to ensure propagation of the dominant quasi-TEM mode, which generally requires the substrate thickness to be greater than the central conductor width plus twice the slot separation distance. Such optimization also contributes to improving the precision of S-parameter measurements [II.30].

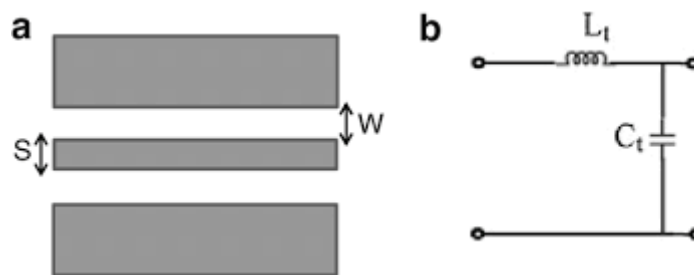


Figure II-10: Coplanar waveguide (a) and its equivalent electrical circuit (b).

II.3.3. Microstrip Lines

The microstrip line is among the most widely used planar transmission lines due to its compatibility with photolithographic fabrication, ease of miniaturization, and seamless integration with both passive and active microwave components. The basic geometry of a microstrip line is illustrated in Figure II-11(a), where a conductor of width (W) is printed on a grounded dielectric substrate of thickness (d) and relative permittivity ϵ_r . The corresponding electric and magnetic field lines are depicted in Figure II-11(b).

If the dielectric substrate were absent ($\epsilon_r = 1$), the structure would reduce to a two-wire line: a flat strip conductor over a ground plane in a homogeneous medium (air). This configuration supports a pure TEM wave with phase velocity $v_p = c$ and propagation constant $\beta = k_0$.

The introduction of a dielectric substrate, particularly the fact that the dielectric does not occupy the region above the strip ($y > d$), complicates the line's electromagnetic behavior. Unlike stripline, where the entire field is confined within a homogeneous dielectric, the microstrip line has most of its field lines in the dielectric between the conductor and the ground plane, while a portion extends into the air above the substrate. As a result, a microstrip line cannot support a pure TEM wave, since the phase velocity in the dielectric differs from that in air, preventing exact phase matching at the dielectric–air interface.

In reality, the microstrip fields form a hybrid TM-TE wave, requiring advanced analysis techniques beyond the scope of this work. However, in practical applications where the dielectric thickness is electrically thin ($d \ll \lambda$), the fields can be approximated as quasi-TEM, meaning they closely resemble the static (DC) field distribution. This approximation allows accurate calculation of phase velocity, propagation constant, and characteristic impedance using quasi-static methods.

Under this quasi-TEM assumption, the phase velocity and propagation constant can be expressed as:

$$v_p = \frac{c}{\sqrt{\epsilon_e}} \quad (II.9)$$

$$\beta = k_0 \sqrt{\epsilon_e} \quad (II.10)$$

Here, ϵ_e is the effective dielectric constant of the microstrip line. Since some field lines lie in the dielectric and others in air, the effective dielectric constant satisfies:

$$1 < \epsilon_e < \epsilon_r \quad (II.11)$$

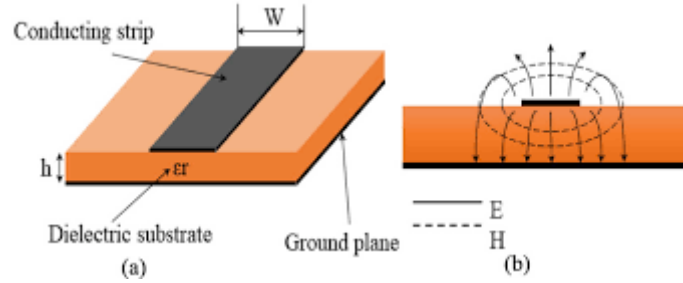


Figure II-11: Microstrip transmission line. (a) Geometry. (b) Electric and magnetic field lines.

II.4. Modeling of the Transmission Line

Transmission line modeling involves representing the distributed nature of electrical transmission lines using lumped-element circuits. A real transmission line has parameters such as resistance, inductance, capacitance, and conductance continuously distributed along its length.

The standard transmission line model consists of series resistance (R) and inductance (L) combined with parallel capacitance (C) and conductance (G). The series resistance and parallel conductance are responsible for the **losses** in the transmission line [II.31].

II.4.1. Distributed Parameter Model

For a lossy transmission line, the line can be represented using distributed parameters: resistance (R), inductance (L), capacitance (C), and conductance (G) per unit length.

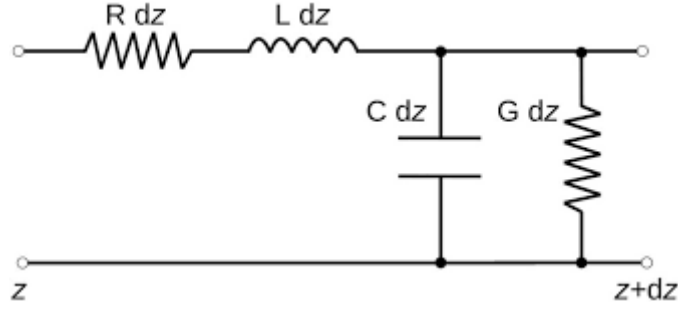


Figure II-12: The distributed circuit for a transmission line modeled by a resistance (R), inductance (L), capacitance (C), and conductance (G) per unit length.

In general, these parameters may vary with frequency. By applying Kirchhoff's voltage and current laws to the infinitesimal segment shown in Figure II-12, the Telegrapher's equations are obtained:

$$\frac{\partial}{\partial x} \tilde{V}(x) = -L \frac{\partial}{\partial t} \tilde{I}(x) - R \tilde{I}(x) \quad (\text{II.12})$$

$$\frac{\partial}{\partial x} \tilde{I}(x) = -C \frac{\partial}{\partial t} \tilde{V}(x) - G \tilde{V}(x) \quad (\text{II.13})$$

Assuming sinusoidal time dependence:

$$I(x, t) = \tilde{I}(x) = I(x) e^{j\omega t} \quad (\text{II.14})$$

$$V(x, t) = \tilde{V}(x) = V(x) e^{j\omega t} \quad (\text{II.13})$$

Then, we insert (II. 14) into (II. 15), and solve the wave equation to define the propagation constant,

$$\gamma = \sqrt{(R + j\omega L)} \sqrt{(G + j\omega C)} \quad (\text{II.16})$$

And the characteristic impedance,

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (\text{II. 17})$$

where ω is the angular frequency [II.32].

II.4.2. Smith Chart

The Smith chart (Figure II-13) is a widely used graphical tool for solving transmission line problems. While alternative charts for impedance and reflection coefficient exist [II.33], the Smith chart remains the most recognized and employed. It was developed in 1939 by P. Smith at Bell Telephone Laboratories [II.34].

Although modern engineers often rely on personal computers and CAD software, the Smith chart continues to be invaluable. Beyond its graphical nature, it is embedded in many CAD tools and measurement equipment for microwave design, providing a visual method for understanding transmission line behavior without intensive numerical computation. Mastery of the Smith chart allows microwave engineers to develop strong intuition for impedance matching, transmission line phenomena, and related design problems [I.1]

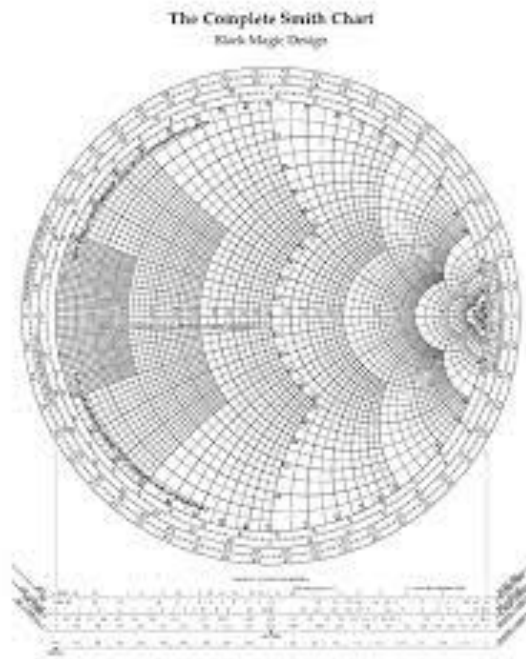


Figure II-13: The Smith chart.

II.5. Microwave Filtering

Filtering is the process of suppressing or enhancing specific frequencies or frequency bands within an electrical signal. Essentially, it involves modifying the spectral content of a signal to retain desired components and reject undesired ones.

Filters are among the most widely used components in telecommunications, with their primary application being frequency multiplexing of signals. The design of microwave filters is particularly challenging because the elements involved exhibit distributed parameters. Unlike lumped-element circuits, there is no universal synthesis method for microwave filters, as the frequency response of components like transmission lines, cavities, and resonators is complex and highly dependent on the operating frequency. Despite these challenges, a variety of techniques have been developed to design effective microwave filters.

Microwave filters can function independently or in combination with other passive devices, such as multiplexers or duplexers, which are commonly used in communication systems. They are also employed in active circuits, including amplifiers and oscillators. Microwave resonators, which operate in a manner similar to tuned circuits with localized elements, find applications in filters, oscillators, frequency meters, tuned amplifiers, and microwave sensors [II.35].

II.5.1. Filter Function

Filters play a critical role in signal processing by removing unwanted components, often considered noise, from a useful signal. Noise can originate externally, from the transmission channel, or internally, from the passive and active components of the system. The signal-to-noise ratio (SNR), defined as the ratio of signal power to noise power, is a key metric in assessing system performance.

Additionally, in many systems, the transmitted and received signals can interfere with each other, making proper signal separation essential. Filters also address issues related to spurious signals or image frequencies, which can arise depending on the system architecture. In both cases, filtering is necessary to ensure accurate and efficient signal processing [II.36, II.37].

II.5.2. Filter Classification

Filters are commonly classified into four main categories based on how they handle different frequency components of a signal: low-pass, high-pass, band-pass, and band-stop [II.38].

- **Low-pass Filters:** Allow frequencies below a specified cutoff to pass while attenuating higher frequencies. (See Figure II-14a for an ideal low-pass response.)

- **High-pass Filters:** Permit frequencies above a cutoff frequency and suppress lower frequencies. High-pass filters are essentially the inverse of low-pass filters, and combining the two can produce a band-pass response. (Figure II-14b)
- **Band-pass Filters:** Pass frequencies within a defined range between a low and high cutoff frequency while rejecting frequencies outside this range. (Figure II-14c)
- **Band-stop Filters (Notch Filters):** Reject frequencies within a specific range, allowing all other frequencies to pass. These are the inverse of band-pass filters. (Figure II-14d)

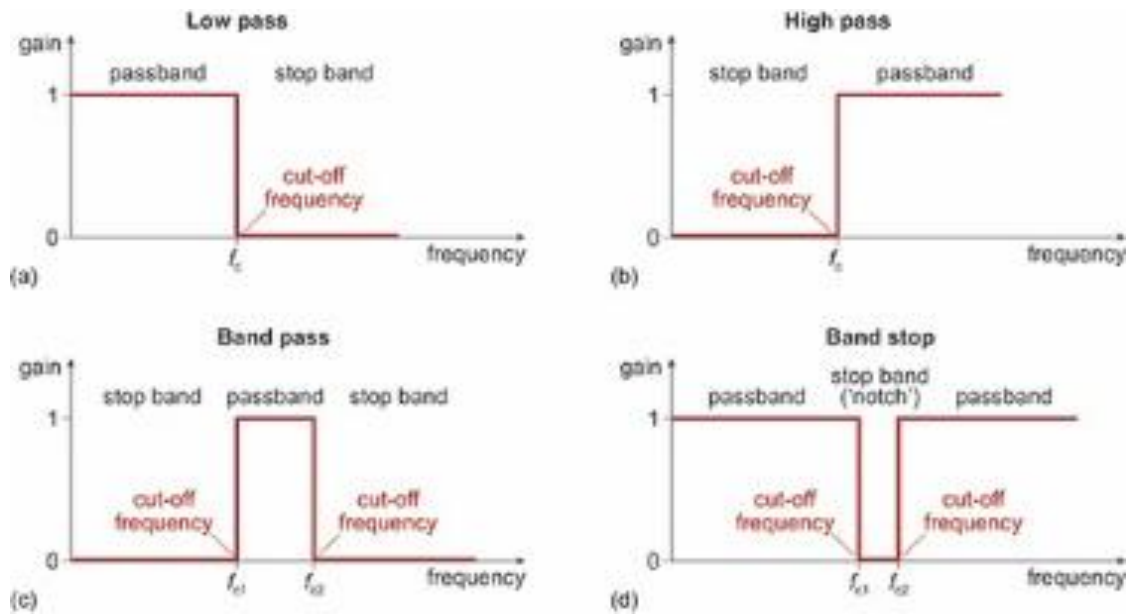


Figure II-14: Ideal filter responses: (a) low-pass (b) high-pass (c) band-pass (d) band-stop.

II.5.3. Filter Families

Filters are broadly classified into two main families:

- **Passive Filters:** These filters are constructed entirely from resistors (R), inductors (L), and capacitors (C). Passive filters, including RC and RLC types, rely on components that store or dissipate electrical energy and do not require an external power supply. They are widely used in high-frequency applications [II.39].
- **Active Filters:** Active filters incorporate active components such as transistors or operational amplifiers. Unlike passive filters, active filters consume power from an external source due to the inclusion of amplifying elements, enabling signal amplification alongside frequency selection.

II.5.3.1. Filter Realization Methods

The method for implementing a filter depends on whether amplification is needed and the target cutoff frequency. Common technologies include:

- **Passive RLC Filters:** Utilize combinations of resistors, inductors, and capacitors to achieve the desired filtering behavior. They can be applied to both low-frequency and high-frequency ranges depending on the components used [II.40].
- **RC and LC Filters:** Employ resistor-capacitor or inductor-capacitor networks tailored for specific frequency ranges and applications [II.41].
- **Quartz Crystal Filters:** Leverage the resonant properties of quartz crystals to pass selected frequencies while attenuating others [II.42].
- **Distributed-Element Filters:** In these filters, circuit elements such as capacitance, inductance, and resistance are distributed along the structure rather than being concentrated in discrete components, enabling operation at microwave and higher frequencies [II.43].

II.5.3.2. Criteria for Selecting Filter Structures

Several criteria must be considered when choosing the appropriate technology and topology for filter design. These criteria include:

- **Electrical criteria:** These relate to the cutoff frequency, passband characteristics, rejection band attenuation, in-band ripple, passband bandwidth, and out-of-band suppression levels.
- **Physical criteria:** These involve the size, weight, and volume of the device, as well as mechanical stability, temperature sensitivity, and power-handling capability.
- **Economic criteria:** These consider the overall cost associated with design, fabrication, and tuning processes, in addition to compatibility with relevant technical standards.

II.5.4. Approximation Functions

Filter design employs various mathematical approximations to achieve desired frequency responses. The most common approximations are Butterworth, Chebyshev, and Elliptic.

II.5.4.1. Butterworth Approximation

The Butterworth approximation is the simplest filter approximation method. Butterworth filters exhibit a maximally flat passband response, meaning that no ripple is present within the passband. However, their out-of-band attenuation is relatively weak compared to other approximation types. This limitation can be partially mitigated by increasing the filter order, although the rejection performance generally remains inferior to that of more selective filter designs. Butterworth filters are typically used in applications requiring low insertion loss [II.44].

The attenuation function expressed in decibels is given by

$$A_{db} = 10 \log_{10} \left[1 + \left(\frac{\omega}{\omega_c} \right)^{2n} \right] \quad (\text{II.18})$$

This corresponds to the amplitude response of the transmission coefficient:

$$|S_{12}(j\omega)|^2 = \left[1 + \left(\frac{\omega}{\omega_c} \right)^{2n} \right]^{-1} \quad (\text{II.19})$$

Where:

- $\omega_c = 2\pi f_c$ is the cutoff angular frequency
- n is the filter order.

At $\omega = \omega_c$, the attenuation A is 3 dB, indicating that the output power is half of the input power at the cutoff frequency since $20 \log_{10} |S_{12}| = -3 \text{ dB}$.

Figure I-15 illustrates transmission responses of the Butterworth approximation for different orders n . This approximation yields a very flat passband response, with flatness increasing as filter order rises.

