Study of compact planar antennas for multi and wide band applications

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Study of Compact Planar Antennas for Multi and Wide Band Applications

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*Dedicated to my beloved parents
& my dear friend Manaswini*

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Abstract

This thesis explores different compact planar antenna structures and their applicability in the field of microwave engineering. Miniaturized compact antennas that can provide wide bandwidth, multiple bands of resonance at desired frequencies, good return loss, high gain have always been in demand in the antenna industry. Keeping these desired aspects in mind, the work presented aims at proposing novel antenna designs that have improved performance characteristics in comparison to the ones reported in the literature till date. The overall contribution can be divided into three parts.

In the first part, an analytical procedure called singular perturbation method has been performed for a double periodic transmission line structure with both inductance and capacitance modulated. The method has been employed to analyze a double periodic composite right/left handed (CRLH) leaky wave structure. Numerical expressions have been derived from which we have obtained the dispersion characteristics, propagation constant and radiation characteristics of the structure. We have also included the modulation of the permeability along with permittivity which increases the degree of freedom of the analysis. The radiation efficiency has been plotted as a function of antennas length to illustrate the effect of modulating indices on the antenna performance. The theoretical, simulated and the measured results seem to be in good agreement, thus justifying the application of the singular perturbation method on multiple scales for similar kind of structures.

The next work proposes a novel leaky wave antenna design based on CRLH metamaterials which has been implemented on a substrate integrated waveguide(SIW) platform. It operates

at a frequency range of millimeter wave 30-50 GHz. The SIW design aspects and its equivalent circuit have been studied. The proposed structure has been simulated using commercially available simulation tool CST MWS. The dispersion characteristics have been obtained to identify the left handed(LH) and the right handed(RH) frequency ranges. The frequency responses and the radiation patterns are simulated and the beam angle scans from -32° in the LH region to 61° in the RH region. The structure has a very wide bandwidth and is very compact. This structure has been extended to be implemented on half mode substrate integrated waveguide(HMSIW). The results of this HMSIW are also in close agreement with its SIW counterpart.

The last part of the work proposes an octa-mode substrate integrated waveguide (OMSIW) which is one eighth sector of a square waveguide resonator. The OMSIW is realized by bisecting the quarter mode substrate integrated waveguide (QMSIW) into two parts along the fictitious quasi- magnetic wall when it operates at TE_{101} mode. The OMSIW can almost preserve the field distribution of QMSIW and can achieve broadside radiation from the dielectric aperture of the OMSIW at 2.4 GHz. Another design is presented which involves etching a C-shaped slot on the top surface of the OMSIW. This contributes to an additional resonance from the structure at 5.8 GHz which finds applications in commercial multiband communication systems. The structures were designed, fabricated and measured. The measurement results match well with the simulation results. In order to study the effect of a complementary split ring resonator (CSRR) on the proposed OMSIW structure, a triangular CSRR was etched on the top surface of the OMSIW. Simulation and measured results indicate that the triangular CSRR introduces a resonance below the cutoff frequency of the OMSIW, which is desirable for many wireless mobile applications requiring compactness and high radiation performance.

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Chapter 1

Introduction

1.1 Background and motivation

The topic of antennas has been of research interest ever since its first discovery in 1888. Antennas have found wide applications in different fields, making them an indispensable component of any communication system. With the progress in technology, modern day antennas have replaced the initial bulky and less efficient antenna models.

The second half of the 20th century saw the development of planar antennas. It was seen that traditional antenna types could be implemented on a planar platform. This opened up newer avenues for antenna research. The planar implementation was done mainly on transmission line (TL) platforms like microstrip, coplanar waveguide (CPW) etc. Deschamps proposed the idea of microstrip radiators and was first implemented by Howell and Munson [1]. Ever since then, extensive research has been carried out in the field of microstrip antennas. Successful attempts were made to exploit their numerous advantages such as lightweight, low volume, low cost, conformal configuration, compatibility with integrated circuits and so forth.

A typical microstrip configuration consists of two parallel metal plates separated by a dielectric substrate [1]. One side is regarded as ground plane and the other plate acts as the radiator. With the invent of this configuration, different designs and ideas have been proposed and implemented for diverse applications. Various materials, alignment techniques have been explored by the researchers in order to improve the existing designs.

There is always a constant effort to improve the performance characteristics of the antennas. The parameters of interest include size of the antenna, frequency response, bandwidth, radiation pattern, gain and efficiency. Size reduction, also known as antenna miniaturization, requires the antenna size to be reduced while retaining its existing characteristics. The aforementioned aspects make antenna research an interesting and a challenging process. One of the topics that has attracted the attention of antenna researchers in recent times is metamaterials. Notwithstanding its applications in the fields of optics, cloaking etc, metamaterials have made a substantial contribution in the field of microwaves.

1.1.1 Basics of metamaterials

It was in early 1898 when Jagadis Chunder Bose conducted the first microwave experiment on twisted structures that were actually artificial chiral elements. In 1914, Lindman worked on artificial chiral medium by embedding many randomly oriented small wire helicals in the host medium. In 1948, Kock made light weight microwave lens by arranging conducting spheres, strips periodically and effectively tailoring the effective refractive index of the artificial media. Ever since then, artificial complex materials have been a matter of research interest for many investigators. The idea of composite materials in which both permittivity and permeability are negative at certain frequencies have been drawing the research attention.

Veselago theoretically investigated plane wave propagation in a material whose permittivity and permeability were assumed to be negative [2]. His studies showed that these materials

exhibit certain different properties in comparison to the ones that the conventional materials display. However, Veselagos work was theoretical. Actual means of implementing these kinds of composite structures were proposed in the late 2000s by D Smith et al. and Pendry et al. [3, 4]. Then, Itoh et al. proposed the concept of composite right/left handed transmission lines[5]. Many research groups all over the world are now studying various aspects and classes of metamaterials. Several ideas and suggestions for future applications of these materials are under investigation.

The first left handed medium was realized as a combination of conducting wires and split ring resonators (SRRs) as shown in Figure 1.1 [6]. However, these structures are resonant, and consequently exhibit high loss and narrow bandwidth [7]. A structure made of resonating elements generally does not constitute a good transmission medium for a modulated signal because of the quality factor intrinsically associated with each resonator.

Figure 1.1: Revolutions on the first experimentally realization of MTMs. (a) Thin-wire structure exhibiting negative- ϵ . (b) Split-ring resonator structure exhibiting negative- μ [8]

Composite right/left handed (CRLH) transmission lines (TLs) which are realized by periodic loading of series capacitance and shunt inductance in a host medium are becoming an interest-

ing and promising topic of research. Since the resonant frequency of CRLH TLs is independent of the physical length [9], the TLs are being used to design many compact devices for communication system applications, like small leaky-wave antennas (LWAs) and compact filters. Another advantage of CRLH TLs is the unusual propagation property of backfire- to-endfire (including broadside) radiation. It can be designed to radiate backward in the left-handed (LH) region and forward in the right-handed (RH) region. Most of the well-known CRLH TLs are periodically loaded with uniform inductance and capacitances (Figure 1.2).

Figure 1.2: Equivalent circuit model of a (a) forward and (b) backward transmission lines. The T-circuit models of the basic cell structures are also indicated in (c) and (d). Equivalent circuit model (basic cell) of the CRLH transmission line (e) [8]

The CRLH structures offer the following advantages [9].

- *Reduction in size.* CRLH TL can be constituted by nondispersive (independent of frequency) elements. Compact devices for communication system applications, like small leaky wave antennas(LWAs) and compact filters are realized based on CRLH TLs since the resonant frequency is independent of the physical length.
- *Bandwidth enhancement.* For the balanced condition of the CRLH TLs, the real CRLH dispersion curve is, precisely the superposition of the LH hyperbolic dispersion function (dominant at lower frequencies) and of the RH linear dispersion function (dominant at higher frequencies). When the hyperbolic linear dispersion characteristic of the CRLH TL is judiciously combined with the linear dispersion characteristic of the conventional pure RH TL section, it is possible to maximize the bandwidth of the component. This is referred as dispersion engineering.
- *Dual-band operation.* Because unbalanced CRLH TLs exhibit a dispersion relation satisfying the double resonance condition where the two frequencies and the two propagation constants are arbitrary, this kind of CRLH TLs with the dual-band property can be used to transform virtually any microwave TL-based component into a dual-band component [10].

One of the effective means of realizing CRLH TLs is through periodic structures. A periodic array of conducting elements can behave as an effective medium for electromagnetic scattering when the wavelength is much longer than both the element dimension and lattice spacing. Thus, CRLH TLs proposed till date have been realized as periodic structures on different transmission line platforms.

1.1.2 Microstrip implementation

Microstrip has been one of the competent platform to realize CRLH TLs. It offers the advantages of ease of manufacturing, lower fabrication cost, efficient radiation and it is easy to integrate with other microwave components on a circuit. The initial CRLH designs were implemented on this platform owing to these advantages. The first distributed CRLH TL was implemented on microstrip line, with series interdigital capacitors and shunt short-circuited stub inductors connected to the ground plane as shown in Figure 1.3 [9].