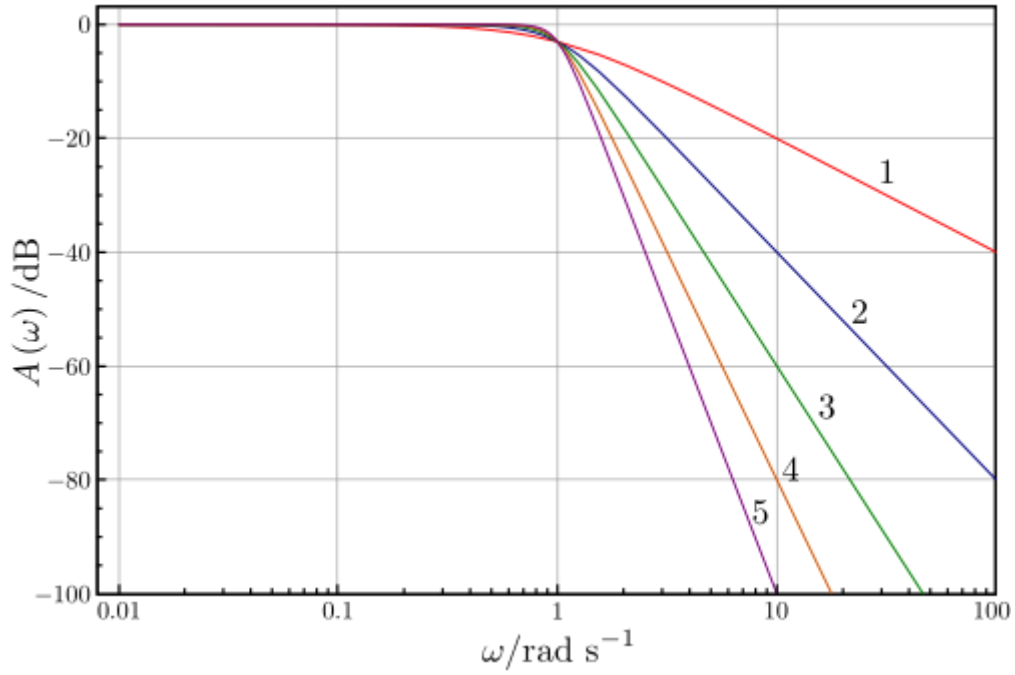


Figure II-15: Transmission responses of the Butterworth approximation for different filter orders n .

II.5.4.2. Chebyshev Approximation

Chebyshev filters tolerate slight passband ripple but provide better rejection than Butterworth filters.

The attenuation function is given by:

$$A_{db} = 10 \log_{10} \left[1 + \varepsilon^2 T_n^2 \left(\frac{\omega}{\omega_c} \right) \right] \quad (\text{II.20})$$

Which corresponds to the amplitude response of the transmission coefficient:

$$\varepsilon = \sqrt{10^{\frac{L_s}{10}} - 1} \quad (\text{II.21})$$

Where:

- T_n is the Chebyshev polynomial of the first kind of order n
- ε relates to the passband ripple level L_s by:

$$\varepsilon = \sqrt{10^{\frac{L_S}{10}} - 1} \quad (\text{II.22})$$

Chebyshev polynomials are defined as:

$$T_n(x) = \begin{cases} \cos(n \cos^{-1}(x)), & |x| \leq 1 \\ \cosh(n \cosh^{-1}(x)), & |x| \geq 1 \end{cases} \quad (\text{II.23})$$

Figure I-16 shows transmission responses of the Chebyshev approximation for different orders n . The passband ripple and sharper cutoff compared to the Butterworth filter are evident.

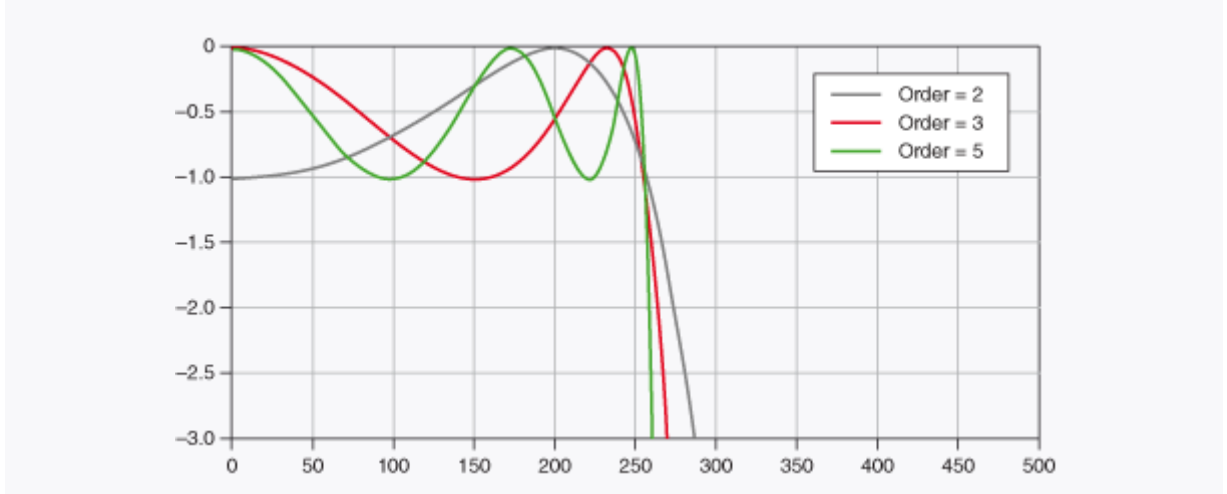


Figure II-16: Transmission responses of the Chebyshev function for different filter orders n .

II.5.4.3. Elliptic Approximation

Elliptic filters exhibit equiripple behavior in both the passband and stopband [II.1], [II.45, II.46]. They also feature transmission zeros in their frequency response, allowing for high selectivity with a relatively low filter order. Despite their efficiency, elliptic filters are more challenging to design and implement physically [II.44]. A detailed treatment of elliptic function theory can be found in references [II.44], [II.47]. The attenuation function is expressed as:

$$A_{db} = 10 \log_{10} [1 + \varepsilon^2 R_n^2(\omega)] \quad (\text{II.24})$$

Where:

- ε is a parameter defining the passband ripple at the cutoff frequency ω_c
- $R_n(\omega)$ is the elliptic rational function of order n

Figure II-17 presents the magnitude response showing parameters used to specify an elliptic filter.

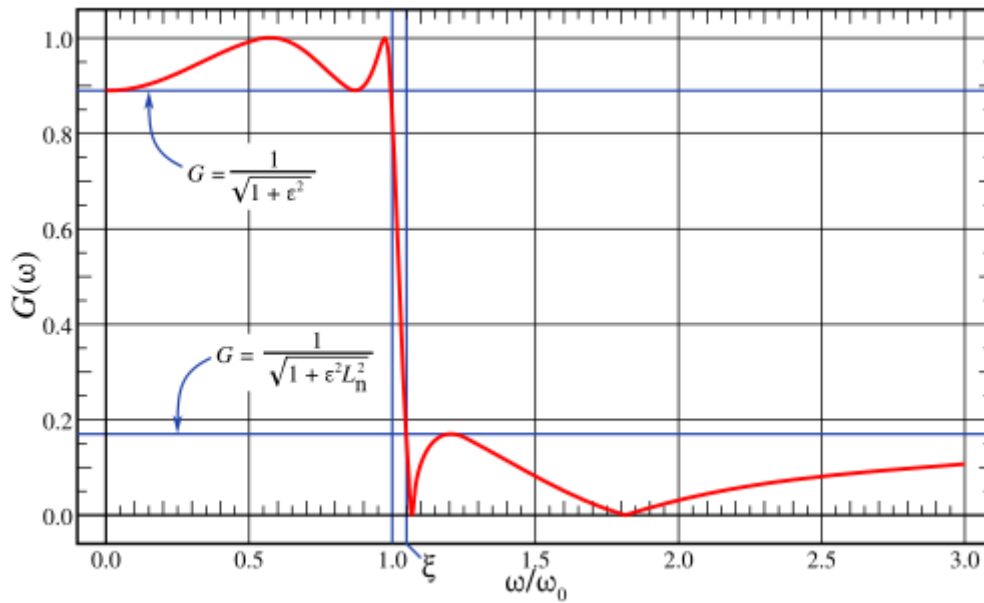


Figure II-17: Magnitude response showing parameters used to specify an elliptic filter.

II.5.5. Band-Pass Filter Specifications

The bandwidth limits of a bandpass filter are typically defined as the frequencies at which the gain decreases by 3 dB relative to the maximum voltage gain, corresponding to approximately half of the maximum output power. These frequencies are referred to as the 3 dB cutoff frequencies. The bandwidth of a bandpass filter is therefore the frequency interval between the lower cutoff frequency FLF_LFL and the upper cutoff frequency FHF_HFH.

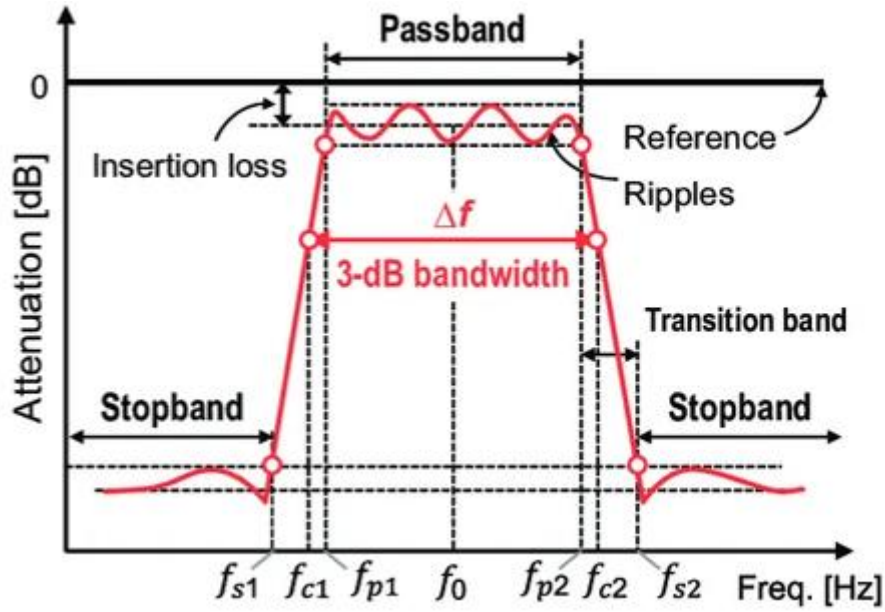


Figure II-18: Frequencies and bandwidth of a band-pass filter.

- **Center frequency (f_c):** The midpoint frequency of the passband.
- **Bandwidth (Δf):** The width of the passband where the filter allows signals to pass with minimal attenuation.
- **Insertion loss:** The reduction in signal power resulting from inserting the filter in the signal path, ideally minimized within the passband.

These electromagnetic parameters can be calculated by transforming a low-pass filter prototype into a band-pass filter through frequency mapping techniques, which will be discussed in the following sections [II.1].

II.5.6. Band-Pass Filter Synthesis

Bandpass filter synthesis involves several fundamental steps. The process begins by determining the corresponding low-pass prototype of the desired filter. Next, the appropriate filter topology is selected based on electrical specifications such as center frequency, bandwidth, and insertion loss. The filter response type must also be chosen, with common options including Butterworth, Chebyshev, elliptic, or pseudo-elliptic responses [II.48]

Figure II-19 shows the equivalent circuit of a low-pass prototype with lumped elements.

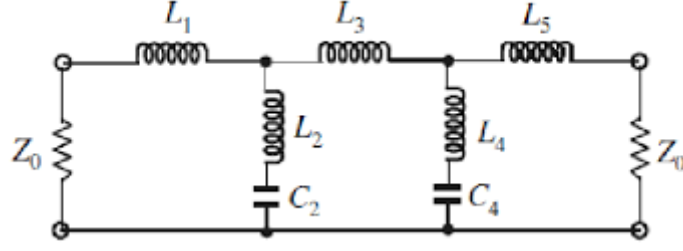


Figure II-19: Equivalent circuit of a low-pass prototype with lumped elements.

II.5.6.1. Low-Pass Prototype Element Values

The element values of a normalized low-pass filter are calculated using classical recurrence formulas:

$$g_0 = 1 \quad (\text{II. 25})$$

$$g_k = \frac{2a_k}{\gamma} \text{ for } K = 1 \quad (\text{II. 26})$$

$$g_k = \frac{4a_{k-1} \cdot a_k}{b_{k-1} \cdot g_{k-1}} \text{ for } K = 2, 3 \dots n \quad (\text{II. 27})$$

$$a_k = \sin \left| \frac{(2K-1)\pi}{2n} \right| \text{ for } K = 1, 2 \dots, n \quad (\text{II. 28})$$

$$b_k = \gamma^2 + \sin^2 \left(\frac{K\pi}{n} \right) \quad (\text{II. 29})$$

$$\gamma = sh \left(\frac{\beta}{2n} \right) \text{ for } K = 1, 2 \dots n \quad (\text{II. 30})$$

$$\beta = Ln \left(coth \frac{A_m}{17.37} \right) \quad (\text{II. 31})$$

A_m is the amplitude of the ripple.

II.5.6.2. Band-Pass Transformation

The design of filters that suppress frequencies outside a specified range while allowing signals within that range to pass is known as bandpass filter synthesis. Bandpass filters can be implemented using several techniques, including prototype-based transformation methods, coupling matrix rotation, and frequency transformation algorithms [I.49].

$$\Delta = \frac{\omega_H - \omega_L}{\omega_0} \quad (\text{II. 32})$$

$$\omega_0 = \sqrt{\omega_L \omega_H} \quad (\text{II. 33})$$

Series elements become series resonant circuits:

$$C'_{bp} = \frac{C}{\omega_0 \Delta} = \frac{g_k}{\omega_0 \Delta} \quad (\text{II. 34})$$

$$L'_{bp} = \frac{\Delta}{C\omega_0} = \frac{\Delta}{g_k\omega_0} \quad (\text{II.35})$$

$$\omega \rightarrow \frac{1}{\Delta} \left(\frac{\omega_0}{\omega_{bp}} - \frac{\omega_{bp}}{\omega_0} \right) \quad (\text{II.36})$$

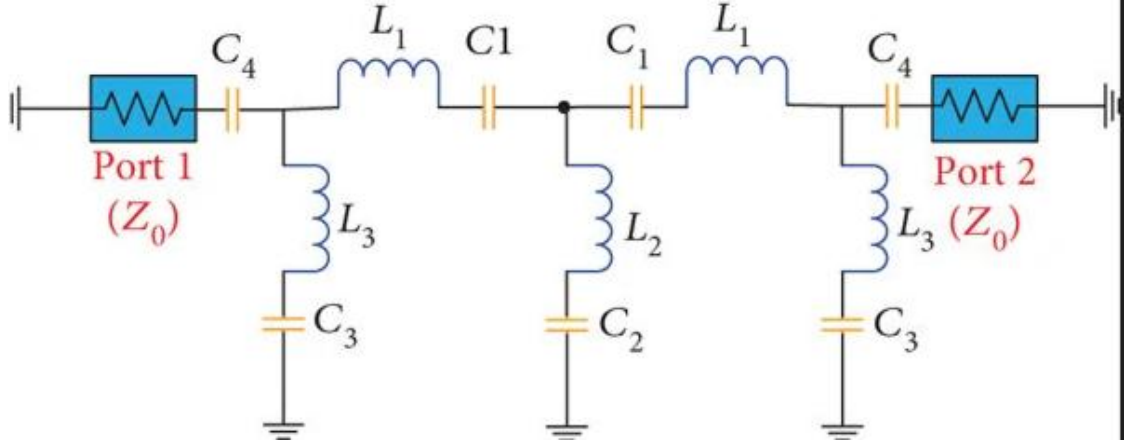


Figure II-20: Equivalent circuit of a lumped-element band-pass prototype.

II.5.7. Distributed Element Filters

At microwave frequencies, filters implemented with discrete (lumped) elements are impractical because the required capacitance and inductance values are extremely small. To overcome this, distributed-element filters use transmission lines themselves as reactive components. The low-pass prototype elements are replaced by equivalent transmission-line resonators and impedance (or admittance) inverters, which define the resonators and the coupling between them.

The synthesis process for a microwave filter proceeds as follows: knowing the values of the low-pass prototype elements, we determine the coupling parameters and the lengths of the resonators, which are usually half-wave ($\lambda/2$) or quarter-wave ($\lambda/4$) lines.

For series resonators connected via impedance inverters, the relationships are given by:

$$K_{01} = \sqrt{\frac{\pi\omega\lambda}{2*\Omega_c*g_0g_1}} \quad (\text{II.37})$$

$$K_{i,i+1} \Big|_{i=1}^{i=n+1} = \frac{\pi\omega\lambda}{2*\Omega_c} \sqrt{\frac{1}{g_i g_{i+1}}} \quad (\text{II.38})$$

$$K_{n,n+1} = \sqrt{\frac{\pi \omega \lambda}{2 * \Omega_c * g_n g_{n+1}}} \quad (\text{II.39})$$

Where:

- $\lambda = \frac{c}{\sqrt{\epsilon} f}$ is the wavelength in the medium.
- θ Represents the electrical length of the resonators.

The dynamic reactance of a series resonator at the central frequency $x_i(\omega_0)$ is given by [I.3]:

$$x_i = \omega_0 L_i = \frac{\omega_0}{2} \frac{dX_i(\omega)}{d\omega} \bigg|_{\omega = \omega_0} \quad (\text{II.40})$$

II.5.7.1. Practical Implementation

In practice, half-wave and quarter-wave resonators are widely used for microwave bandpass filters. Half-wave resonators are suitable for relative bandwidths up to 20%, whereas quarter-wave resonators can handle up to 40%.

The electrical length of a resonator is calculated as:

$$\theta_i = \pi - \frac{1}{2} \left(\tan^{-1} \left(\frac{2X_{i-1,i}}{Z_0} \right) + \tan^{-1} \left(\frac{2X_{i,i+1}}{Z_0} \right) \right) \quad (\text{II.41})$$

where $Z_0 = 50 \Omega$ is the characteristic impedance.

Figure I-21 illustrates an impedance inverter-based band-pass filter circuit.

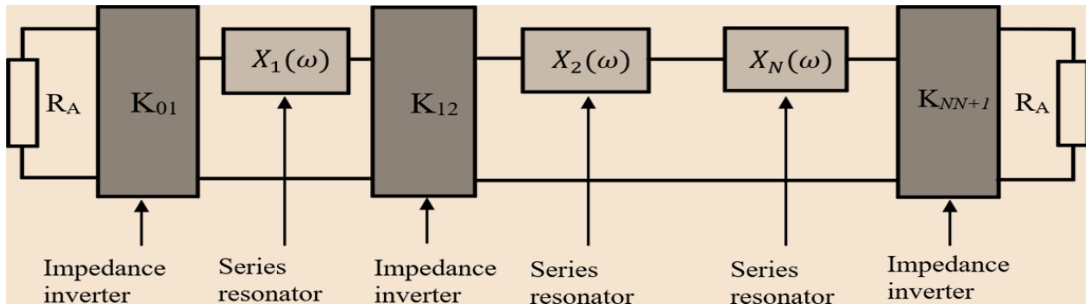


Figure II-21: Impedance inverter-based band-pass filter circuit.

The length of a waveguide cavity is given by:

$$L = \frac{\lambda \theta}{2\pi} \quad (\text{II.42})$$

Where:

$$\lambda = \frac{c}{\sqrt{\epsilon} f} \quad (\text{II.43})$$

II.5.8. Filter Characterization

The performance of a filter depends primarily on the characteristics of its resonators. Several parameters are used to evaluate a resonator's performance based on its electrical response.

II.5.8.1. Cutoff Frequency and Center Frequency

- **Cutoff frequency (f_c):** The frequency at which the filter begins to significantly attenuate the signal, typically defined as the point where the signal amplitude drops by 3 dB.
- **Center frequency (f_0):** The midpoint of the filter's passband, representing the frequency around which the filter operates symmetrically.

II.5.8.2. Insertion Losses

- **Insertion loss** is the attenuation introduced when a filter is placed in a signal path. It reflects losses due to ohmic resistance, dielectric losses, and radiation.
- Minimizing insertion loss is critical for ensuring that the filter passes the desired signals with minimal power degradation.

Figure I-22 illustrates how insertion losses manifest in a resonator and provides insight into evaluating filter performance.

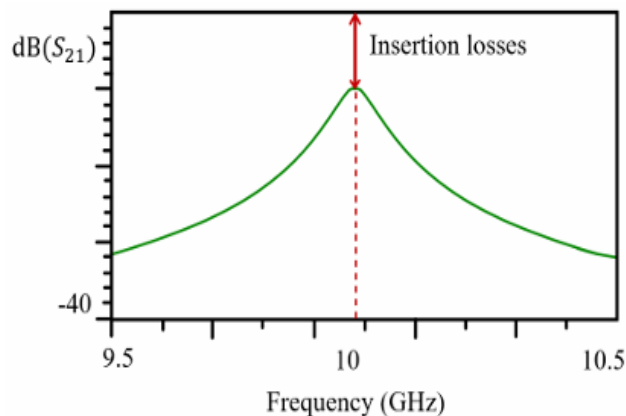


Figure II-22: Illustration of resonator insertion loss.

II.5.8.3. Loaded Quality Factor

The loaded quality factor (Q_L) is a dimensionless metric quantifying resonator selectivity. It is determined by the resonant frequency (f_r) and the -3 dB bandwidth ($f_2 - f_1$) [II.51]:

$$Q_L = \frac{f_r}{f_2 - f_1} \quad (\text{II.44})$$

Figure I-23 demonstrates extraction of parameters f_c , f_0 , and f_1 from the transmission response S_{21} . Higher Q_L indicates better selectivity. However, Q_L does not account for intrinsic resonator performance, as it depends on frequency and does not consider insertion losses.

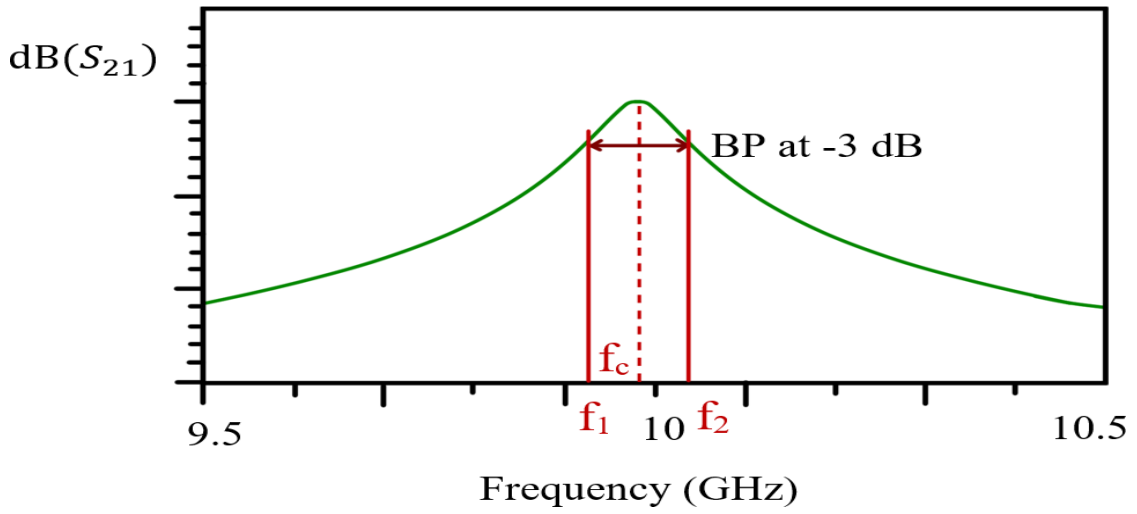


Figure II-23: Parameter extraction for loaded quality factor calculation from resonator transmission response.

II.5.8.4. Unloaded Quality Factor

The unloaded quality factor (Q_0) is influenced by three loss mechanisms and defined as:

$$\frac{1}{Q_L} = \frac{1}{Q_{est}} + \frac{1}{Q_0} \quad (\text{II.45})$$

Losses associated with the resonator excitation mechanism are represented by the external quality factor Q_{ext} , which can be computed as:

$$Q_{est} = \frac{Q_L}{|S_{21}(f_r)|} \quad (\text{II.46})$$

The insertion loss at resonance is denoted by $S_{21}(f_r)$. The unloaded quality factor (Q_0) accurately represents resonator performance, increasing as insertion loss decreases and rejection level increases. From equations (I.45) and (I.46), we obtain [I.51]:

$$Q_0 = \frac{Q_L}{1 - |S_{21}(f_r)|} \quad (\text{II.47})$$

II.6. Reconfigurable Microstrip Filters

Reconfigurable microstrip filters have gained considerable attention in recent years due to their ability to adapt to different operating conditions [II.52, II.61]. These filters allow the center frequency, bandwidth, or other parameters to be dynamically modified without changing the physical structure. A notable example is a reconfigurable microstrip band-pass filter (BPF) using a varactor diode, designed to maintain a constant impedance bandwidth by tuning the resonance frequencies of both odd and even modes independently, with minimal mutual coupling [II.53].

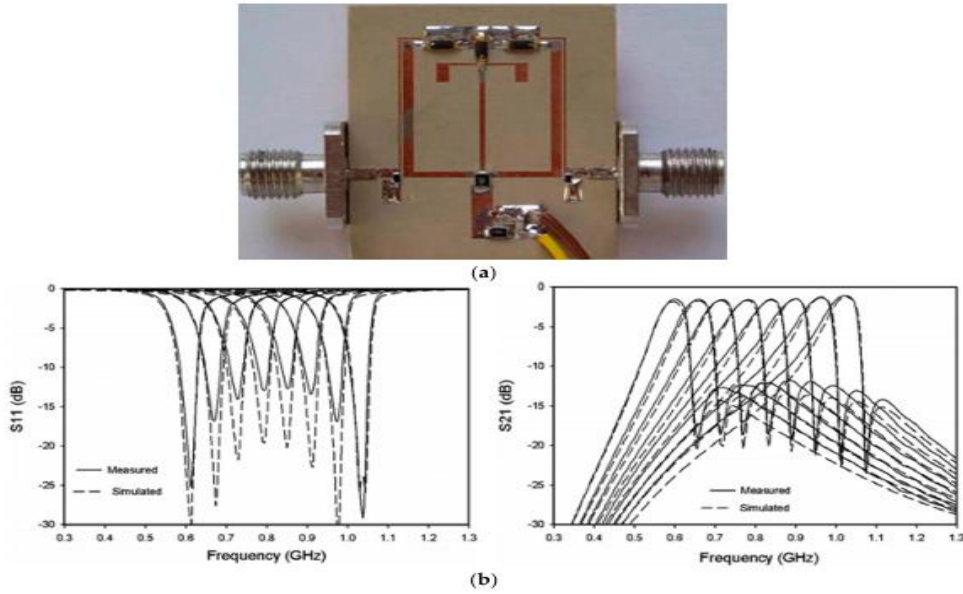


Figure II-24: Reconfigurable filter reproduced from [II.29]. 2020, IEEE: (a) prototype structure; (b) S-parameter performance.

II.6.1. Reconfigurable Microstrip Filter Concepts Using Diodes

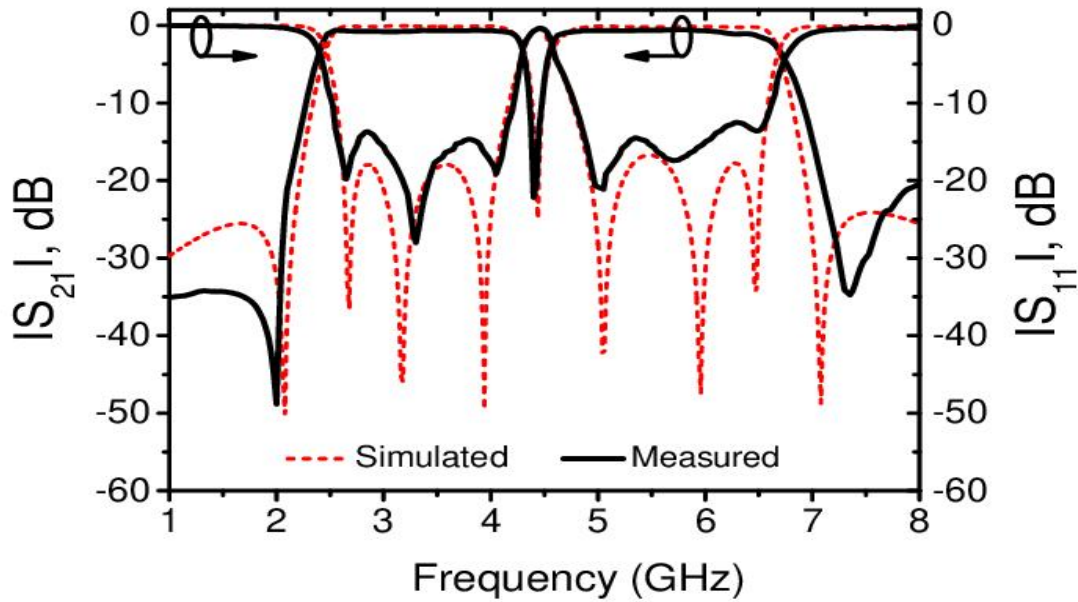
Several reconfigurable microstrip filter designs have been proposed over the years. Notable examples include:

II.6.1.1. Reconfigurable BPF Using One Varactor Diode

The practical BPF performance demonstrates good roll-off skirt on the low edge of the transmission band, with insertion loss below 2.2 dB and return loss exceeding 10 dB. A

reverse bias voltage of 2.2--22.0 V applied across the varactor diode achieves a 40% tuning rate over the 0.60--1.0 GHz range, with 91 MHz impedance bandwidth for all configurations.

Figure II-24(a) shows the design, while Figure II-25 presents (a) simulated and measured S-parameters and (b) photograph of a notch-band BPF with controllable notch bandwidth.



(a)



(b)

Figure II-25: (a) Simulated and measured S-parameters, and (b) photograph of NWB BPF with controllable notch bandwidth.

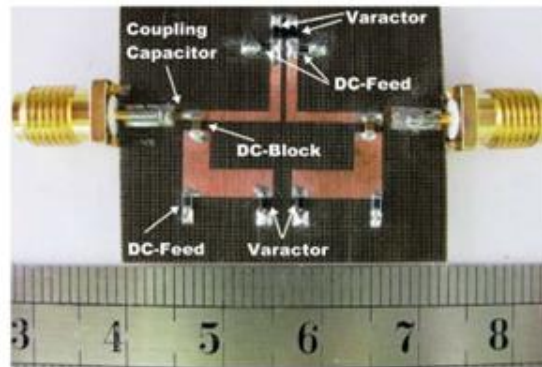
II.6.1.2. Reconfigurable BPF Using Two Varactor Diodes

In [II.54], a reconfigurable microstrip bandpass filter (BPF) is implemented using two varactor diodes to control two finite transmission zeros (TZs). The filter's center frequency and bandwidth can be adjusted over a wide tuning range of approximately 600 MHz, covering

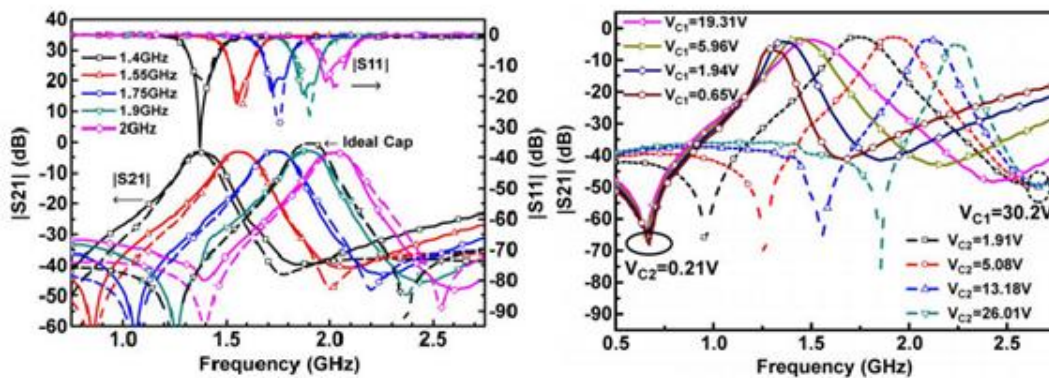
the frequency interval from 1.4 GHz to 2.0 GHz by modifying the reverse bias voltage applied to the varactors, as illustrated in Figure II-26.

Experimental results indicate that the filter exhibits an insertion loss of less than 4 dB and a return loss greater than 18 dB, with a fractional bandwidth of approximately 10%. A stopband rejection level exceeding 25 dB is achieved through the introduction of two transmission zeros. The tuning operation is realized by applying a bias voltage ranging from 0.21 V to 30.02 V across the diodes to modify the resonance characteristics.

In [II.55], a compact tunable planar BPF with nearly constant fractional bandwidth is presented. By increasing the reverse bias voltage applied to the switching elements, the filter's center frequency is tuned from 3.4 GHz to 3.8 GHz while maintaining a fractional bandwidth of approximately 11%. This design is characterized by its compact structure and simplicity, as it requires only a single varactor diode switch.