Figure 1.3: (a) Microstrip CRLH TL using series interdigital capacitors and short-circuited stub inductors, (b) CRLH TL LW antenna showing the three regions: backward, broadside, and forward [9].

This structure is based on microstrip line, with series capacitance and parallel inductance per unit length added to make an artificial transmission line. It was first shown in [11] that such a structure may be used to achieve a continuous scanning from the backward to the forward region, passing through broadside. One interesting feature of this design is that the fundamental quasi-TEM mode of the microstrip line becomes a fast wave due to the loading, so that radiation occurs from the fundamental $n = 0$ space harmonic. The $n = -1$ space harmonic is a slow wave (nonradiating wave) for this design. If the period is small enough, the structure acts as a quasi-uniform 1D structure, which supports a fast wave.

It was experimentally proven that while implementing the CRLH TL on the planar plat-

form, the inductances can be realized in the form of via holes connecting the top metal plate and the ground plane, while the capacitances can be realized by interdigitated gap patterns. The microstrip line inherently has series inductance(L_R) and parallel capacitance (C_R) which contribute to right handed behaviour. The LH behaviour is induced by adding extra series capacitance (C_L) and parallel inductance (L_L).

1.1.3 Substrate integrated waveguide (SIW) implementation

There has been a constant effort to develop better platforms for realizing planar microwave systems. Currently microstrip and coplanar waveguide(CPW) technologies have been the most popular choices for fabricating planar systems. However, it has been observed that with increase in frequency, these platforms pose problems of transmission and radiation losses. This calls for a need for different means of implementing planar technology.

Recently, a new concept for implementing microwave planar components known as Substrate Integrated Waveguide has been proposed [12]. This technique integrates non planar 3D structures within a uniplanar circuit. Ever since its introduction, substrate integrated waveguide has been widely developed for integrated microwave and mm wave components and antennas.

Figure 1.4: Basic SIW layout

Basically, SIW are integrated waveguide like structures fabricated by using two rows of conducting cylinders or slots embedded in a dielectric substrate that electrically connects two parallel metal plates. This way of realization facilitates the implementation of nonplanar rectangular waveguide in a planar form. These conducting cylinders keep the fields confined within the structure.

Figure 1.4. shows a typical implementation of a basic substrate integrated waveguide. The dimensions d is the diameter of the metal vias, w represents their transverse spacing and s is their longitudinal spacing. These SIW structures have similar propagation characteristics as that of classical rectangular waveguides [13]. These were initially called post wall waveguides or laminated waveguides and were used for feeding networks in antenna arrays. But now, SIW technology has been applied to implement different microwave devices.

The main advantage of SIW which is beneficial to CRLH implementation is its ability to provide negative permittivity when operated below cutoff frequency. Currently, realizing CRLH on SIW is gaining lot of research interest. Different CRLH structures especially LWA are being tried out on SIW. CRLH structures have already been tried out on microstrip and coplanar waveguide.

It is challenging to design lumped series capacitance, which is a necessity for CRLH structures. This is overcome by etching interdigital capacitance on the surface of SIW for large series capacitance. Basic principle of implementing CRLH on SIW is by using series interdigital capacitors and via holes for shunt inductors.

CRLH TLs are realized by using series interdigital capacitors and via holes for shunt inductors. The values of the inductance and capacitance can be suitably controlled by adjusting the distance between two vias, width of the interdigital structure, number of fingers etc. There have been many SIW slot antennas that have been developed for specific applications, all of which are operated in the RH region [14].

1.2 Objectives

It has been seen in the previous sections that the topic of metamaterial antennas offer lot of scope to improve the existing performance characteristics of the existing antennas. Also, the miniaturization of antennas is always in demand. Keeping these criteria in mind, the current research focuses on simpler designs to implement leaky wave antennas (wideband) and resonator antennas (multiband). The objective of this thesis is to present a physical insight into the proposed structures in addition to the proposed designs. It attempts to balance the concepts as well as the applicability of the proposed antenna structures. The main goals of the thesis can be summarized as follows.

- 1) To study the basic concepts of the metamaterials and their applicability to the field of microwaves especially antennas. It is very essential to know the theories of composite right left handed transmission lines and its various means of implementation. In order to identify the research gap in this area, the existing designs need to be reviewed. This review also aims at investigating the pros and cons of the CRLH TLs as compared with SRR and other implementation means. The analytical methods for improvement and the future direction of research aid the motivation for the proposed research.
- 2) To obtain a physical insight into the antenna structures theoretically. Studying the propagation characteristics of the wave as it travels in the antenna is necessary to gain better understanding of the structure. It helps in identifying the shortcomings and makes way for antennas with improved designs. Manually solvable numerical methods have the advantage of providing a basis for the understanding of wave propagation in the structure, along with giving accurate numerical data as a benchmark result against which approximate ones can be judged. In this regard, an analytical technique called singular perturbation method is employed to understand the impact of varying the reactive parameters of the circuit as the wave propagates through it.

3) Based on the review of CRLH LWAs implemented on the SIW platform, following were identified as areas that needed improvement.

- Need for simpler slot design on the CRLH SIW TL.
- Wider bandwidth of the overall LWA
- Better values of gain at higher frequencies
- Size reduction of the LWA

In order to address these concerns, a novel CRLH SIW LWA with V shaped slot was proposed. Owing to CRLH characteristics, it operates below its cut-off frequency and provides a continuous backfire to endfire scanning with change in frequency.

4) The review of substrate integrated waveguide applications leads to exploring its adaptation to resonator antennas. The SIW resonator antennas are similar to microstrip patch antennas except for the fact that it has metal via holes that connect to the ground plane through the substrate. However, it has more advantages compared to the traditional patch antennas. These ones are more compact at their operating frequencies. When dissected along symmetric planes, they allow further miniaturization of the structure without compromising on its performance and resonant frequencies. Also, from concept perspective, they are very attractive since they provide lot of scope for implementing various microwave components. The SIW resonator can be reduced to half of its original size by cutting it along a fictitious line and the resulting structure is called half mode SIW (HMSIW). If this HMSIW is further bisected, it gives a quarter mode SIW (QMSIW). Proceeding with this theory, the QMSIW can be cut further into half to yield an octa mode SIW (OMSIW) which has been explored extensively in this work.

5) Continuing with the concept of OMSIW, the following research gaps were identified from the existing works.

- The resonant frequencies were not tuned to usable frequency bands.
- The resonances had narrow bandwidth.
- Need for a parameter to tune the frequencies other than the SIW patch dimensions.

These gaps in the existing reported works on SIW resonator antennas offered a scope for improvement of the structures. These issues are overcome in this work with the introduction of slot structure on the OMSIW antenna. The slots provide an extra degree of freedom in the OMSIW slot antenna designs.

These are the specific objectives of this thesis. In general, the thesis also aims to open new avenues for further research in these areas.

1.3 Major contributions of the thesis

The research outcome achieved in this thesis includes designs of size miniaturized and high performance wide/multiband planar antennas implemented mainly on substrate integrated wave guide platform. It also involves an analytical technique which has been extended to include two modulated parameters and is validated by applying to a double periodic microstrip based LWA structure. The major contributions of the thesis can be summarized as follows.

1) Numerical analysis technique called Singular Perturbation Method has been extended to accommodate an extra parameter variation. This method provides the dispersion characteristics and illustrates the left-handed and right-handed regions of operation. The effectiveness of the method has been validated by applying it to a double periodic LWA structure which is loaded with alternate capacitances and inductances. The numerical results obtained from

the analysis have been compared with the simulated results of the structure and they seem to be in good agreement. The radiation efficiency and the pattern results were also obtained from the analysis. In order to substantiate these results, the prototype of the double periodic LWA structure was measured to procure the radiation pattern of the structure. The results have been plotted and they match.

- 2) A wide bandwidth leaky wave antenna based on CRLH has been implemented on a SIW platform. The structure is designed to operate at a high frequency range between 30-50 GHz. V-shaped slots were implemented on the top surface of the proposed SIW. The elaborate design procedures of the SIW and the slot have been discussed. The simulated and the measured frequency response indicate wide bandwidth of almost 15 GHz with good gain values. The proposed antenna offers a continuous backfire to endfire scanning with broadside radiation.

The structure has been adapted to the half-mode SIW (HMSIW) implementation while retaining the entire V-shaped slots. This proposed antenna provides backward and forward radiation with a stop band in between. The antenna has been fabricated and the frequency response has been measured. They seem to be in good agreement. The simulated efficiency results indicate better efficiency for the HMSIW structure when compared to its SIW counterpart.

- 3) An octamode substrate integrated waveguide (OMSIW) resonated antenna that operates at 2.4 GHz has been proposed. Its evolution from SIW resonator antenna has been studied in depth and the design techniques have been provided. A thorough comparison of the structure with its predecessor designs has been made to highlight the advantages of the proposed resonator antenna. The antenna has also been experimentally fabricated to verify its simulation performance. It is seen that the measurement results agree with the simulated ones, thus making it suitable for commercial 2.4 GHz applications.

4) In order to study the impact of the slot on the proposed OMSIW antenna, two slot types were etched on the top patch surface of the proposed OMSIW.