(a)



(b) **Figure II-26:** (a) Filter design. (b) $|S_{11}|$ and $|S_{21}|$ for different tuning states.

II.6.1.3. Reconfigurable Band-Stop Filters

Ebrahimi et al. [II.56] proposed a dual-mode tunable notch band-stop planar filter using two varactor diodes. The filter operates by introducing inductive and capacitive coupling elements into the input and output microstrip transmission lines. The inductive components are realized using thin inductive strip structures.

As illustrated in Figure II-27, the designed second-order filter has a compact footprint of approximately $0.13 \lambda_g \times 0.17 \lambda_g$. It provides a continuous tuning capability for the resonance frequency across the range of 0.8 GHz to 1.1 GHz, with a stopband fractional bandwidth of about 17%.

Experimental measurements show that the filter achieves a stopband return loss of approximately 0.9 dB and a passband insertion loss of approximately 0.6 dB throughout the entire tuning interval. Unlike some multilayer implementations, the inductive coupling in this design is achieved using a bottom-layer inductor beneath the patch filter. This approach eliminates the need for a more complex three-layer structure, increases flexibility in controlling coupling coefficients, and preserves a simple top-layer configuration, contributing to a more compact overall design.

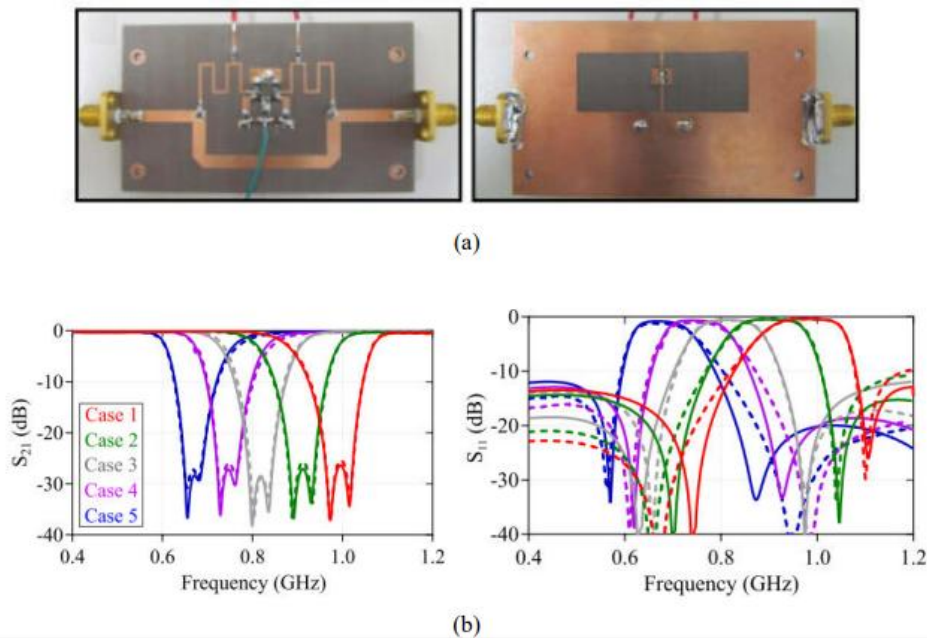


Figure II-27: (a) Proposed tunable filter upper and bottom layers. (b) S_{21} and S_{11} coefficients across the tuning range.

II.6.1.4. Compact 2-Pole BPF

Chen et al. [II.57] proposed a fully tunable two-pole planar bandpass filter characterized by compact dimensions, continuous frequency tuning, and nearly constant impedance bandwidth. The design employs two varactor diodes to control the resonance frequencies associated with the high and low resonant modes.

The filter structure consists of two resonators, each formed by a pair of transmission lines connected through a reverse-biased varactor diode. A bias voltage ranging from 0.4 V to 18 V is applied to the diodes, allowing the capacitance to vary between approximately 0.3 pF and 2.4 pF. This capacitance tuning enables resonance frequency adjustment over the range of 1.2 GHz to 1.9 GHz, with an operational impedance bandwidth of approximately 39 MHz.

The proposed filter exhibits a very compact footprint of about $0.06 \lambda_g \times 0.27 \lambda_g$, along with continuous tunability and a relatively simple configuration. These characteristics make the design suitable for modern and emerging wireless communication systems. Figure II-28 presents the measured insertion loss and return loss responses of the fabricated filter prototype.

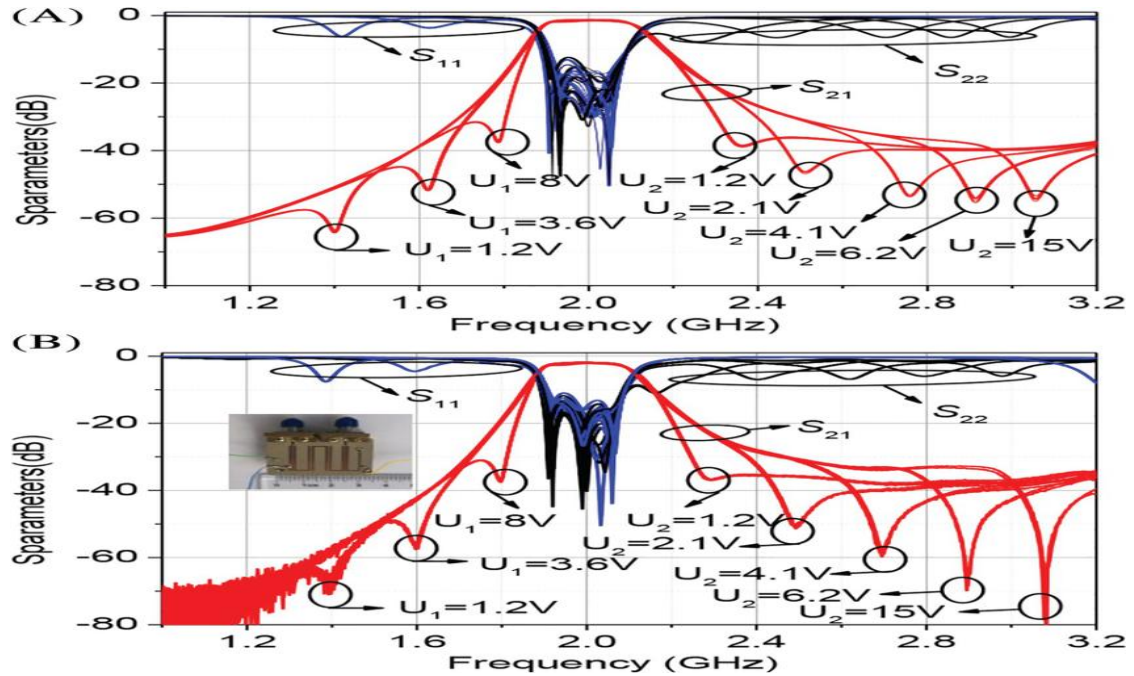


Figure II-28: S_{11} and S_{21} coefficients of the proposed reconfigurable filter, with photograph of the fabricated model.

II.6.1.5. Tunable Microstrip Filter for 4G and 5G Applications

In [II.60], a highly compact reconfigurable microstrip filter suitable for fourth-generation (4G) and sub-6 GHz fifth-generation (5G) communication systems is presented. The design is based on a hybrid co-simulation approach and employs three coupled-line resonators combined with quarter-wavelength ($\lambda/4$) open-circuit stubs.

The coupling coefficients between adjacent and non-adjacent resonators are adjusted to tune the filter's center frequency across the range of 2.5 GHz to 3.8 GHz. This enables the filter to support multiple operating bands required in modern wireless communication systems Figure II-29 shows the simulated insertion and return losses of the proposed reconfigurable filter.

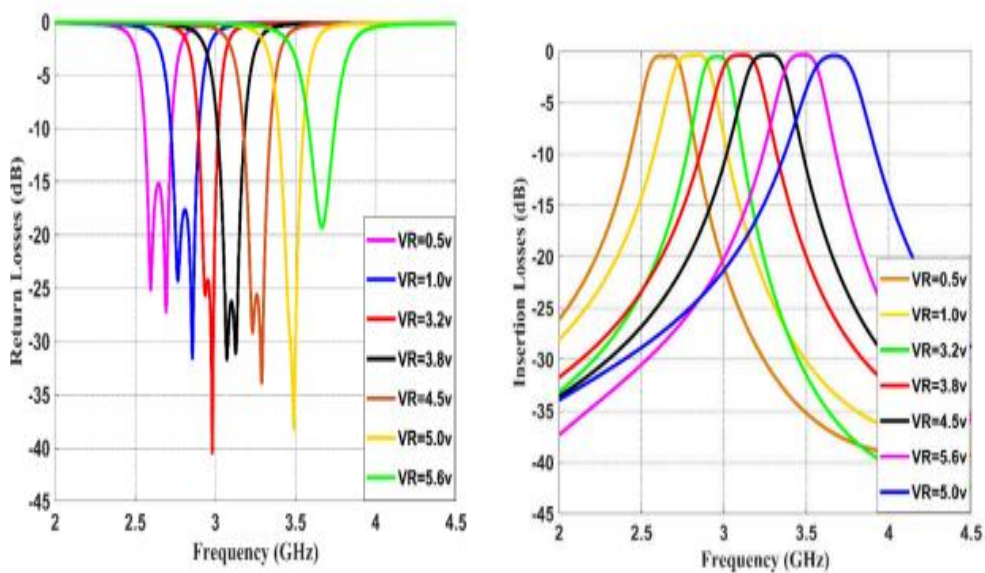


Figure II-29: Simulated S-parameters of the proposed filter.

As wireless communication technologies such as 4G and 5G continue to evolve, there is increasing demand for compact filters that provide wide frequency tuning ranges. Several tunable filter designs have been proposed to meet these requirements by improving key performance characteristics.

The comparative analysis presented in Table II-1 shows that the filter proposed in [II.60] achieves a wider tuning range, lower insertion loss, broader impedance bandwidth, and smaller physical dimensions compared to the designs reported in [II.53], [II.54], [II.56], and [II.60].

Furthermore, the tunable filters described in [II.53] and [II.57] exhibit relatively narrow impedance bandwidths of approximately 40 MHz. The designs presented in [II.59] and [II.60] utilize only two varactor diode switches combined with a simple biasing circuit to achieve frequency tuning while maintaining efficient electrical performance.

Overall, the filter proposed in [II.60] demonstrates favorable S-parameter behavior, including improved group delay and S21 phase characteristics. Its compact size, limited number of tuning elements, and simplified structure make it a promising candidate for 5G communication systems.

Table II-1: Comparison of tunable microstrip filters

Ref.	Year	Topology	Tuning Range (GHz)	BW (MHz)	Switches	IL (dB)	Size (mm ³)
[53]	2010	Dual-Mode	0.6–1.0	85–95	3	2.2	30×23×1.27
[54]	2011	Coupled lines	1.5–2.0	110	4	4	36×30×0.80
[56]	2018	Dual-Mode	0.66–0.99	108	4	0.75	72×70×1.6
[57]	2018	Ring-resonator	1.1–2.1	40	7	6	52×12×1.6
[58]	2018	Dual-Mode	1.7–2.9	40	7	4	36×35×0.8
[59]	2018	Multimode	0.76–2.0	75–150	2	1.2	100×8×0.50
[60]	2019	Coupled lines	2.5–3.8	95–115	2	0.8	13×8×0.80

II.7. Conclusion

This chapter presented an overview of transmission line theory and the electromagnetic spectrum, along with key concepts of coplanar and microstrip technologies. It also discussed filter theory, emphasizing the role of filters in communication and signal processing systems, and classified them into two families and four main categories based on their frequency response.

In addition, three common filter design approximations—Butterworth, Chebyshev, and Elliptic—were examined, highlighting their characteristics. Finally, recent advances in reconfigurable microstrip filters using diode-based tuning were reviewed, with a comparison to conventional fixed filters outlining their respective advantages and limitation.

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CHAPTER III:

Frequency-Tunable Metamaterial Stopband Filter

III.1. Introduction

Modern communication systems increasingly rely on highly integrated wireless devices, which dictates a stringent need for both the miniaturization and adaptability of radio frequency (RF) components. Within this context, microwave filters serve as fundamental building blocks, fulfilling the critical functions of signal selection and interference mitigation. Stopband filters, in particular, play an indispensable role in the targeted suppression of unwanted frequency bands.

To address the aggressive size constraints of contemporary wireless systems, metamaterial-inspired designs have emerged as a highly effective approach. Specifically, architectures utilizing Split-Ring Resonators (SRRs) facilitate a significant reduction in physical footprint by exploiting the unique sub-wavelength resonant characteristics of these structures. Concurrently, the growing demand for system versatility has elevated the importance of reconfigurable filters capable of dynamically adjusting their frequency responses. While early adaptive filters relied heavily on mechanical switching, modern iterations leverage advanced semiconductor technologies. By directly embedding active components—such as PIN or varactor diodes—into the filter circuitry, rapid and efficient electronic tuning can be achieved.

Addressing these dual imperatives of extreme compactness and dynamic functionality, this chapter presents the design, development, and analysis of a novel miniaturized stopband filter that exhibits both mode and frequency reconfigurability. The foundation of this proposed design is a newly synthesized metamaterial unit cell. Although derived from traditional SRR topologies, this optimized cell is explicitly engineered to maximize electrical length, thereby achieving superior compactness compared to conventional counterparts. This fundamental building block is subsequently expanded into a multi-resonator stopband architecture that integrates six strategically positioned PIN diodes. Functioning as high-speed electronic switches, these diodes dictate the resonant characteristics of the overall filter.

To systematically articulate this research, the remainder of the chapter is divided into two primary sections:

- **The first section** explores the design evolution of the proposed metamaterial unit cell, detailing its fundamental operating principles and comparative advantages over

existing designs. It further outlines the methodological progression of integrating these individual cells into the complete multi-resonator filter architecture.

- **The second section** is dedicated exclusively to the system's reconfigurable characteristics. It details the physical incorporation of the PIN diodes, elucidates the underlying mechanisms by which distinct biasing states enable dynamic frequency and mode switching, and concludes with a comprehensive performance analysis of the filter across its diverse operational configurations.

III.2. Resonant-type CRLH transmission lines

Split-ring resonators (SRRs) were first introduced in their circular form by Pendry [III.1], marking a key development in metamaterial research. Subsequent works have proposed a variety of alternative geometries to enhance or tailor their electromagnetic response. The original study primarily investigated the metamaterial behavior, with particular emphasis on achieving negative effective permeability.

Structurally, an SRR unit cell is composed of two concentric metallic rings, each incorporating a gap located opposite to that of the other ring. The geometrical parameters, including the gap width, ring spacing, and overall dimensions, are carefully designed, as they critically determine the resonant characteristics of the structure. Owing to the split configuration, displacement currents are induced to flow between the inner and outer rings. As a result, the SRR are effectively represented using an equivalent LC circuit model [III.2].

III.2.1. The extraction techniques for effective parameters

A key objective of the present analysis is to verify that the considered metamaterial structures exhibit a negative n in the vicinity of their resonant frequency. The effective permittivity and permeability can be extracted from the reflection and transmission coefficients using several established retrieval methods. The choice of method depends on factors such as the available S-parameters (S_{11} and S_{21}), sample length, desired material properties, computational efficiency, and accuracy requirements.

In this context, the S-parameters are obtained by illuminating a well-defined metamaterial slab with a normally incident wave. Under the assumption of well-characterized boundaries, the Fresnel equations are applied at the air–metamaterial interface, where S_{11} represents the reflection coefficient and S_{21} corresponds to the transmission coefficient.

$$S_{11} = M_{11}e^{-i\varphi_{11}} \quad (\text{III. 1})$$

$$S_{21} = M_{21}e^{-i\varphi_{21}} \quad (\text{III.2})$$

The moduli and phases of reflection and transmission are represented by M_{11} , M_{21} , φ_{11} , and φ_{21} , respectively. The Nicolson-Ross-Weir (NRW) technique calculates the permittivity and the permeability directly from S parameters [III.3,III.4]. The four coefficients (S_{11} , S_{21} , S_{12} , S_{22}) or a pair (S_{11} , S_{21}) of the coefficients of the material under test must be simulated or measured in order to extract the reflection coefficient and transmission coefficient. We will go through the NRW method's premise as well as its limitations in the following sections.

III.2.1.1. Extraction by the Nicolson-ross-weir method

Originating from the inversion of the Fresnel equations, the Nicolson-Ross-Weir (NRW) extraction approach utilizes a composite medium's wave impedance and refractive index to model its normal-incidence reflection and transmission coefficients. These derived quantities subsequently allow for the determination of the material's average permittivity and permeability. Frequently referred to as the distributed impedance method, the NRW protocol is widely implemented in both computational simulations and experimental setups to characterize natural media as well as granular, densely packed composite structures [III.3, III.4]. Its initial adaptation for analyzing metamaterials was pioneered by Smith et al. [III.5]. Nonetheless, the NRW methodology possesses distinct constraints, most notably the requirement of a single propagating mode at the operating frequency. Furthermore, a precise extraction of effective constitutive parameters can only be achieved if the incident wavelength is substantially larger than both the geometric dimensions and the spatial distribution of the individual components within the composite structure.

Fundamentally, the NRW technique relies on straightforward interference calculations to evaluate the transmission and reflection profiles of a material layer based on its thickness, effective index, and effective impedance. Smith proved that by inverting the Fresnel equations we can obtain the values of n_{eff} and z_{eff} as a function of the simulated layer thickness d at the coefficients of transmission $t' = s_{21}$ and reflection $r = s_{11}$ [III.5]:

$$Re(n_{eff}) = \pm Re\left(\frac{\arccos\left(\frac{1}{2t'}[1 - (r^2 - t'^2)]\right)}{kd}\right) + \frac{2\pi m}{kd} \quad (\text{III.3})$$

$$Im(n_{eff}) = \pm Im\left(\frac{\arccos\left(\frac{1}{2t'}[1 - (r^2 - t'^2)]\right)}{kd}\right) \quad (\text{III.4})$$

$$z_{eff} = \pm \sqrt{\frac{(1+r)^2 - t'^2}{(1-r)^2 - t'^2}} \quad (III.5)$$

Where m is an integer, the thickness of the material is represented by d , and the vacuum wave vector of the incident plane wave is represented by $k = \omega/c$.

The effective parameters ϵ_{eff} and μ_{eff} can be calculated from the effective index and impedance using the following equation:

$$\epsilon_{eff} = \frac{n_{eff}}{z_{eff}} \quad (III.6)$$

$$\mu_{eff} = n_{eff} \cdot z_{eff} \quad (III.7)$$

III.2.2. Magnetically active resonator

III.2.2.1. Circular SRR

Initiated by Pendry, the circular split ring resonator (SRR) was originally developed for operation within the X-band frequency range [8.2 GHz; 12.4 GHz] [III.1]. As shown in Figure III-1, the structure is defined by an outer ring radius of $r = 3$ mm, a ring width of $c = 0.33$ mm, an opening gap of $g = 0.33$ mm, and an inter-ring spacing of $d = 0.33$ mm. The design was simulated on a ROGERS RO4003C substrate characterized by a relative permittivity of 3.38, a loss tangent of approximately 0.0027, and a substrate thickness of $h = 0.81$ mm.

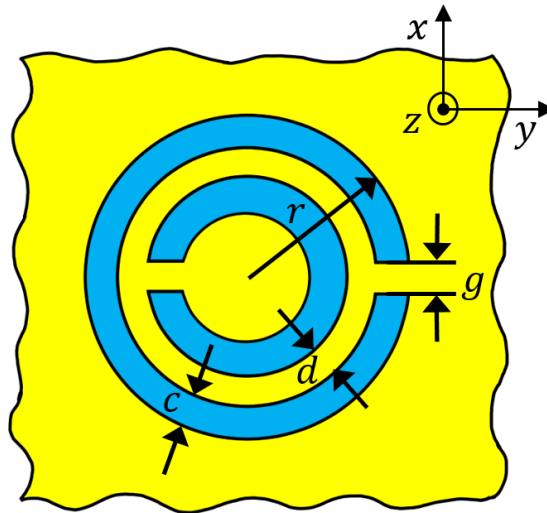


Figure III-1: A circular split-ring resonator [III.5].

The inductance L and capacitance C can be used to describe a subwavelength Split-Ring resonator (Figure III-2) [III.6].

The resonant frequency is:

$$f_r = \frac{1}{2\pi\sqrt{LC}} \quad (\text{III. 8})$$

$$L = \mu_0 r_m \left(\ln \left(\frac{8 r_m}{h + c} \right) - 0.5 \right) \quad (\text{III. 9})$$

Where μ_0 denotes vacuum permeability, and r_m the ring's mean radius.

$$r_m = r + \frac{w}{2} \quad (\text{III. 10})$$

Where r is the inner radius of the ring.

The surface capacity (C_{surf}) and the gap capacity (C_{gap}) are the two capacities used in this study. Where the gap capacitance is the result of the charges in the gap, which can be used as a parallel plate capacitance. Meanwhile, the surface capacity is provided by loads on the surface of the ring [III.6].

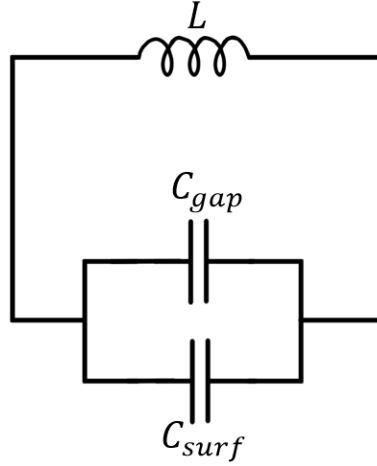


Figure III-2: The circular SRR equivalent circuit.

$$C_{\text{gap}} = \epsilon_0 \left[\frac{wh}{g} + \frac{2\pi h}{\ln \left(\frac{2.4h}{w} \right)} \right] \quad (\text{III. 11})$$

$$C_{\text{surf}} = \frac{2\epsilon_0 h}{\pi} \ln \left(\frac{4R}{g} \right) \quad (\text{III. 12})$$

Assuming that the gap and surface capacitances are parallel, the total capacitance is:

$$C = C_{gap} + C_{surf} \quad (\text{III.13})$$

Carried out simulations aim at extracting the S-parameters of a network of SRR structures using the HFSS electromagnetic solver. In particular, the coefficients S_{11} and S_{21} are employed for determining the effective permeability μ_{eff} . Prior to running the simulation, appropriate boundary conditions must be defined at the extremities of the structure. In this context, electric walls are imposed along the y-axis, while magnetic walls are assigned along the x-axis, with wave propagation oriented along the z-axis. To guarantee proper magnetic coupling, the magnetic field is aligned parallel to the axis of the rings, ensuring magnetic excitation of the SRR; consequently, periodic boundary conditions are applied to the unit cell under study. Furthermore, the Rogers 4003C dielectric substrate is widely used in the fabrication of multilayer electronic boards and is commonly associated with double-sided PCB applications. In this work, however, only a single-layer configuration is considered to model electromagnetic wave propagation through the structure.

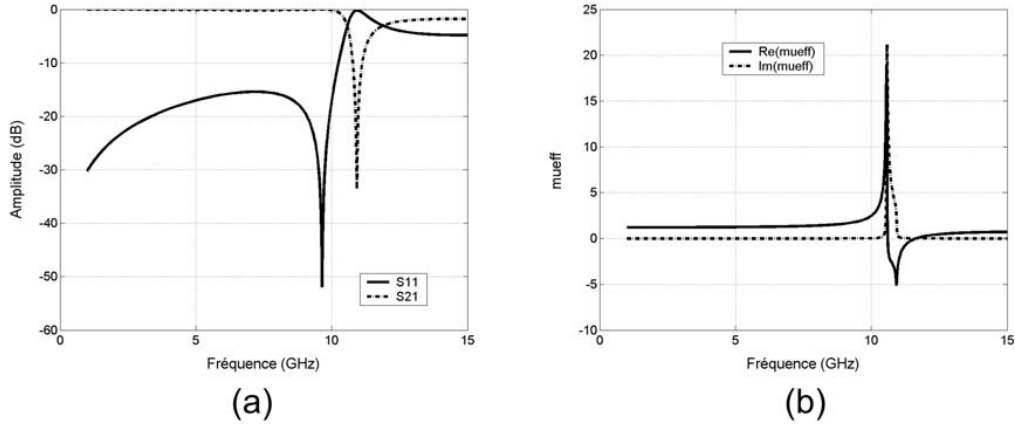


Figure III-3: The E-field along the z-axis: (a) the parameters S of the circular SRR structure and (b) Real and imaginary parts of the effective permeability.

In the configuration where the electric field is oriented along the z-axis and the propagation direction is along the y-axis, the simulated S-parameters are presented in Figure III-3 (a). The resonance occurs at 10.925 GHz, reaching a minimum of approximately -34 dB. The real part of the effective permeability exhibits negative values over a narrow frequency band, ranging from 0 down to -5, whereas outside this region $\text{Re}(\mu_{eff})$ remains positive, as illustrated in Figure III-3 (b).

When the electric field is polarized along the z-axis, the resulting E-field distribution preserves the symmetry of the structure and does not effectively excite the gaps of the SRR rings. As a consequence, only magnetic resonance is observed. It is further noted that in cases

where both electric and magnetic resonances are simultaneously excited, the resonance frequency tends to shift slightly higher compared to the scenario where only magnetic resonance is present.

III.2.3. Electrically active resonator (metal line)

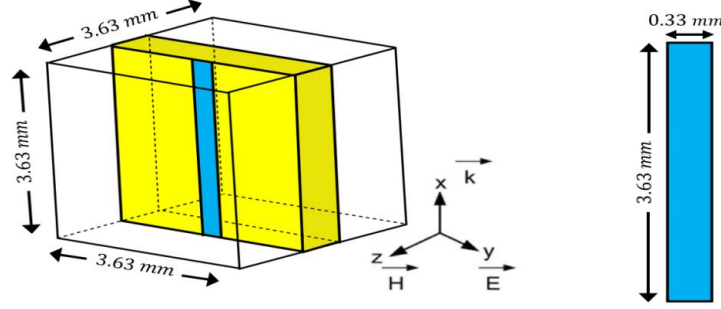


Figure III-4: The structure of metal line unit cell.

The electromagnetic behavior of a metallic line unit cell can be modeled in a manner similar to a low-density plasma, which leads to an effective negative permittivity [III.7]. In this configuration, boundary conditions are defined such that electric walls are imposed on the two faces perpendicular to the z-axis, while magnetic walls are assigned to the faces perpendicular to the x-axis on both sides, as illustrated in Figure III-4.

The metallic strip has a length of 3.63 mm, corresponding both to the height of the radiation box and to the periodicity of the structure, while its width is 0.33 mm, matching the gap or opening associated with the magnetically active resonator rings.

The transmission and reflection characteristics of the metallic line are presented in Figure III-5. The investigated structure exhibits a plasma frequency at 20.9 GHz, corresponding to a -3 dB point and a high-pass transmission response. It is further observed that the structure provides full rejection of electromagnetic waves below the plasma frequency.

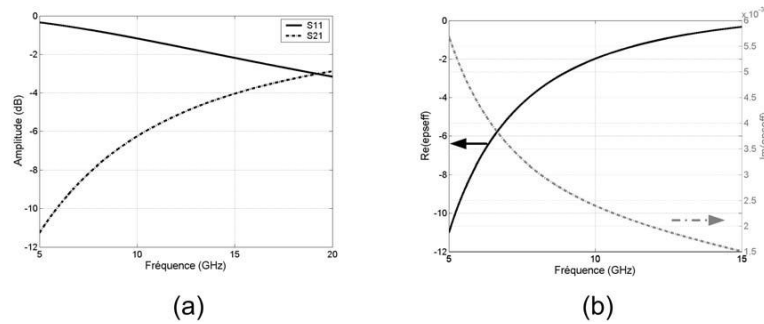


Figure III-5: The simulated results of the metal line: (a) Reflection and transmission in dB. (b) Real and imaginary parts of the effective permittivity [III.7].

III.3. Frequency and mode reconfigurable microstrip stopband filter using PIN diodes and new metamaterial cells

In this section, a new SRR design is proposed. This SRR is used to structure a reconfigurable stopband filter. This SBF is designed and optimized to resonant with three important wireless communication bands. Furthermore, the reconfigurability is achieved by loading six PIN diodes onto the filter structure. This method allowed the achievement of SBF with seven different cases, in terms of the resonant frequency and the number of rejected bands (triple, dual, or single-mode stopband filter).

Besides the above, the presented SBF model has a small size compared to similar papers that have been reported, and this is thanks to the new SRR. Moreover, the PIN diodes are loaded in such a way that the design is simple. The measured results showed minimum losses and good electromagnetic properties.

III.3.1. Proposed Split-Ring Resonator Design

The proposed planar metamaterial unit cell, shown in Figure III-6 (a), is implemented on a dielectric substrate (shown in green) and incorporates a conductive pattern (depicted in copper orange). The design features two primary concentric resonating elements that exhibit vertical symmetry. The outer element is a large, nearly circular ring serving as the primary split-ring resonator (SRR), characterized by a narrow vertical split positioned at the apex (12 o'clock).

Nested within this ring is a smaller, elliptic inner structure that also includes a vertical split at its apex. This inner gap is precisely aligned with that of the outer ring and is electrically connected via two conductive rectangular stubs—one on each side of the splits—thereby enabling strong capacitive coupling between the two resonators. Notably, the lower segment of the inner structure (near the 6 o'clock position) deviates significantly from a continuous arc. Instead, it comprises two parallel vertical rectangular stubs extending upward from the ends of the semi-elliptic inner segments. These stubs are separated by a narrow, vertically oriented gap centered along the vertical axis of symmetry, introducing an additional capacitive gap within the inner resonator.

This innovative unit cell design offers several advantages:

- ✓ **Enhanced miniaturization:** The modification of the inner resonator increases the effective electrical path within a compact footprint. By directing current along the semi-elliptic contours and the vertical stubs, the structure achieves a lower fundamental

resonant frequency compared to conventional double SRRs with comparable overall dimensions (Figure III-6 (b)). This results in effective size reduction, supporting the development of miniaturized metamaterial-based devices.

- ✓ **Improved tuning flexibility:** The resonant frequency and electromagnetic response of the unit cell are highly sensitive to the geometry of the modified inner structure. Adjustments to the length or width of the vertical stubs, or to the elliptic curvature of the inner ring, provide independent tuning parameters. This flexibility facilitates the design of devices tailored to specific operational frequencies or bandwidths.
- ✓ **Simplified fabrication:** Replacing the traditionally curved lower section of the inner ring with straight vertical stubs can simplify certain fabrication steps. This design may reduce manufacturing complexity and enhance tolerance or yield, especially when compared to structures involving intricate curved features.
- ✓ **Enhanced coupling control:** The central gap between the vertical stubs introduces an additional capacitive element that impacts the intra-resonator coupling within the inner structure. This alters the overall coupling dynamics and electromagnetic field distribution relative to standard double SRRs. The gap also becomes a focal point for strong electric field concentrations at resonance, which can be exploited for engineering specific filter responses.

In this study, three target frequencies—3.6 GHz, 5.2 GHz, and 5.8 GHz—are investigated. The corresponding unit cell dimensions required to achieve resonance at each of these frequencies are detailed in Table III-1.

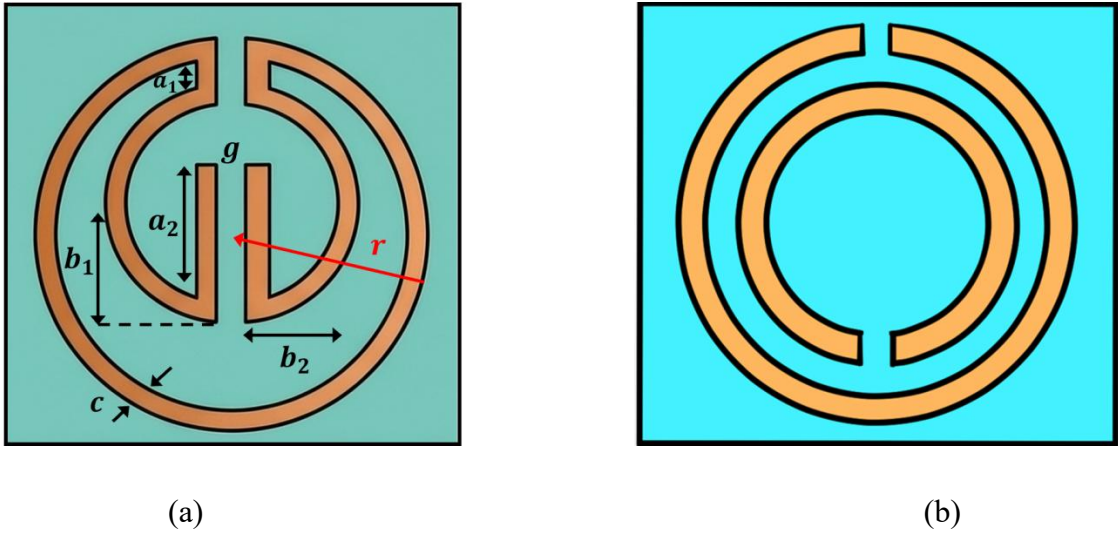


Figure III-6: (a) The layout of the proposed split-ring resonator. (b) The layout of a traditional circular split-ring resonator.