- The first design implements a C-shaped slot on the OMSIW antenna. It was seen that this C-shaped slot added to another resonant frequency at 5.8 GHz, while retaining its original 2.4 GHz resonance. This makes the structure suitable for commercial 2.4 GHz and 5.8 GHz applications. This structure is fabricated and the measured results match the simulated ones.
- The second design involves a triangular shaped complementary split ring resonator (CSRR) implemented on OMSIW antenna. The triangular CSRR has the advantages of compactness and broader bandwidth in comparison to its square CSRR counterpart. Also, the CSRR introduces an extra frequency band below the original cut off frequency of the OMSIW resonator at 1.8 GHz. This demonstrates the LH behaviour of the antenna. The CSRRs contribute to the metamaterial features of the structure. The structure has been fabricated and the measured results of frequency response and radiation patterns match the simulated ones.

1.4 Organization of the thesis

The thesis has been organized into seven chapters. The first chapter primarily explains the motivation behind this work by identifying limitations of previous designs, and scope of improvement. It establishes that the proposed solution, in principle, can overcome most of those problems of traditional structures.

Chapter 2 provides the necessary background to understand the evolution of metamaterials and their applications to leaky wave antennas. The types of metamaterials have been discussed with main focus on composite right/ left handed transmission lines as applied to periodic

structures. The contributions of different research groups on these topics have been presented. The CRLH LWA implemented on two platforms of microstrip and SIW have been discussed. A scheme of numerical technique for analyzing periodic structures called singular perturbation method has been provided. The chapter concludes with the discussion of SIW based resonator antennas.

Chapter 3 presents singular perturbation analysis implemented by modulating two parameters simultaneously. In order to analyse the radiation characteristics of the double periodic composite right/left handed (DP-CRLH) TL proposed in [15], a singular perturbation procedure has been used. The leaky-wave (LW) characteristics from the fast wave region in the DP-CRLH TLs is implemented by the microstrip loaded with double periodic interdigital capacitors and shunt short-circuited stub inductances. The analysis has been done with the help of the singular perturbation procedure. The new LW region is found below the left-handed (LH) passband for the DP-CRLH TLs based on this analysis method. The analysis shows that wave radiates forward and the TLs work as a right-handed material (RHM) in the new LW region. Dispersion characteristics and the radiation efficiency are investigated with the help of the singular perturbation procedure. The radiation efficiency is numerically estimated as a function of the antenna length, and the radiation angle is estimated in the LW region. Validating experimental results have been provided too.

Chapter 4 presents the implementation of CRLH LWA on substrate integrated waveguide and half mode substrate integrated waveguide (HMSIW) platforms. We discuss the design of v shaped slot leaky wave antenna implemented on a SIW platform which radiates in both backward and forward directions. The design procedures of the SIW and the slot are discussed and the measured frequency responses have been presented.

Chapter 5 focuses on the substrate integrated resonator antenna which is designed to work at octa mode. It provides the necessary background of the SIW square resonator and its

subsequent extension and adaptation to obtain the proposed octa mode SIW. It discusses the structure details, field distributions and resonant frequency modes of half mode, quarter mode and octa mode SIW resonators. It gives the steps involved in the design of the proposed antenna structures while elaborating on the design procedure starting from the choice of substrate to obtaining a desired frequency of resonance of the proposed resonator. It also focuses on the implementation of a quarter mode substrate integrated waveguide antenna whose design has been extended to obtain the proposed octa mode resonator antenna. The simulated and the measured results are discussed. The proposed octa mode resonator antenna that operates at dominant resonant frequency of 2.4 GHz is discussed. The variation of various design parameters and its impact on the structure have been presented. The last Section concludes with the performance comparison of the proposed structure with its counterparts.

Chapter 6 deals with OMSIW resonator antenna with two different types of slots etched on the top patch surface. Two types of slot structures have been discussed. The first one is a C-shaped slot which is realized along the feed line side of the OMSIW structure. Initial sections discuss the structure in detail along with its simulated and fabricated results. Different variations of this slot have been studied and the simulated results have been presented. The later part of the chapter focuses on a triangular CSRR slot etched on the OMSIW. The design details of the structure have been discussed while providing the simulated and measured results.

Chapter 7 concludes the thesis by summarizing the contributions of this research. Few suggestions for future explorations leading to possible improvements/extensions of the proposed work have been listed in this last chapter.

Chapter 2

Literature Review

This chapter provides the necessary background to understand the evolution of metamaterials and their applications to wideband and multiband antennas. The types of metamaterials have been discussed with main focus on composite right/left handed transmission lines as applied to periodic structures. The contributions of different research groups on these topics have been presented. The CRLH LWA implemented on two platforms of microstrip and SIW have been discussed. A scheme of analytical technique for analyzing periodic structures called singular perturbation method has been provided. The chapter concludes with the discussion of SIW based resonator antennas.

2.1 Metamaterials

Metamaterials can be defined as manmade artificial composite structures constructed in such a way that their electric permittivity and permeability are negative, which results in the display of unique properties not exhibited by conventional materials. This concept was first proposed theoretically by Veselago in 1968 [2]. The first experimental realization was achieved by incorporating the well-known split-ring resonators (SRRs) and metallic thin-wires in combination

[16] [17]. Metamaterials usually encompass different classes of structures, which when brought together, exhibit unusual properties not readily seen in nature. Hence metamaterials include a large class of different structures. So the term metamaterials has been used for different arrangements of various structures[6, 18, 7, 19, 8, 20, 21] . The various structures reported till date, which can be classified under metamaterials are [2]

- DNG(Double Negative)
- Left handed transmission lines
- NRI (Negative Refractive Index)
- Magneto-optic materials
- Soft and hard Surfaces
- High Impedance Surfaces
- Artificial Magnetic Conductors

Amongst many other applications, these structures are used in practice to design compact microwave components like leaky-wave antennas (LWAs), filters, couplers and many other applications.

2.1.1 Background of metamaterials

The concept of materials with both negative permittivity and permeability was first proposed in literature by Veselago in 1967 [2]. His work focussed on the theory behind plane wave propagation in a material whose permittivity and permeability were assumed to be negative. It was predicted that these type of materials have properties that are not displayed by naturally occurring materials. The investigation of implementing these kind of materials gained momentum in late 1990s. Three research groups almost simultaneously proposed actual means of engineering materials with both permittivity and permeability negative [16, 17, 22].

These materials were designed by individually realizing negative ε and negative μ , with their combination resulting in metamaterials. Caloz and Itoh [9] proposed the concept of composite right/left handed transmission lines. Ever since then the microwave applications of these materials are being explored by many research groups. They are being implemented to realize various microwave components like filters, couplers, antennas and so on.

2.1.2 Characteristics of metamaterials

The quadrant system gives a better insight of metamaterials where permittivity is plotted against permeability. We can differentiate four regions based on four quadrants as shown in Figure 2.1.

Quadrant I : DPS- Double Positive

Quadrant II : ENG-Epsilon Negative

Quadrant III: DNG- Double Negative

Quadrant IV: MNG- Mu Negative

It can be seen that most natural materials occur at first quadrant where the μ and ε have positive values. Few of the naturally occurring electric plasma and magnetic plasma materials fall in the second and fourth quadrants [23, 24, 13]. The best-known examples of media with negative permittivity are low-loss plasmas, metals and semiconductors at optical and infrared frequencies (sometimes called solid-state plasmas) [25].

 Figure 2.1: Classification of materials based on μ and ϵ

Ferrites which are magnetized upto saturation, present a negative magnetic permeability near the ferrimagnetic resonance (which usually occurs at microwave frequencies) [26]. Media with negative permeability are not so commonly found as compared to the ENG in nature due to the weak magnetic interactions in most solid-state materials [27]. In the $\epsilon - \mu$ domain, the point $\mu = 0$ and $\epsilon = 0$ represent a nihility, which can support a perfect tunneling effect; the line $\epsilon_r = \mu_r$ in both DPS and DNG regions denote airline, which have perfect impedance matching with air.

Metamaterials have attracted a lot of research attention because they exhibit properties that were different from existing ones. The characteristic properties of metamaterials that differ from the conventional systems are [28, 9, 8, 20, 21]

- 1) The double negative nature of the constitutive parameters, ϵ and μ , results in the propagation of electromagnetic waves exhibiting anti parallel phase and group velocities.

RH medium: $v_p > 0$ ($\beta > 0$) and $v_g > 0$; LH medium: $v_p < 0$ ($\beta < 0$) and $v_g > 0$, where v_p is the phase velocity and v_g is the group velocity.

- 2) *Necessary frequency dispersion of the constitutive parameters.* A frequency-dispersive or simply dispersive medium is a medium in which the propagation constant (β) is a nonlinear function of frequency, which results in frequency-dependent group velocity and leads to distortion of modulated signals. In a dispersive medium, simultaneously negative ε and μ are allowed, as long as the frequency dependent ε and μ satisfy the necessary entropy conditions [29].
- 3) *Reversal of Doppler Effect.* The Doppler frequency of a receding source is blue-shifted, whereas that of a proceeding source is red-shifted. This phenomenon of reversal of Doppler Effect in a LH medium was pointed out by Veselago back in 1967 as an immediate consequence of left-handedness[30].
- 4) *Reversal of Vavilov-Cerenkov radiation.* Vavilov-Cerenkov radiation is the visible electromagnetic radiation emitted by liquids and solids when bombarded by fast-moving electron beams with high velocity. It has been seen that, in a RH medium, Vavilov-Cerenkov radiation occurs at positive θ angles (in the half-plane in the direction of motion of the particle), since $\beta > 0$. If in contrast the medium is LH, $\beta < 0$, radiation occurs at negative θ angles (in fact the half-plane in the direction of motion of the particle but opposite to the group velocity of the beam including it) [31].
- 5) *Reversal of the boundary conditions relating the normal components of the electric and magnetic fields at the interface between a conventional/right-handed (RH) medium and a LH medium.* Thus the tangential components of electric and magnetic field remain continuous while their normal components become antiparallel at the interface between a RH medium and a LH medium .
- 6) *Reversal of Snells law.* A wave incident upon the interface between two media with different Handedness (RH and LH) undergoes negative refraction, corresponding to negative refraction angle or NRI [19, 32].

2.1.3 Classification of metamaterials

Metamaterials can be broadly classified as resonant and nonresonant types. This classification is made based on the means of implementation.

Figure 2.2: Revolutions on the first experimentally realization of MTMs. (a) Thin-wire structure exhibiting negative $-\epsilon$. (b) Split-ring resonator structure exhibiting negative $-\mu$. (c) Bidimensionally LH structure exhibiting negative $-\epsilon$ and μ . [33]

Veselago predicted and proposed only theoretical concept of metamaterials. However, unlike as expected by Veselago, LH material was not a naturally occurring material. Almost after a few decades, practical means of realizing metamaterials was proposed by Pendry et al at Imperial College London. They introduced resonant structures that contributed negative permittivity and negative permeability individually. Also, these structures were effectively homogenous [34]. Features of metamaterials beyond the negative refraction are studied and the gradient refractive index medium was realized to bend electromagnetic waves in 2005 [3], including the exciting discovery of three-dimensional optical MTM with a negative refractive index (NRI) [29] and the optical transformation making invisible cloaks [33, 22, 35, 36].

It was found that the negative ε can be realized in the form of thin wires and negative μ is achieved in the form of metal split ring resonators (SRR). Smith et al combined the thin wire and SRR structures to form the composite structure as shown in the Figure 2.2. It represented the first experimental LH metamaterial prototype. The frequency ranges of thin wire and SRR were chosen such that they overlapped. This structure exhibited simultaneous negative permittivity and negative permeability allowing the presence of a passband in the desired frequency range [9].

These resonant type of metamaterials have high quality factor but they compromise heavily on losses and have narrow bandwidth. Even though they are very exciting from a physics perspective, they did not seem to offer much prospect as far as microwave applications were concerned. It has been observed that a modulated signal cannot be transmitted efficiently through a resonating propagating medium. Therefore, this necessitated an alternative approach that could satisfy the desired criterias for microwave applications.

Due to these shortcomings of the resonant-type LH structures, there existed a need for other methods of realization. Thus, by identifying the similarity between LH waves and conventional backward waves, three research groups almost simultaneously, introduced a transmission line (TL) approach of metamaterials which will be elaborated in the next section.

2.2 Composite Right/Left-Handed (CRLH) Metamaterials

The concept of composite right/left-handed (CRLH) MTM, introduced by Caloz et al. in 2004 [6, 18, 37, 19]. LHM has attracted growing attention in both theoretical exploration [6, 7, 38, 39, 34, 32, 31, 40, 41, 42] and experimental study [37, 43, 44, 45, 46, 47, 48, 49, 50, 51, 52]. A conventional transmission line is a combination of inductances and capacitances as a result of varied voltages and currents. It displays right handed behaviour. It was observed that as wave propagates along the structures, the associated currents and voltages induce other natural

effects.

As the currents flow along the series capacitance C_L , it induces magnetic flux and hence contributes a series inductance L_R . Similarly, the voltage gradient between the metal surface and the ground plane gives rise to a shunt capacitance C_R in addition to shunt inductance L_L . As a consequence of these phenomena, it was shown that a purely LH structure cannot exist in nature, since a real LH structure necessarily includes (L_R, C_R) contributions in addition to the (L_L, C_L) reactances. This lead to the coining of the term "composite right/left-handed" (CRLH) transmission line (TL).

2.2.1 CRLH circuit model

Figure 2.3: Equivalent circuit of the unit cell of (a) Pure RH TL, (b) Pure LH TL, (c) CRLH TL.

In order to understand the characteristics of CRLH TL, it is important to analyze its circuit parameters. Figure 2.3 shows the unit circuit model of the CRLH TL. When the circuit is operating at lower frequencies, the series inductance(L_R) and the shunt capacitance (C_R) are

short and open respectively. This reduces the equivalent circuit to just series capacitance(C_L) and shunt inductance (L_L) which contribute to left handedness(antiparallel phase and group velocities) and causes the circuit to have high pass behaviour. At high frequencies, C_L and L_L tend to be short and open, respectively, such that the equivalent circuit is essentially reduced to the the series inductance(L_R) and the shunt capacitance (C_R), which is RH(parallel phase and group velocities). This RH circuit is of lowpass in nature. Above a certain cutoff, a RH stopband is present[53]. The series resonant frequency is represented by ω_{se} and the shunt resonant frequency is represented by ω_{sh} . They are related to the circuit parameters as shown below.

$$\omega_{se} = \frac{1}{\sqrt{L_R C_L}}$$

$$\omega_{sh} = \frac{1}{\sqrt{L_L C_R}}$$

Figure 2.4: Dispersion diagram for a purely LH (PLH) structure and for a purely RH (PRH) structure based on circuit parameters [9]

The combined effect of CRLH on dispersion characteristics is very different from individual effect of pure LH and pure RH systems. According to properties of slow-fast wave effect and right-left handed propagation, the dispersion diagram in Figure 2.4 is divided into four different

regions [20].

Region I: Left-slow wave region. This region supports guided wave with phase and group velocities are anti-parallel with each other.

Region II: Left-fast wave region: Wave propagation in this region is naturally leaky, and phase and group velocity are anti-parallel with each other, and leaky wave propagates backwards.

Region III: Right-fast wave region: It is right handed leaky wave region as opposed with Region II, phase velocity and group velocity are parallel with each other, the leaky wave in this region propagates forward.

Region IV: Right-slow wave region: This is normal guided wave region with parallel phase and group velocity.

Different regions are used for different applications. Backward and forward propagation leaky wave antenna can be engineered in Region II and III, left and right handed guided wave application such as filters and guided transmission lines, fall in to region I and IV, respectively.

2.2.2 Balanced and Unbalanced CRLH TL

The dispersion diagram of Figure 2.4 identifies two curves. One for pure LH circuit and one for pure RH circuit. It also shows the way in which the CRLH dispersion curves tends to the pure LH and pure RH curves at lower and higher frequencies, respectively. There exists a predicted gap in the CRLH curve. This gap is owing to the different series and shunt resonances (ω_{se}, ω_{sh}). Two cases can be defined based on their criteria as[9]

$$case1 : \omega_{se} = \omega_{sh}$$

$$case2 : \omega_{se} \neq \omega_{sh}$$

If case 1 is true then the CRLH TL is said to be balanced and if case 2 holds good, the CRLH TL is unbalanced. For the balanced case, the CRLH gap closes up with nonzero group velocity.

The gap closes up and the characteristic impedance becomes a frequency-independent quantity which means that the balanced condition allows matching over an infinite bandwidth. The transition from LH region to RH region occurs at a transition frequency ω_0 . For unbalanced case, the impedance conditions correspond to zero group velocities. There will exist a stopband between the LH and the RH regions.

The balanced case has a few advantages over the unbalanced case. Firstly, its model is simpler than that of the general unbalanced model because series combined LH and RH contributions are decoupled from each other. Secondly, unlike the unbalanced TL, the balanced TL is gapless (no stop band). Lastly, the characteristic impedance of the balanced TL is a constant frequency independent quantity.

2.3 CRLH based periodic structures

As discussed before CRLH metamaterials can be implemented in different forms. Implementing them as periodic structures by far has been the popular means. The reason for periodicity has been more of convenience than physics [10, 28, 54, 55, 11, 5, 9]. It is simpler to build a metamaterial by repeating periodically a unique cell than by using a collection of different cells. Also, another advantage of periodic structures, in comparison to non uniform structures is that, theoretical analysis becomes easier since we can use already well established theories [56, 13]. We will discuss some of the basic concepts related to periodic structures [57, 9, 58, 59, 60, 61, 62].