Table III-1 The different dimensions of the SRR for each case according to the target frequencies.

Variables	Dimensions (mm)		
	3.6 GHz SRR	5.2 GHz SRR	3.6 GHz SRR
a_1	1	1	1
a_2	2	1.4	0.6
b_1	2	1.4	1.2
b_2	1.5	1.2	1
c	0.3	0.3	0.3
g	0.5	0.5	0.5
r	5.5	4.5	4.2

III.3.2. The effective refractive index

In order to examine the metamaterial behavior of the proposed filter, the real component of the effective refractive index n is retrieved from the simulated S-parameters obtained via CST Microwave Studio. The study is conducted at three designated resonant frequencies: 3.6 GHz, 5.2 GHz, and 5.8 GHz. At each of these frequencies, the SRR pair constituting the single-mode stopband filter (as depicted in Figure III-7) is accordingly adjusted, following the geometrical parameters listed in Table III-1.

The methodology, briefly outlined below, was previously detailed in Chapter Two of this dissertation:

- The S11 and S21 simulation results are exported from CST.
- The exported data are imported into MATLAB as column vectors.
- The Nicolson-Ross-Weir (NRW) retrieval technique is implemented in MATLAB using the imported S-parameters.
- The real part of the extracted effective refractive index is then evaluated.

The obtained results, illustrated in Figure III 7, indicate that the real part of the refractive index assumes negative values at all the targeted resonant frequencies. This negative index behavior is characteristic of metamaterial media and confirms the metamaterial response of the proposed SRR-based filter at the specified resonance points.

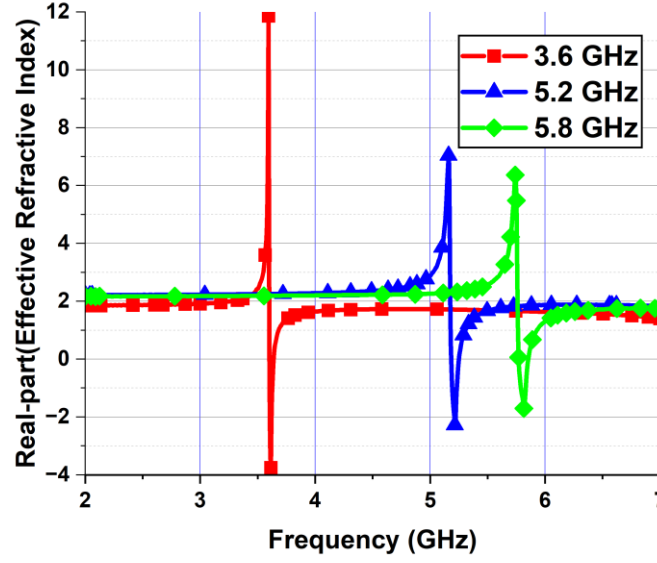


Figure III-7: Real-part of the effective refractive index n_{eff} of the different SRRs.

III.3.3. The triple-mode metamaterial-based stopband filter

III.3.3.1. Step one: stopband filter loaded with one SRR pair

The proposed unit cell design is employed to construct a stopband filter configuration. The overall design procedure is carried out through three main stages to obtain the final filter topology. The first stage consists of implementing a preliminary stopband filter, as shown in Figure III-8. This structure is composed of a $50\ \Omega$ microstrip transmission line with a length of $L = 11.50$ mm and a width of $S = 1.75$ mm. On both sides of the transmission line, two identical split-ring resonators (SRRs) derived from the newly introduced design are placed symmetrically. These resonators are tuned to operate at 3.6 GHz and are referred to as SRR1 in subsequent discussions.

The separation distance between each SRR and the microstrip line is a key parameter influencing the overall filter response. After multiple simulation iterations and fine-tuning, an optimal spacing is determined to achieve strong electromagnetic coupling between the transmission line and the resonators. The entire filter structure is realized on a Rogers 6002 substrate, which exhibits a relative permittivity of $\epsilon_r = 2.94$, a loss tangent of $\tan \delta = 0.0012$, and a thickness of $h = 0.76$ mm. This substrate is maintained as the common material platform for all subsequent designs presented in this chapter.

The S-parameter response of the proposed stopband filter is illustrated in Figure III-9, where the transmission coefficient S_{21} (red curve) and the reflection coefficient S_{11} (blue curve) are

plotted over the frequency range from 2 to 7 GHz. A pronounced dip in S_{21} is observed at the resonance frequency of 3.6 GHz, reaching an attenuation level of approximately -18 dB, which indicates strong signal suppression within the stopband region. Outside this band, S_{21} remains nearly constant, demonstrating stable transmission characteristics in the passband.

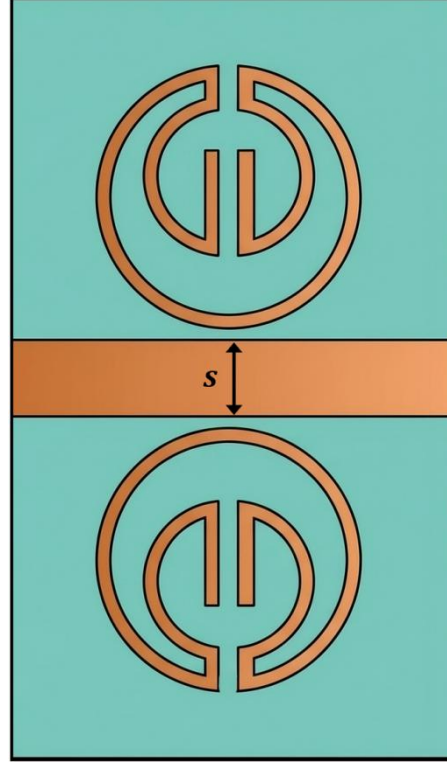


Figure III-8: Top view of the SBF loaded with one SRR pair (step one).

In parallel, S_{11} exhibits a clear peak at 3.6 GHz, coinciding with the minimum of S_{21} , which confirms that most of the incident energy is reflected at the resonant frequency—an expected behavior for an efficiently designed single-mode stopband filter. This response validates the effectiveness of the structure in selectively rejecting the targeted frequency while maintaining transmission elsewhere, in agreement with the intended design objectives for frequency-selective applications.

Additionally, given the compact geometry of the filter and the relatively low operating frequency, the obtained results highlight the superiority of the newly introduced SRR configuration compared to conventional double-ring SRR unit cells, which generally require larger physical dimensions to achieve similar filtering performance.

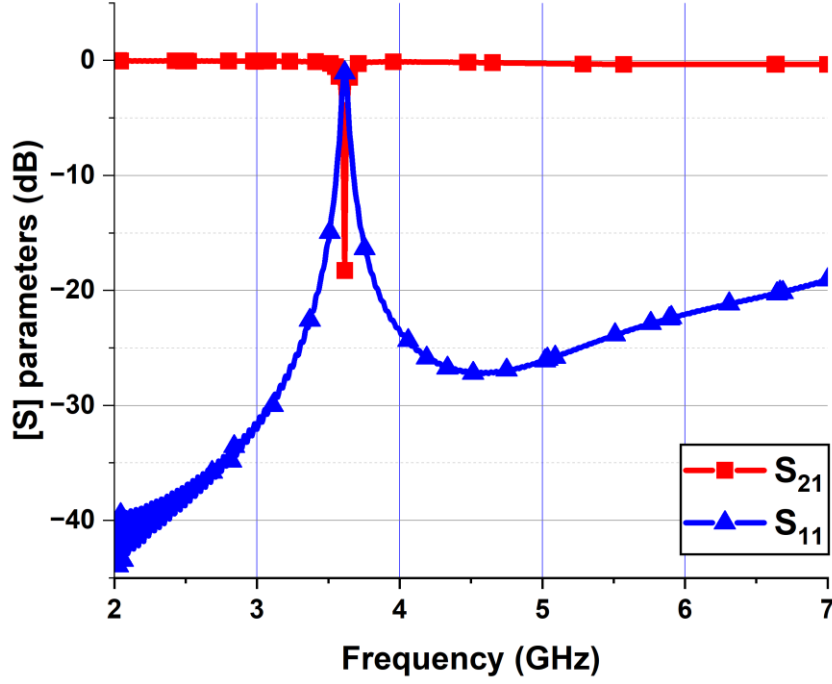


Figure III-9: Simulation results of the transmission and reflection coefficients of step one (one SRR pair).

III.3.3.2. Step two: stopband filter loaded with two SRR pairs

The top layer corresponding to the second stage of the design process is shown in Figure III-10. In this configuration, the microstrip line, substrate, and ground plane are all extended to a total length of $L = 17.7$ mm in order to integrate an additional pair of split-ring resonators (SRRs). This added pair, referred to as SRR2, is engineered using unit cell dimensions specifically tuned to achieve resonance at 5.2 GHz.

The distance separating the SRR1 pair (operating at 3.6 GHz) from the newly introduced SRR2 pair represents a key design consideration. It is precisely optimized to ensure that electromagnetic coupling between the two resonator groups remains minimal or negligible. Such isolation is necessary to preserve independent resonant behavior for each SRR set, ensuring that each target frequency is governed exclusively by its corresponding resonators. Concurrently, the design objective of compactness is preserved by carefully balancing this isolation with spatial efficiency, thereby limiting the overall footprint of the structure.

The S-parameter response of the second-stage structure, presented in Figure III-11, demonstrates a clear evolution from a single-mode to a dual-mode stopband filter. The transmission coefficient S_{21} reveals two well-defined attenuation minima, corresponding to effective rejection bands centered at approximately 3.6 GHz and 5.2 GHz. At 3.6 GHz, the rejection level reaches about -18.64 dB, while at 5.2 GHz a deeper attenuation of nearly -22 dB is achieved. These outcomes confirm the successful realization of dual-band stopband functionality.

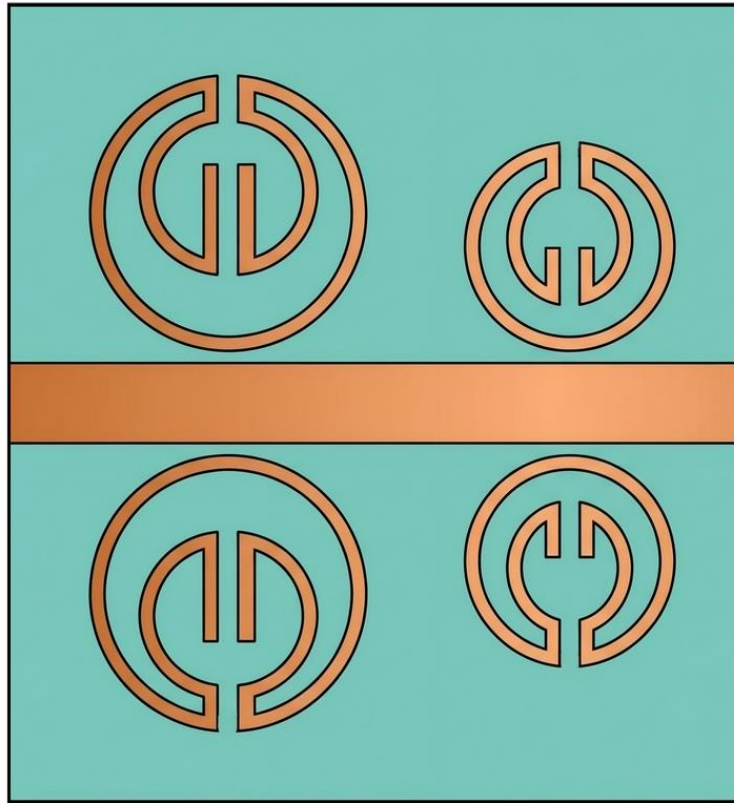


Figure III-10: Top view of the SBF loaded with two SRR pairs (step two).

Consistent with the previous configuration, pronounced stopband behavior is maintained, highlighting the capability of the multi-resonator architecture to provide selective frequency suppression at multiple bands while preserving a compact overall structure.

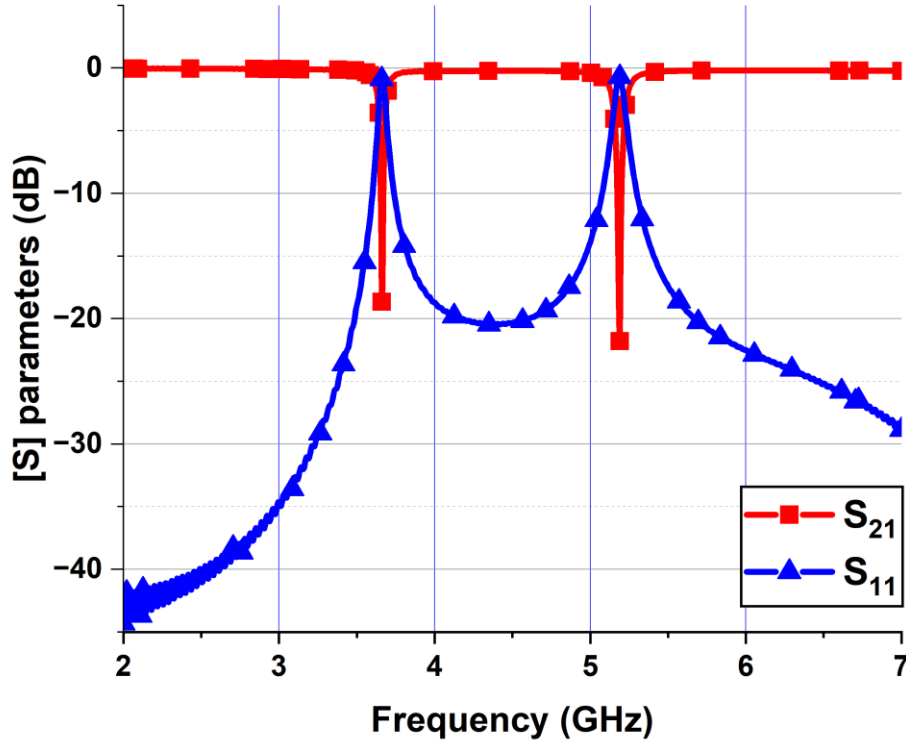


Figure III-11: Simulation results of the transmission and reflection coefficients of step two (two SRR pair).

III.3.3.3. Step three: stopband filter loaded with three SRR pairs

For the final stage of the design, a third pair of split-ring resonators (SRRs), referred to as SRR3, is incorporated into the overall structure. This additional pair is engineered with geometrical parameters adjusted to resonate at 5.8 GHz, following the same methodological approach adopted in the previous stages (Figure III-12). It should be emphasized that, with each successive design iteration, the earlier SRR pairs are subjected to slight dimensional refinements to preserve their target resonant frequencies. This adjustment is required due to the progressive extension of the microstrip line, substrate, and ground plane, where the increased transmission line length can affect the resonance conditions.

The complete filter structure has final dimensions of 25.62 mm × 18 mm × 0.76 mm. The full three-stage configuration is optimized and simulated using CST Microwave Studio to guarantee accurate electromagnetic performance.

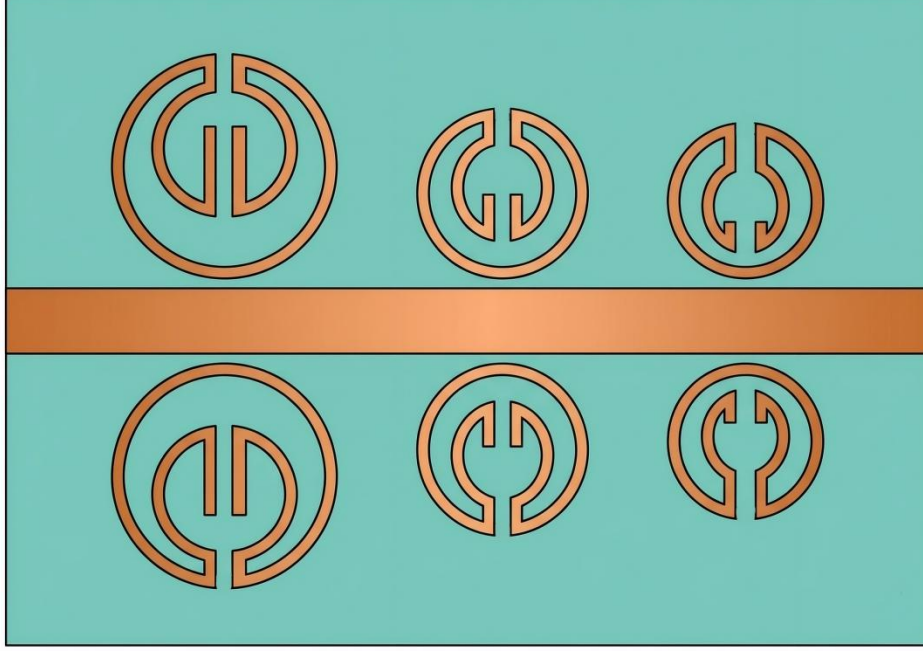


Figure III-12: Top view of the SBF loaded with three SRR pairs (step three).