As early as 1887, Lord Rayleigh proposed a new concept about the propagation of waves in periodically stratified media [63]. Since then, Kelvin and Vincent built a theory for analyzing the dispersion properties of lattices which were made up of a series of identical particles and an electric filter was designed by the same group during 19th century [64]. In 1928, the most important theory about periodic structures was found by Floquet [65]. Floquet theorem [66] (or Bloch's theorem [67]) is a basic theorem underlying the theory of wave propagation in periodic

structures. It gives a method to determine the solution of the wave equation in a periodic structure. After that, the interest in periodic structures increased. The applications including the relatively slow electron beam and antennas were designed [68].

Periodic structures can be classified into two basic types, structures with continuous but periodically varying electrical properties and structures with periodic boundary conditions as shown in Figure 2.5 . There are two extraordinary properties that make periodic structures essential, (a) In the periodic structures there are infinite number of space-harmonics with phase velocities varying from zero to infinity; (b) They can support waves with well-specified propagation and stop bands [69, 56].

The first property is important because that it expresses a separate set of passbands differentiated by stopbands. The second property is attractive since it provides an application which can support waves with phase velocities much less than the velocity of light. The metallic diaphragm in a hollow metallic wave guide behaves as a lumped capacitance and inductance added at interval, which forms the series capacitance and shunt inductance per unit length. So the negative permittivity and permeability will be realized at certain frequency of interest.

Figure 2.5: Periodic structures. (a) The structure with continuous but periodically varying electrical properties. (b) The structure with periodic boundary conditions.

2.4 CRLH TL based leaky wave antennas

The concept of leaky wave antennas (LWA) was introduced around 1940s [54, 23]. The initial realization of LWA consisted of a rectangular waveguide with a slot cut on one of its side walls (Figure 2.6). This topic has received significant contributions ever since then and in recent times the focus has been more on planar LWAs [57, 9, 58, 59, 60, 61, 62]. This is owing to the latest research interest in metamaterials which have enhanced the performance of conventional LWAs.

Figure 2.6: Rectangular waveguide with a slit cut on one of its side walls.

LWAs are quite sought after in the microwave band and above, because they can achieve high values of directivity with a simple structure, without the need for a complicated and costly feed network as typically needed in a phased array. The pattern bandwidth (defined from the radiated power density at a fixed observation angle) is usually fairly narrow. This is owing to the fact that as the beam scans with frequency change, the pattern bandwidth decreases as the beamwidth decreases. However, LWAs are often best suited for applications that can take advantage of frequency beam scanning.

Leaky wave antennas can be defined as structure that radiates a type of wave which leaks EM energy continuously as it propagates on the guiding structure [70]. It supports a fast wave on the guiding structure, where the phase constant β is less than the free-space wavenumber k_0 . A LWA is based on a guiding structure that propagates a wave in one direction, either unidirectionally (a source is placed at one end of the structure) or bidirectionally (a source is placed in the middle of the structure).

1-D LWAs can be broadly classified into two categories: uniform (or quasi-uniform) and periodic. A uniform LWA is one where the guiding structure is uniform along the length. In this case, the structure supports a wave that is fast with respect to free space. A quasi-uniform LWA works in a similar manner as a uniform one, except that a periodic structure is used to guide the wave. The periodicity, which is usually much less than a wavelength, does not affect the radiation.

A periodic LWA is one where the guiding structure supports a wave that is slow with respect to free space, so that $\beta > k_0$. The fundamental nonradiating mode is made to radiate by introducing periodicity along the length of the structure. The field on the structure is then characterized by an infinite number of space harmonics. One of the advantages of a periodic LWA is that a beam can radiate either in the forward or backward directions. In addition to the two main types of guiding structures (uniform or quasi-uniform versus periodic) there are also two ways in which an LWA can be fed: from one end structure, or from the center of the structure.

A major difficulty with periodic LWAs is that it is often difficult to obtain a good beam exactly at broadside. This is because it corresponds to an open stopband point. At the stopband point, the periodic traveling-wave antenna turns into a standing-wave antenna. This is the main limitation for achieving a continuous scanning from the backward to the forward quadrants with a single antenna[70].

This difficulty is addressed by using composite right/left handed transmission line structures implemented on different platforms. In principle, any open CRLH TL structure can operate as a LW antenna since the CRLH dispersion curve always penetrates in the radiation region [71, 59, 14, 72]

Backfire to endfire frequency-scanning capability is provided by an open balanced CRLH structure [73]. This backfire-to-endfire capability, first demonstrated experimentally in [10] and explained by the CRLH concept in [14], is a very unique feature for a LW antenna, which cannot be obtained in conventional (uniform or periodic) LW structures. Broadside radiation is a consequence of the fact that the usual gap is suppressed (by mutual cancelation of the series/shunt branches of the CLRH structure under the balanced condition). Because there is no gap, group velocity is different from zero at the origin.

Finally, a further advantage of the present CRLH LW antenna is that, operating in a fundamental mode, it can be fed by a very simple and efficient (small and broadband) mechanism, such as a simple line, in contrast to typical LW structures.

2.4.1 Microstrip implementation of CRLH LWAs

Leaky wave antennas implemented on microstrip lines was first reported by W. Menzel [74]. Microstrip line provides the benefits of simple construction, low-profiles, easy impedance matching, narrow beam-width, frequency scanning abilities and inherently more broader than resonant type antennas [75]. The CRLH TLs realized by introducing the series capacitances and shunt inductances in host mediums [72, 76, 77] can be used to design compact broadband LWAs which can radiate backward in the LH region and forward in the RH region.

(a)

(b)

Figure 2.7: (a) Microstrip CRLH TL using series interdigital capacitors and short-circuited stub inductors, (b) CRLH TL LW antenna showing the three regions: backward , broadside, and forward [9].

The first distributed CRLH TL was implemented on microstrip line, with series interdigital capacitors and shunt short-circuited stub inductors connected to the ground plane as shown in Figure 2.7. This structure is based on microstrip line, with series capacitance and parallel inductance per unit length added to make an artificial transmission line. It was first shown in [72] that such a structure may be used to achieve a continuous scanning from the backward to the forward region, passing through broadside. One interesting feature of this design is that the fundamental quasi-TEM mode of the microstrip line becomes a fast wave due to the loading, so that radiation occurs from the fundamental $n = 0$ space harmonic. The $n = -1$ space harmonic is a slow wave (nonradiating wave) for this design. If the period is small enough, the structure acts as a quasi-uniform 1D structure, which supports a fast wave.

It was experimentally proven that while implementing the CRLH TL on the planar platform, the inductances can be realized in the form of via holes connecting the top metal plate and the ground plane, while the capacitances can be realized by interdigitated gap patterns[5]. The microstrip line inherently has series inductance(L_R) and shunt capacitance (C_R) which contribute to right handed behaviour. The LH behaviour is induced by adding extra series capacitance (C_L) and shunt inductance (L_L).

Figure 2.8: (a) The microstrip interdigital(C)/stub(L) CRLH LW antenna. (b) Vialess microstrip implementation of the CRLH resonator [8]

Figure 2.8 shows a microstrip interdigital-C/stub-L SP-CRLH TL providing a unique property of backfire-to-endfire frequency-scanning capability. An alternative to via hole approach is shown in Figure 2.8. This structure is the via-less configuration which may be advantageous in processes where implementation of vias is problematic.

Recently, the leaky wave structure proposed by Itoh et al.[5] was extended to realize a double periodic LWA(DP-CRLH TLs). A TL which is periodically loaded with capacitances and inductances can be considered as an effective transmission medium provided that the dimensions of the unit cell are small compared to the operating wavelength. The model of the DP-CRLH TLs proposed in [15] consists of double periodically loaded shunt inductances L_{Li} (Figure 2.9 (a)) and double periodically loaded series capacitances C_{Li} ($i=1,2$) (Figure 2.9 (b)).

Figure 2.9: (a) DP-CRLH TL with different shunt inductances loaded. (b) DP-CRLH TL with different series interdigital capacitances loaded.

The realization of the DP-CRLH TLs requires a continuous medium with small physical dimensions provided by a host TL medium (Z_0 , k) double periodically loaded with alternate inductances (L_{L1} , L_{L2}) and capacitances (C_{L1} , C_{L2}) that may be realized using either lumped printed or discrete (chip) inductors and capacitors. Z_0 and k refer to the characteristic impedance and wave number of the TL used. Many cells are arranged periodically into a one-dimensional (1-D) array, and the array is excited at guide wavelengths longer than the cell dimension to enforce the homogeneity condition, so that the resulting periodic structure can be regarded as distributed.

A series of microwave applications such as dual-band branch-line coupler [78], asymmetric backward-wave directional coupler [76], zeroth order resonator [79] have been proposed.

A number of compact filters have been developed before, such as, using meander spurline and capacitively loaded stubs [5], sierpinski-based resonator [80], defected ground structure combined with lumped capacitor [81], multilayered TL [82], MCM-D technology [83], and superconductor [84]. Unfortunately, these kinds of compact filters have few disadvantages: the size of filters in [85, 80] are 23mm length resonating at 5.6 GHz and 40mm length resonating at 2.2 GHz respectively, the filters in [81, 82, 83] are difficult to fabricate because they are based on MCM-D technology or lumped capacitor. The superconductor[84] filter can only work at very low temperature and it is too expensive for these filters to be widely used.