The corresponding S-parameter response of the final design is shown in Figure III-13. As anticipated, the inclusion of SRR3 transforms the system into a triple-mode stopband filter. The transmission coefficient S_{21} exhibits three distinct attenuation notches located at 3.6 GHz, 5.2 GHz, and 5.8 GHz, with rejection levels of approximately -23 dB, -27 dB, and -26 dB, respectively. These results confirm the capability of the proposed structure to suppress multiple frequency bands effectively while maintaining a relatively stable response outside the stopbands.

In parallel, the reflection coefficient S_{11} presents pronounced peaks at the same resonant frequencies where S_{21} reaches its minima, indicating strong reflection of incident energy, which is consistent with the expected behavior of an efficient stopband filter. This triple-band response significantly improves the selectivity of the filter, enabling sharp rejection across multiple frequencies, which is highly advantageous for applications requiring simultaneous multi-band interference suppression.

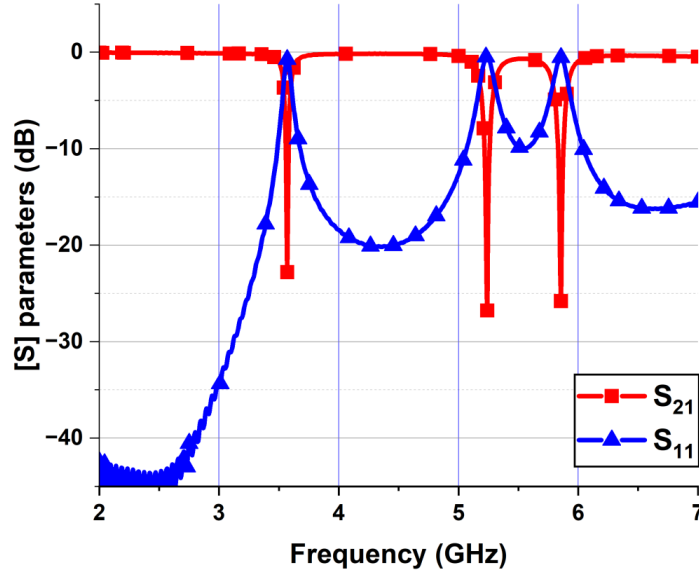
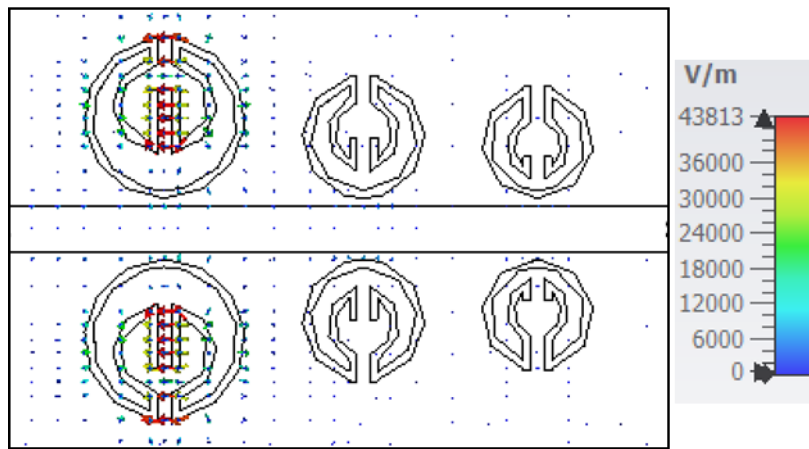


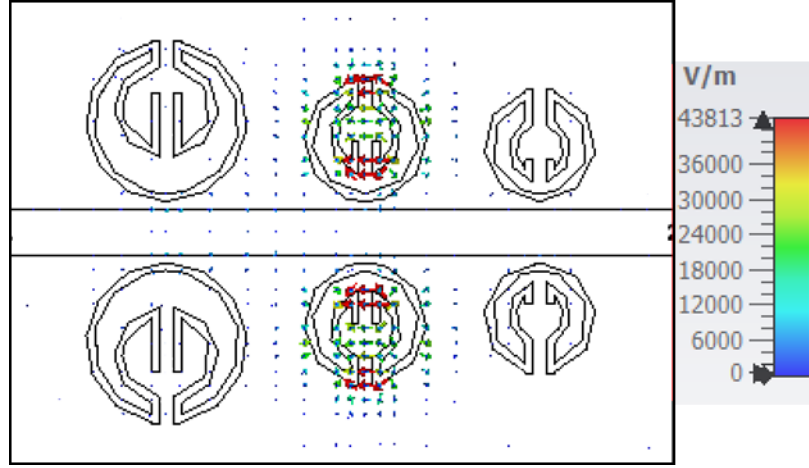
Figure III-13: Simulation results of the transmission and reflection coefficients of step three (three SRR pair).

III.3.3.3.1. The E-field distribution

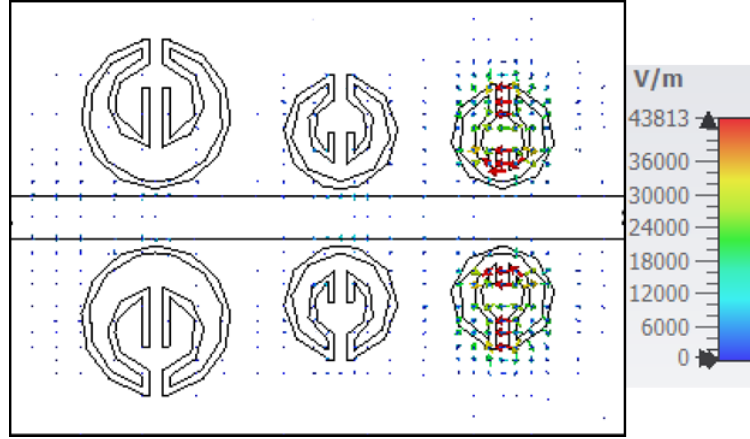
Figure III-14 presents the simulated electric field (E-field) distribution obtained using CST, showing the spatial confinement and concentration of the field around specific SRR pairs depending on the corresponding resonant frequency. This behavior confirms that coupling effects between the different resonator pairs have been effectively suppressed. Each SRR pair operates independently, with negligible impact on the rejection bands associated with the other pairs.



(a)



(b)



(c)

Figure III-14: The E-field distribution of the triple-mode stopband filter at: (a) 3.6 GHz; (b) 5.2 GHz; (c) 5.8 GHz.

This decoupled response is achieved through the careful optimization of the spacing between the three SRR pairs, which ensures sufficient electromagnetic isolation. Such a characteristic is of particular importance for the subsequent stage of the study, namely reconfigurability, where independent operation of each SRR pair at its respective frequency is required without interference or unwanted interactions from neighboring resonators. This independence establishes a robust basis for the development of flexible and frequency-selective filtering structures.

III.4. Reconfigurable Metamaterial Stopband Filter Using PIN Diodes

This section introduces the design of an electronically reconfigurable stopband filter (SBF). The reconfigurability is realized by integrating PIN diodes into the triple-mode stopband filter described previously. As shown in Figure III-15, one PIN diode is inserted across the split of the outer ring of each SRR, allowing control over the filter's characteristics. This configuration enables dynamic tuning of both the resonant frequency and the operational mode (single-mode, dual-mode, or triple-mode), depending on the state of the diodes.

PIN diodes function as voltage-controlled switches in microwave and radiofrequency systems. Their behavior is dictated by the applied bias voltage [III.8]:

- In the forward-biased (ON) state, the diode exhibits low impedance, effectively shorting the SRR gap and suppressing resonance.
- In the reverse-biased (OFF) state, the diode presents high impedance and low capacitance, allowing the SRR to behave as a resonant structure.

To achieve full reconfigurability, six BAR50-02 PIN diodes from Infineon are used—one for each SRR. The diodes are placed at the outer ring splits, a location proven optimal for switching and for ease of physical placement. Due to the metamaterial nature of the SRRs, when exposed to an external magnetic field, each SRR induces a circulating current, resulting in resonance—but only when the diode is OFF. When the diode is ON, the SRR ceases to resonate, thereby removing its influence on the stopband response. Importantly, each SRR pair (SRR1, SRR2, SRR3) must be switched simultaneously for consistent filter behavior.

This configuration yields seven distinct reconfigurable cases, summarized in Table III-2. The PIN diodes controlling the SRR1 pair are labeled P1, while those on SRR2 and SRR3 are labeled P2 and P3, respectively. These labels reflect the fact that the diodes in each pair switch together.

Table III-2 Possible different cases of the SBF based on ON/OFF PIN diodes combinations.

Cases	P1	P2	P3
011	OFF	ON	ON
101	ON	OFF	ON
110	ON	ON	OFF
001	OFF	OFF	ON
010	OFF	ON	OFF
100	ON	OFF	OFF
000	OFF	OFF	OFF

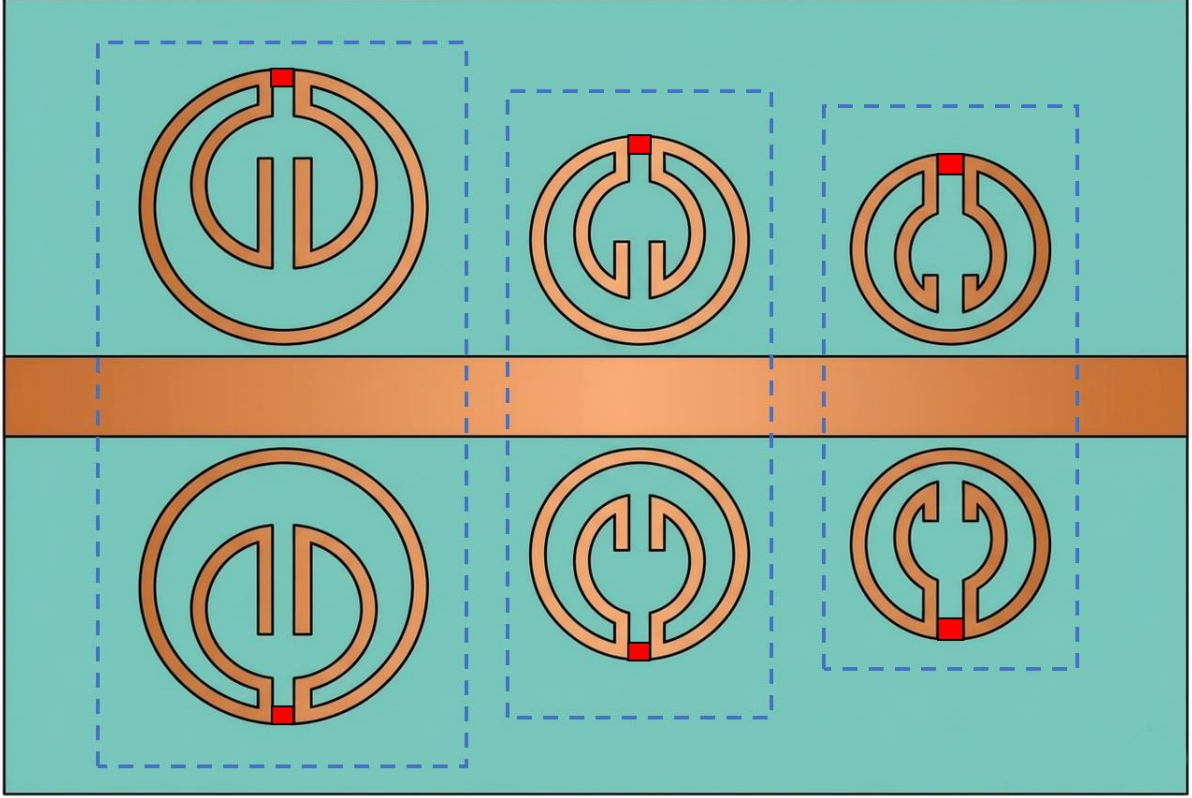


Figure III-15: Top view of the proposed reconfigurable SBF structure loaded with SRR pairs and PIN diodes.

Each SRR pair responsible for a specific stopband can be modeled as an LC resonant circuit. Due to the negligible coupling between different SRR pairs, the equivalent circuit of the proposed reconfigurable triple-mode stopband filter is presented in Figure III.16.

$$C = \frac{1}{2\pi Z(f_{U-10} - f_{L-10})} \quad (\text{III-14})$$

$$L = \frac{1}{4\pi^2 f_r^2 C} \quad (\text{III-15})$$

Where f_{U-10} and f_{L-10} represent the upper and lower frequencies of the -10 dB stopband bandwidth, respectively. The input impedance of the filter is denoted by Z .

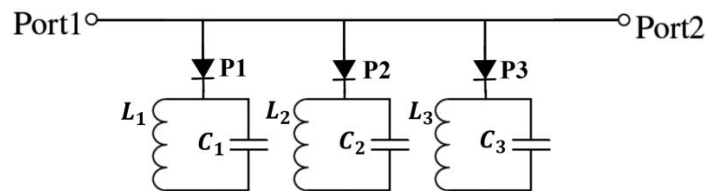


Figure III-16: The equivalent circuit of the proposed tunable SBF.

The PIN diodes are represented using lumped-element models. In the ON state, a diode is modeled by a series resistance and inductance, corresponding to its low-impedance behavior. In the OFF state, it is represented by an inductance in series with a parallel RC network to account for its high impedance and parasitic effects. These models provide an accurate representation of the filter's reconfigurable operation.

III.4.1. The outcome of the different SBF Configurations.

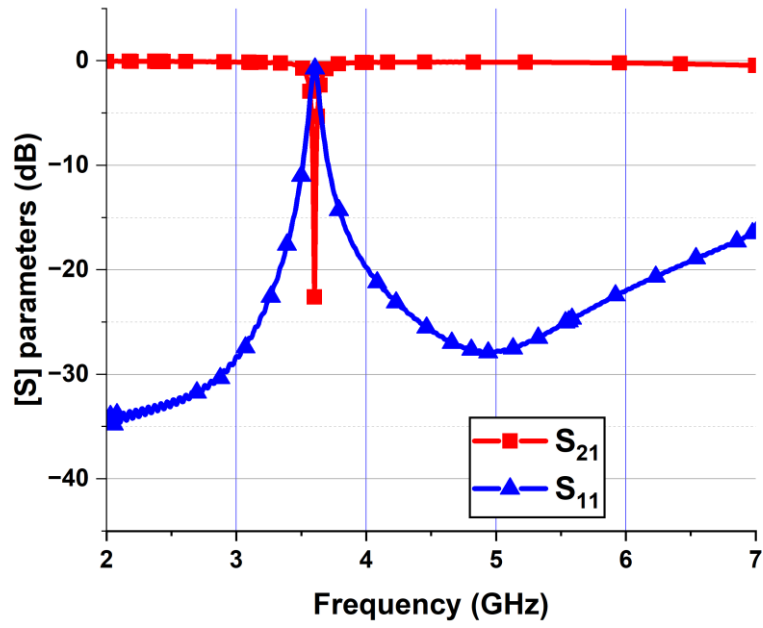
The simulation results depicted in Figure III.17, Figure III.18, and Figure III.19 demonstrate the performance of the proposed reconfigurable stopband filter under the seven switching configurations summarized in Table III.3. The obtained responses validate the intended stopband characteristics for each operating mode.

A notable observation is that the resonant frequencies remain unchanged at 3.6 GHz, 5.2 GHz, and 5.8 GHz throughout all switching states. This behavior indicates that the reconfiguration process does not introduce any frequency deviation, thereby ensuring excellent frequency stability during transitions between operating modes.