2.4.2 SIW implementation of CRLH LWA

There has been a constant effort to develop better platforms for realizing planar microwave systems. Currently microstrip and coplanar waveguide technologies are the most popular choices for fabricating planar systems. However, it has been observed that with increase in frequency, these platforms pose problems of transmission and radiation losses. This calls for a need for different means of implementing planar technology. Recently, a new concept for implementing

microwave planar components known as Substrate Integrated Waveguide has been proposed. This technique integrates non planar 3D structures within a uniplanar circuit [86]. Ever since its introduction, substrate integrated waveguide has been widely developed for integrated microwave and mm wave components, such as post and cavity filters [87], directional couplers [88], circulators [89] and slot array and leaky antennas [90].

Basically, SIW are integrated waveguide like structures fabricated by using two rows of conducting cylinders or slots embedded in a dielectric substrate that electrically connects two parallel metal plates. This way of realization facilitates the implementation of nonplanar rectangular waveguide in a planar form. These conducting cylinders keep the fields confined within the structure [12, 87, 91].

Figure 2.10: SIW layout

Figure 2.10 shows a typical implementation of a basic substrate integrated waveguide. The dimensions d is the diameter of the metal vias, w represents their transverse spacing and s is their longitudinal spacing. These SIW structures have similar propagation characteristics as that of classical rectangular waveguides [92]. These were initially called post wall waveguides or laminated waveguides and were used for feeding networks in antenna arrays. But now, SIW technology has been applied to implement different microwave devices [87, 93]

As discussed, the main advantage of SIW which is beneficial to CRLH implementation is its ability to provide negative permittivity when operated below cutoff frequency. Currently, realizing CRLH on SIW is gaining lot of research interest. Different CRLH structures especially LWA are being tried out on SIW. CRLH structures have already been tried out on microstrip and coplanar waveguide. It is challenging to design lumped series capacitance, which is a necessity for CRLH structures. This is overcome by etching interdigital capacitance on the surface of SIW for large series capacitance.

Figure 2.11: The unit cell of the uniplanar CRLH structure in SIW [94].

Basic principle of implementing CRLH on SIW is by using series interdigital capacitors and via holes for shunt inductors. Tie Jun Cui et al. proposed a basic CRLH structure implemented on a SIW [94]. Figure 2.11 shows the CRLH structure on SIW. The values of the inductance and capacitance can be suitably controlled by adjusting the distance between two vias, width of the interdigital structure, number of fingers etc.

There have been many SIW slot antennas that have been developed for specific applications, all of which are operated in the RH region. Dong et al. proposed a new SIW slot antenna built on a negative zeroth order resonance [95]. They are operated far below initial waveguide cutoff frequency which also leads to miniaturization of the structure. They are distinguished as CRLH resonant antennas and they are realized by introducing a series capacitor into the circuit. This new capacitance along with the waveguide inherent shunt inductance provides the required loading, thus providing a CRLH response.

Few CRLH SIWs have been designed, such as the antenna with transverse slots etched on the waveguide surface which are bridged by metallic strips located in a plane slightly below the surface as shown in Figure 2.12 (a) [92], the antenna with interdigital capacitors etched on the waveguide surface as shown in Figure 2.12(b) [95]. Figure 2.12(c) shows the two SP-CRLH half mode substrate integrated waveguide (HMSIW) LWAs [95].

Their propagation properties and radiation characteristics have been studied extensively to demonstrate the backfire-to-endfire beam-steering capabilities through frequency scanning. The radiating structures are realized by etching periodically transverse slots on the upper side of the waveguide. The slot contributes the series capacitor as well as a radiator leading to a CRLH LW application.

The concept of miniaturization has been applied to SIW structures by reducing the size of the original structure to make it work in half mode. This configuration is called Half mode SIW abbreviated as HMSIW. Under this configuration, wave can propagate and radiate both below and above the cutoff frequency. It retains the half mode field distribution of the original structure. These antennas have the advantages of low profile, low cost, and low weight, while they also have unique features, such as high directivity, quasi-omnidirectional radiation, miniaturized size, continuous beam-steering capabilities covering both the backward and forward quadrants, etc., thus providing much needed design flexibility for the real applications [95].

Figure 2.12: Examples of the SIW implementations of SP-CRLH TLs. (a) A SP-CRLH SIW with transverse slots [92]. (b) SP-CRLH SIWs with interdigital slots. (c) SP-CRLH HMSIW with interdigital slots [95].

2.5 Singular perturbation method applied to periodic structures

In order to gain better physical insights into the transmission line structures, there is a necessity to obtain a method to determine the propagation constant (phase constant and attenuation constant) and radiation characteristics. The typical numerical methods available for this analysis include TL theory, the Floquet Theorem [5] and the equivalent circuit approach [96]. However, in these methods, the analysis used is static methods or quasistatic methods. The nature of the mode of propagation is considered to be a pure transverse electric-magnetic (TEM) mode and the TL characteristics are calculated from the electrostatic capacitance and inductance of the TLs, but the hybrid nature of the mode of the propagation should be taken into account which is more appropriate [13].

Especially for microstrip lines, they are not capable of supporting a pure TEM mode wave, and small longitudinal components of both the electric and magnetic fields need to be considered to satisfy the boundary conditions at the dielectric-air interface. These hybrid modes cannot be fully described in terms of static capacitances and inductances only. Since the power density distributions of the symmetric E component, which is very close and similar to the transverse magnetic (TM) mode as opposed to other higher order modes [75], the dispersion characteristics of the TM mode are investigated to determine the propagation constant.

The perturbation procedure provides a better physical insight of the analysis [97]. It starts with the exact solution of a related problem, and then it uses mathematical methods to find an approximate solution to a problem which cannot be solved exactly by full wave analysis. Perturbation method provides an expression for the desired solution in terms of a formal power series in a parameter that quantifies the deviation from the exactly solvable problem. This parameter can be expressed and expanded as a series which will be an approximation to the full solution. Singular perturbation method is one of the mathematical methods that is used

to find an approximate solution to a problem which cannot be solved exactly. It identifies a small parameter, such that when this parameter is set to zero, the problem (for example, the differential equation) becomes solvable. Singular perturbation method is applicable when there is a problem containing a small parameter that cannot be approximated by setting the parameter value to zero.

$$H_y = \delta^0 H_{y0} + \delta^1 H_{y1} + \delta^2 H_{y2} + \cdots + \delta^n H_{yn}, \quad (2.1)$$

where H_{y0} represents the known solution to the exactly solvable unperturbed initial problem and $H_{y1}, H_{y2}, \cdots, H_{yn}$ represent the higher-order terms which may be found iteratively by some systematic procedure [97].

It is noted that in addition to the dispersion characteristic derived from the zeroth-order problem, the perturbation method also gives information about the radiation efficiency and the radiation angle for the optimum radiation efficiency, which are derived from the first-order and second-order problem [98].

2.6 SIW resonator antennas

Substrate integrated waveguides (SIW) have been studied extensively in recent times [99]. SIW takes advantage of the well-known characteristics of conventional rectangular waveguides, such as high Q-factor and high power density. Several kinds of antennas [100, 101] and filters [102, 103, 104] based on SIW have been realized.

A cavity resonator is formed when the two open ends of a rectangular wave guide are closed by the metal plates (Figure 2.13). This introduction of the metal walls to the waveguide structure generates a pattern of standing waves in the cavity, resulting in the conventional high pass behaviour of the waveguide being transformed into very narrow bandpass behaviour. It is due

to the change of waveguide cut off frequency to the oscillatory frequency of the resonator [13]. In the resonator, the resonant modes exist in correspondence of discrete resonance frequencies. The typical cavity modes of a resonator are represented by TE_{mnp} and TM_{mnp} . The indices m , n and p correspond to periodicity (number of sine or cosine waves) in the three directions. If z -axis is set as reference direction, then the possible indices are

$$TE_{mnp} : m, n, p = 0, 1, 2, 3..$$

$$TM_{mnp} : m, n, p = 0, 1, 2, 3..$$

Figure 2.13: Typical cavity resonator

The dominant mode of the basic cavity resonator is TE_{101} . In case of TE mode only one m or n zero index is allowed, whereas for TM mode it is possible to have the third index as zero. The cavity resonator can be implemented on a planar platform in the form of substrate integrated waveguide (SIW). SIW has metallic posts which are equivalent to side metallic walls of a conventional waveguide (Figure 2.14). The field and frequency expressions for the conventional rectangular waveguide resonator hold good for the SIW resonator too.

Figure 2.14: SIW equivalent of cavity resonator

Half mode substrate integrated waveguide (HMSIW) was proposed and exploited to design some high performance microwave components in [105]. It is demonstrated that substrate integrated waveguide can be bisected along a fictitious quasi-magnetic wall and the SIW becomes an HMSIW. The HMSIW can almost preserve the field distribution of the original SIW and a leaky wave radiation can be obtained from the dielectric aperture of the structure [106]. Because the center symmetrical plane of the HMSIW can be equivalently regarded as a quasimagnetic wall for a few particular modes (Figure 2.15), the HMSIW can be further bisected into two parts again along the symmetrical plane.

These resonator antennas can be used for both linear and circularly polarized antenna designs. In [107] HMSIW technology is used to realize low-profile, single-fed CP cavity-backed antennas operating between 8.6 GHz and 8.7 GHz. A wearable antenna based on the aperture field distribution of a half-mode substrate-integrated resonant cavity is proposed which operates at 4.45 GHz [108]. Owing to its applicability in wearable technology, the design requires silver coated fabric to form the ground and the top planes. In [109] a HMSIW topology has been implemented in textile materials for a dual band on body application. It uses shorting pins on the top metal layer.

Figure 2.15: (a) Configuration of the square SIW resonator and (b) Configuration of the HMSIW [101].

After bisecting the SIW twice along the central symmetrical fictitious magnetic wall, a new structure termed as quarter-mode substrate integrated waveguide (QMSIW) is obtained. A quarter-mode substrate integrated waveguide (QMSIW) is realized as shown in Figure 2.16. The field distribution of the QMSIW is one fourth as the field distribution of the original SIW. Depending on the resonant mode analysis of the QMSIW structure, it is found that the QMSIW structure can be used for antenna designs. Radiation occurs in the discontinuity between the lateral edge and the extended dielectric-filled ground plane [101]. This structure provides an attractive and promising way to design resonator antennas.

Figure 2.16: Configuration of the QMSIW resonator [101].

It has been shown that the QMSIW structure almost preserves the electric field distribution of quarter sector of the original square SIW and half sector of the corresponding HMSIW. The initial results of the structure indicate that the TE_{101} mode of the QMSIW structure can be used to design linear polarization antennas [101]. This antenna has many advantages, such as compact structure, simple matching network, planar configuration and convenient integration with the planar circuit systems.

2.7 Summary

This chapter provides the background for metamaterials and their applications as antennas on different platforms. The evolution of metamaterials and their adaptation to the field of microwaves have been discussed. Non-resonant type of MTM was proposed by three research groups which were implemented as composite right/ left handed transmission lines. The appli-

cation of these TLs as leaky wave antennas has been of research interest. Different contributions in this area have been provided. It is seen that these structures have opened up new avenues of implementing periodic structures. They overcome the shortcomings of the existing designs, like for instance, CRLH TLs when applied as LWAs, are able to radiate in the broadside direction which the conventional LWAs are incapable of radiating in that direction. The implementations of these structures on microstrip and SIW platforms have been presented.

An analytical method called singular perturbation has been discussed, which provides advantages like obtaining dispersion, radiation characteristics by manually deriving the rigorous equations, thus providing a reference to compare the simulation results.

The last section focussed on the concepts of SIW resonator antenna and its subsequent miniaturization methods. Its adaptation to HMSIW and QMSIW have been discussed. Proceeding with the same concept we propose and implement an octa mode SIW resonator antenna at usable frequency bands.

Chapter 3

Singular perturbation method on a double periodic structure

This chapter presents singular perturbation analysis implemented for two parameter modulation. In order to analyse the radiation characteristics of the double periodic composite right/left handed (DP-CRLH) TL proposed in [15], a singular perturbation procedure has been used. The leaky-wave (LW) characteristics from the fast wave region in the DP-CRLH TLs is implemented by the microstrip loaded with double periodic interdigital capacitors and shunt short-circuited stub inductances. The analysis has been done with the help of the singular perturbation procedure. The new LW region was found below the left-handed (LH) passband for the DP-CRLH TLs based on this analysis method. The analysis shows that wave radiates forward and the TLs work as a right-handed material (RHM) in the new LW region. Dispersion characteristics and the radiation efficiency are investigated with the help of the singular perturbation procedure. The radiation efficiency is numerically estimated as a function of the antenna length, and the radiation angle is estimated in the LW region. Validating experimental results have been provided too.

3.1 Problem definition

A double periodic leaky wave structure based on CRLH microstrip was proposed in [15] with alternately loaded inductances and capacitances. However, the perturbation analysis was performed only for the modulation of single parameter namely ε . There was a scope to include the modulation of another parameter μ . In order to have a better understanding of the wave propagation and its field interaction as it travels through this DP-CRLH LWA, both its parameters need to be included in the analysis. The aim of this chapter is to perform the singular perturbation analysis on multiple space scales including both the ε and μ modulations and to compare the theoretically obtained results with the measured frequency responses and radiation pattern results.

3.2 Introduction

3.2.1 Overview of numerical methods

With the evolution of highly efficient simulation software, the necessity of rigorous manual mathematical analysis techniques have become obsolete. However, it is still desirable to have a numerical model to compare and verify the simulation results. The numerical methods provide a better insight into the electromagnetic phenomena occurring in the microwave structures. Most of the commercially available softwares are based on popular numerical techniques like transmission-line matrix method, finite difference time domain (FDTD) method, mode matching method, finite element method etc. These methods can get very complex when attempted to solve manually and have their own shortcomings. For instance, the transmission-line matrix method, however efficient in solving field problems, cannot explicitly provide dispersion results. Similar drawbacks exist with other methods as well. The mode matching method does not provide details on the radiation characteristics of the structure. These shortcomings can be

overcome by adopting a method which can be solved manually, be developed as a model which can be used for different types of structures by modifying a few design parameters. One such method is singular perturbation method (SPM) based on multiple space scales.

3.2.2 Applicability and Advantages of Singular perturbation method (SPM)

SPM provides radiation pattern and radiation efficiency of the structure under consideration. We need to input the inductor (L) and capacitor (C) values along with the permittivity and the permeability values. The initial round of substitution will yield results which are frequency dependent. The radiation angle also can be found out using the same data. For any structure whose permittivity and permeability are both modulated, this analysis holds good. The method has many advantages [110].

- 1) To provide a basis for the understanding of wave propagation in the structure.
- 2) To identify different types of wave interaction.
- 3) To give accurate numerical data as a benchmark result against which approximate ones can be judged.
- 4) It approaches the accuracy obtainable from the rigorous treatments while retaining the simplicity of perturbation methods.
- 5) It yields a description of the EM fields in terms of a transverse transmission line network which brings considerable physical insight into the overall behaviour of the periodic structure.
- 6) It provides a better physical insight into the structure.
- 7) Theoretically calculated radiation efficiency and radiation pattern can be obtained, which can be used to verify the experimentally observed results.

- 8) Once derived, the same model can be used for different structures with appropriate input parameters.

3.2.3 SPM overview

The singular perturbation procedure provides a better physical insight of the field analysis. It starts with the exact solution of a related problem, and then it uses mathematical methods to find an approximate solution to a problem which cannot be solved exactly by full wave analysis. Perturbation method provides an expression for the desired solution in terms of a formal power series in a parameter that quantifies the deviation from the exactly solvable problem. This parameter can be expressed and expanded as a series which will be an approximation to the full solution.

Singular perturbation method is one of the mathematical methods that is used to find an approximate solution to a problem which cannot be solved exactly. It identifies a small parameter, such that when this parameter is set to zero, the problem (for example, the differential equation) becomes solvable. The method is applicable when there is a problem containing a small parameter that cannot be approximated by setting the parameter value to zero.

$$H_y = \delta^0 H_{y0} + \delta^1 H_{y1} + \delta^2 H_{y2} + \cdots + \delta^n H_{yn}, \quad (3.1)$$

where H_{y0} represents the known solution to the exactly solvable initial problem and H_{y1} , H_{y2} , $\cdots$, H_{yn} represent the higher-order terms which may be found iteratively by some systematic procedure [97]. The flow chart of the singular perturbation procedure for the analysis of the periodic structure is shown in Figure 3.1. It is noted that in addition to the dispersion characteristics derived from the zeroth-order problem, the perturbation method also gives information about the radiation efficiency and the radiation angle for the optimum radiation efficiency, which are derived from the first-order and second-order problems [98].

Figure 3.1: Flowchart depicting the singular perturbation method

3.2.4 SPM as applied to periodic structures

The wave propagation in periodic structures has been a subject of continuing interest in the literature. Periodic structures have been studied, well understood and have been utilized in many optical as well as microwave applications. Recently, double periodic structures, which form a special category of almost-periodic structures, have been presented in the literature [15]. Interesting phenomena like leaky wave, CRLH behaviour has been conjectured and has been verified by experimental results. Their potential applications to device design have been suggested in [111].

While novel simulation methods are desirable for obtaining analytical results that are useful for practical design considerations, they often require a priori knowledge of basic physical processes involved. Hence a clear physical picture of wave process is needed for the understanding of the wave phenomena in periodic structures and also for developing a theoretical basis on which the desired simulated approximation can be established.

We propose to apply the singular perturbation analysis to the periodic structure shown in Figure 3.2. This is a double periodic transmission line with leaky wave radiation characteristics. The details of the structure are provided in subsequent sections.

Figure 3.2: Layout of the double periodic TL leaky wave structure

3.3 Initial analysis

The steps involved in solving the perturbation problem have been summarized in the flowchart shown in Figure 3.3. This section discusses the initial procedures to obtain the problem statement.