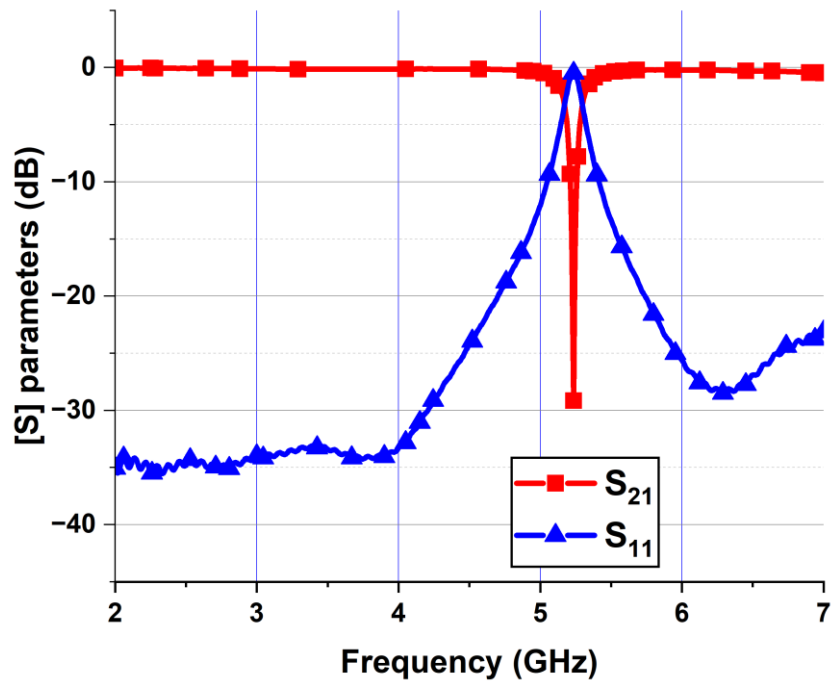
Based on the activation states of the SRR resonators, the seven operating configurations can be classified into three categories:

1. **Single-mode stopband operation:** This mode is achieved when only one SRR pair is active, corresponding to the condition where one pair of PIN diodes is in the OFF state while the remaining two pairs are ON. The cases are 011, 101, and 110.
2. **Dual-mode stopband operation:** This configuration occurs when two SRR pairs are activated, which corresponds to two PIN diode pairs being OFF and one pair being ON. The relevant switching cases are 001, 010, and 100.
3. **Triple-mode stopband operation:** This mode is obtained when all three SRR pairs are active, corresponding to all PIN diodes being in the OFF state (case 000). This configuration represents the fundamental operating state of the filter and is equivalent to the original structure without the incorporation of PIN diodes.

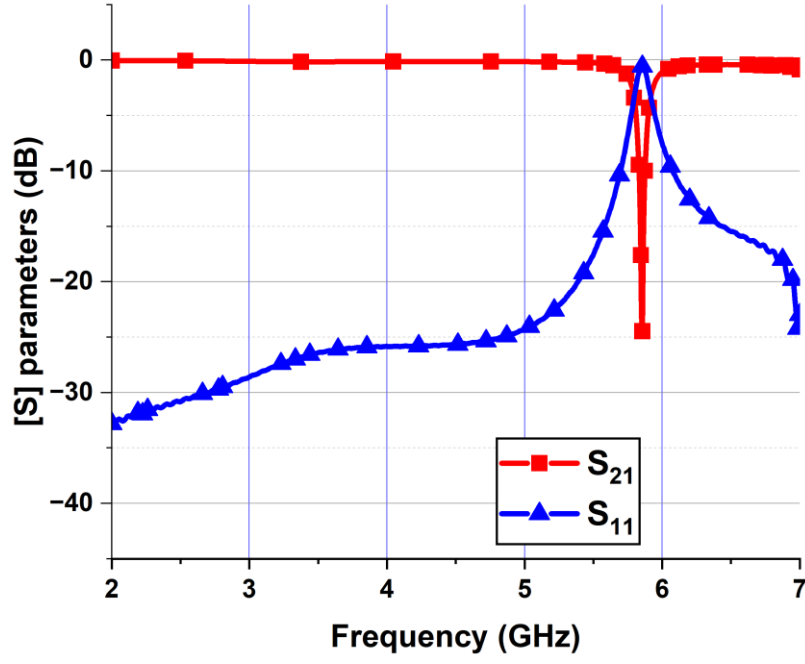
Furthermore, a particular configuration, designated as case 111, is observed when all PIN diodes are switched ON, thereby deactivating all SRR resonators. Under this condition, the stopband response disappears completely, and the structure exhibits no rejection band. This result further verifies the successful implementation of the reconfiguration mechanism and demonstrates the ability of the proposed design to fully control the filter response through PIN-diode switching.



(a)

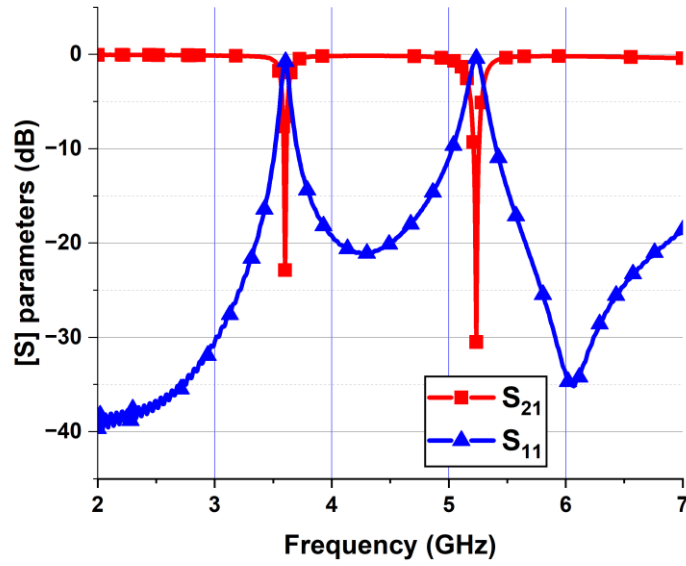


(b)

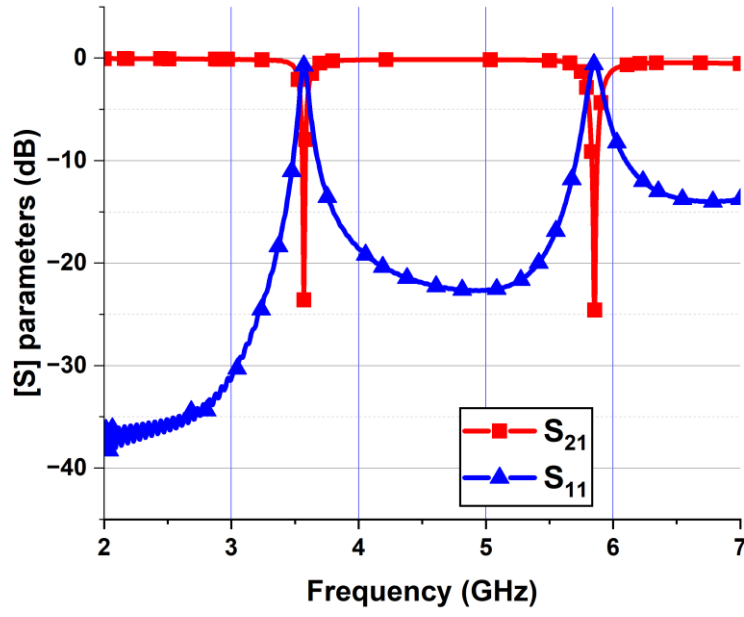


(c)

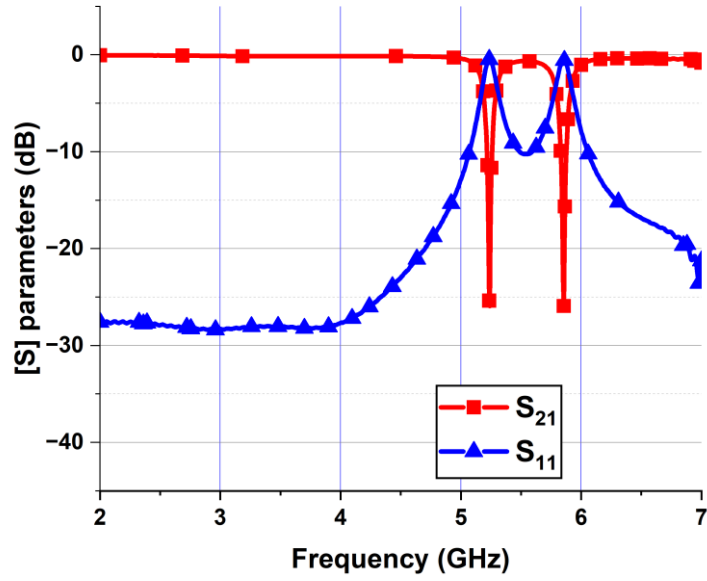
Figure III-17: Simulated S-parameters of the reconfigurable filter in its single-modes: (a) case 011; (b) case 101; (c) case 110.



(a)



(b)



(c)

Figure III-18: Simulated S-parameters of the reconfigurable filter in its dual-modes: (a) case 001; (b) case 010; (c) case 100.

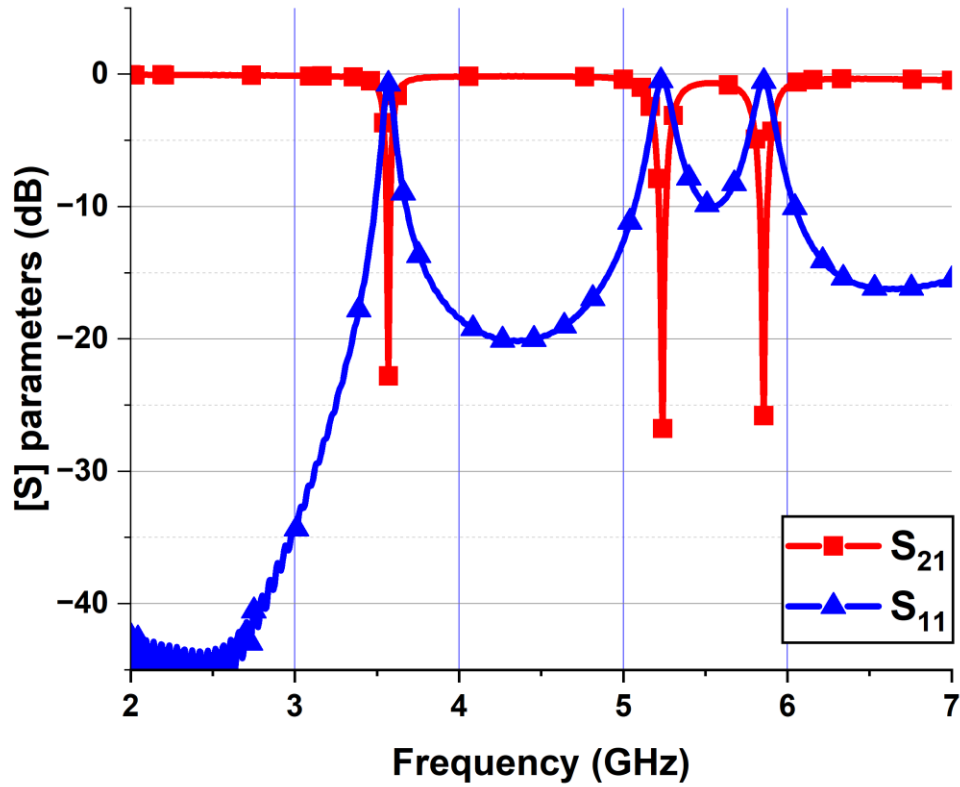


Figure III-19: Simulated S-parameters of the reconfigurable filter in its triple-mode: case 000.

Table III-3 The extracted parameters of the seven cases.

Cases	Resonant frequency (GHz)	Rejection level (dB)	Mode of the filter
011	3.6	-22	Single-band-rejection
101	5.2	-29	Single-band-rejection
110	5.8	-24	Single-band-rejection
001	3.6/5.2	-23 / -30	Dual-band-rejection
010	3.6/5.8	-23 / -24	Dual-band-rejection
100	5.2/5.8	-26 / -26	Dual-band-rejection
000	3.6/5.2/5.8	-22 / -27 / -25	Triple-band-rejection

III.5. Conclusion

This chapter presented the design and analysis of a compact reconfigurable microstrip stopband filter. Miniaturization was achieved using metamaterial-inspired Split Ring Resonator (SRR) unit cells, while frequency reconfigurability was enabled through six strategically positioned PIN diodes. The filter was designed and evaluated in CST Microwave Studio for rejection frequencies at 3.6 GHz, 5.2 GHz, and 5.8 GHz. By controlling the diode states, seven operating modes were obtained, including three single-band, three dual-band, and one triple-band stopband configurations. Simulation results demonstrated stable resonant frequencies, high selectivity, and reliable performance across all modes, highlighting the filter's suitability for compact and adaptable wireless communication applications.

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GENERAL CONCLUSION:

General conclusion

This dissertation has presented a comprehensive investigation into the design, optimization, and application of advanced metamaterial unit cells, culminating in the successful realization of an electronically reconfigurable microstrip stopband filter. The principal innovation of this research lies in the engineering of novel metamaterial building blocks that exhibit superior electromagnetic performance and physical compactness compared to conventional structures. By strategically integrating PIN diodes into these optimized unit cells, dynamic reconfigurability of both the filter's operational frequency and its mode has been achieved. This architecture allows for rapid adaptation of the filter's response while maintaining robust performance metrics and an ultra-compact physical footprint across all operational states.

Consequently, the developed reconfigurable filter demonstrates strong potential for integration into modern multi-standard wireless systems, enabling dynamic interference mitigation and coexistence enhancement across several key applications. The rejection at 3.6 GHz targets sub-6 GHz 5G bands, while the suppression at 5.2 GHz and 5.8 GHz addresses WLAN systems and ISM-band applications widely used in IoT and short-range wireless communications. This adaptability allows the proposed design to operate efficiently in heterogeneous and spectrum-congested environments, making it well suited for advanced RF front-ends, cognitive radio systems, and next-generation wireless devices.

To systematically address these objectives, the research was structured into three foundational chapters:

- **Chapter 1** established the theoretical and historical framework of metamaterials. It explored their origins, detailed fundamental classifications—such as Epsilon-Negative (ENG), Mu-Negative (MNG), and Double-Negative (DNG) media—and highlighted their diverse applications across the electromagnetic spectrum.
- **Chapter 2** provided the requisite microwave engineering background, offering a detailed overview of microwave filter theory and the guiding planar technologies central to modern filter realization. It further presented a comprehensive survey of established methodologies and techniques for implementing reconfigurable filter architectures.

- **Chapter 3** detailed the primary original contributions of this dissertation. It commenced with the rigorous design and electromagnetic characterization of the novel metamaterial unit cell. Building upon this foundation, the chapter systematically outlined the development of the final electronically reconfigurable stopband filter. An in-depth analysis of the design methodology, operational principles, and PIN diode integration was provided. The chapter concluded with a comprehensive discussion of the simulation results across seven distinct operational states, successfully validating the filter's capacity to target specific frequency bands (specifically 3.6 GHz, 5.2 GHz, and 5.8 GHz) through various discrete diode switching permutations.

Future Perspectives and Avenues for Research

While the findings of this dissertation strongly validate the efficacy of combining optimized metamaterial cells with PIN diode switching to achieve dynamic frequency and mode control, several promising avenues remain for further research and performance enhancement:

- **Implementation of Continuous Tuning Mechanisms:** Although PIN diodes provide highly effective discrete state switching, substituting or complementing them with varactor diodes could introduce continuous tuning capabilities. This integration would enable finer, granular adjustments of the resonant frequencies, significantly enhancing the filter's adaptability to dynamic and unpredictable signal environments.
- **Expansion to Higher-Order Multi-Band Functionality:** The current design successfully demonstrates single-band, dual-band, and triple-band operational modes. Future iterations could logically extend this architecture to support quad-band or even penta-band functionality. Broadening this multi-band capacity would dramatically increase the filter's versatility within complex communication networks, allowing for simultaneous operation across a wider array of wireless standards.
- **Adaptation to Alternative Microwave Components:** The underlying design principles and novel metamaterial concepts developed in this work are broadly applicable beyond stopband filters. Future research could adapt these structures to engineer other critical reconfigurable RF devices, including bandpass filters, phase shifters, and intelligent radiating elements for smart antenna arrays. This could yield a new suite of compact, high-performance solutions for next-generation telecommunications.

- **Scaling to Millimeter-Wave and Terahertz Domains:** As global telecommunication infrastructures rapidly migrate toward the higher frequency spectra required for advanced 5G, emerging 6G networks, and next-generation radar systems, scaling these reconfigurable designs to millimeter-wave (mmWave) and terahertz (THz) bands is a crucial next step. Such an extension will require meticulous re-optimization of unit cell geometries, careful selection of low-loss substrate materials, and the utilization of specialized high-frequency switching elements to ensure optimal, low-loss electromagnetic performance in these challenging frequency regimes.