Figure 3.3: Flowchart depicting the analysis procedure of the singular perturbation method

3.3.1 Choice of propagation mode

The double periodic structures analysed by TL method or Floquet theorem have used either static methods or quasi-static methods. The mode of propagation in these methods is considered to be a pure transverse electric-magnetic (TEM) mode. The results are evaluated from the electrostatic capacitance and inductance of the TLs. However, for all practical purposes the mode of propagation is hybrid in nature and hence pure TEM mode assumption is not very accurate.

Especially, microstrip lines are not capable of supporting a pure TEM mode wave [41]. The longitudinal components of both the electric and the magnetic fields require to be taken into account for satisfying the boundary conditions at the dielectric-air interface. These hybrid modes cannot be fully described in terms of static capacitances and inductances. For the microstrip line, the fields in x-direction do not vary significantly owing to thin homogenous isotropic medium. Also, the power density distributions of the symmetric E-component are almost similar to the transverse magnetic (TM) mode [110]. Keeping these factors in mind, we consider the TM mode to determine the propagation characteristics.

The DP-CRLH TLs can be constructed based on the geometry shown in Figure 3.2. It is assumed that the wave is propagating along the z-direction, and has no variation in the y-direction ($\frac{\partial}{\partial y} = 0$). In a thin homogeneous isotropic medium, such as microstrip lines, a TM mode wave is considered because of a weak x-variation [10]. We consider the TM mode with non-vanishing field components H_y , E_x and E_z , independent of y-direction. The substrate material is assumed to be lossless.

3.3.2 Equivalent circuit parameters

The composite right/left handed transmission line structures have a direct L-C equivalent circuit. Both the single periodic and the double periodic CRLH TLs can be represented in terms of their circuit parameters. We know that the shunt capacitance and series inductance contribute to the right handed behaviour and the shunt inductance and series capacitance result in the left handed nature of the structure. This combination is responsible for the display of simultaneous negative equivalent permittivity and negative equivalent permeability. The equivalent permittivity(ϵ) and permeability (μ) are related to the capacitance and inductance as $C = \epsilon d$ and $L = \mu d$. The dimension d is the period of the unit cell with $d=10\text{mm}$. The cell dimension is smaller in comparison to the guided wavelength(λ_g). The equivalent circuit is as shown in the Figure 3.4.

Figure 3.4: Equivalent circuit of double periodic TL

It has been experimentally proven in [15] that the TLs double periodically loaded with either alternate capacitances or alternate inductances will contribute the new LW behaviour. Transmission lines can either be only with double periodically loaded alternate capacitances or only with double periodically loaded alternate inductances or both. The singular perturbation analysis for just double periodically loaded alternate inductances has been provided. In our case we consider transmission line which is double periodically loaded with alternate inductance and capacitance. Figure 3.5. shows the unit cell of the proposed structure for which the singular perturbation analysis will be extended to modulation of two parameters. The substrate details of the structure are $\epsilon_r=4.9$ with thickness=1.6 mm. $l_1=4.35$, $l_2=3.7$, $l_3^*=4.5$, $l_3^\# =7.9$, $l_4=1$, $l_5=1, l_6=0.5$, $S_1=0.15$, $S_2=0.15$, $w=0.2$, $d=5$. All units are in mm.

Figure 3.5: Equivalent circuit of double periodic TL

For the TL with double periodically loaded alternate inductances and alternate capacitances, the expansion is given by Eqn 3.2.

$$L = L_0(1 + \delta_L \eta_{L1} \cos(Kz_0 + \theta_1) + \delta_L^2 \eta_{L2} \cos(2Kz_0 + \theta_2) + O(\delta_L^3)) \quad (3.2a)$$

$$C = C_0(1 + \delta_C \eta_{C1} \cos(Kz_0 + \theta_1) + \delta_C^2 \eta_{C2} \cos(2Kz_0 + \theta_2) + O(\delta_C^3)) \quad (3.2b)$$

Where L_0 and C_0 are average inductance and capacitance values respectively, η_{L1} and η_{C1} are the amplitude of the fundamental (first harmonic) component of the periodically loaded inductance while η_{L2} and η_{C2} are the amplitudes of the second harmonic, δ is a perturbation coefficient smaller than 1, and grating vector K is satisfied by the Bragg condition $K = 2\pi/d$. θ_1 is the phase of the first harmonic component and θ_2 is the phase of the second harmonic.

3.3.3 Expansions for ε and μ

The capacitances and the inductances of the structure affect the dielectric and magnetic properties of the medium directly. The series capacitance (C_L) and shunt inductance (L_L) contribute to the left handed behaviour, while the series inductance (L_R) and shunt capacitance (C_R) result in negative dispersive characteristics. Negative dispersion is characterized by negative refractive index(NRI). For a frequency dispersive medium the equivalent permittivity and permeability is given by [41]

$$\begin{aligned} \varepsilon(\omega) &= \varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_L d}, \\ \mu(\omega) &= \mu_0 - \frac{1}{\omega^2 C_L d} \end{aligned} \quad (3.3)$$

where L_L and C_L represent the double periodically loaded inductance and capacitance respectively. ε_{eff} is the effective permittivity of the unperturbed host medium (Fig 3.6).

Figure 3.6: Model of the DP TL showing the dielectric variations

For instance, ϵ_{eff} of the microstrip lines is

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \sqrt{\frac{1}{1 + 12s/W}}, \quad (3.4)$$

where s/W is the ratio of substrate thickness to the conductor width of the microstrip line and ϵ_r is the relative permittivity of the substrate. The Bloch impedance is given by $Z_B = g\sqrt{\mu/\epsilon}$, where g is the geometric factor, and it can be determined using the width and substrate height of the microstrip line. Previous works have focussed only on the variation of permittivity ϵ . In this work, the analysis has been carried out for the variation of both permittivity ϵ and permeability μ .

In order to include the variation of ϵ in the analysis, we obtain a form which is suitable for substitution. We consider the TM mode with nonvanishing field components H_y , E_x and E_z having no variation in the y -direction. By substituting (3.2) into (3.3), a periodically modulated permittivity profile in z -direction can be expressed as

$$\begin{cases} \varepsilon = \varepsilon(z) = \varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_0 d} + \frac{\delta \eta_{L1} \cos(Kz_0 + \theta_1)}{\omega^2 L_0 d} + \frac{\delta^2 \eta_{L2} \cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d} + \dots, \\ \frac{\partial \varepsilon}{\partial z} = \frac{1}{\omega^2 L_0 d} (-\delta \eta_{L1} K \sin(Kz_0 + \theta_1) - 2\delta^2 \eta_{L2} K \sin(2Kz_0 + \theta_2) - \dots), \\ \mu = \mu(z) = \mu_0 - \frac{1}{\omega^2 C_0 d} + \frac{\delta \eta_{C1} \cos(Kz_0 + \theta_1)}{\omega^2 C_0 d} + \frac{\delta^2 \eta_{C2} \cos(2Kz_0 + \theta_2)}{\omega^2 C_0 d} + \dots, \end{cases} \quad (3.5)$$

3.3.4 Derivation of Helmholtz equation for TM mode

The y-component of the magnetic field H_y satisfies the differential Maxwell's equations. The Maxwell's equations are given as (Eqn 3.6)

$$\begin{cases} \nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t} \\ \nabla \times \vec{H} = \vec{J} + \frac{\partial \vec{D}}{\partial t} \\ \nabla \cdot \vec{D} = \rho \\ \nabla \cdot \vec{B} = 0 \end{cases} \quad (3.6)$$

where $\vec{E}$ is the electric field intensity, $\vec{H}$ is the magnetic field intensity, $\vec{B} = \mu \vec{H}$ is the magnetic flux density, where $\mu = \mu_r \mu_0$ is the permeability of the medium, μ_r is the relative permeability of the medium and μ_0 is permeability of freespace. $\vec{J}$ is the electric current density, and $\vec{J} = 0$ in source-free regions of space. $\vec{D} = \varepsilon \vec{E}$ is the electric flux density, where $\varepsilon = \varepsilon_r \varepsilon_0$ is the permittivity of the medium, ε_r is the relative permittivity of the medium and ε_0 is the permittivity of free-space. ρ is the electric charge density, and $\rho = 0$ in source-free regions of space.

Starting from these equations, we perform a few vector operations to obtain suitable form to make the substitutions. Applying curl operator on the $\nabla \times \vec{H}$ and substituting for $\nabla \times \vec{E}$

gives

$$\nabla \times \nabla \times \vec{H} = \nabla \times \left(\frac{\partial \vec{D}}{\partial t} \right) = \nabla \times \left(\frac{\partial(\varepsilon \vec{E})}{\partial t} \right) = \frac{\partial}{\partial t} \nabla \times (\varepsilon \vec{E}) \quad (3.7)$$

The field components of the TM mode are given as $\vec{H} = \vec{e}_y H_y$ and $\vec{E} = \vec{e}_x E_x + \vec{e}_z E_z$. So we have

$$\frac{\partial}{\partial t} \nabla \times (\varepsilon \vec{E}) = \frac{\partial}{\partial t} \nabla \times (\varepsilon (\vec{e}_x E_x + \vec{e}_z E_z)) \quad (3.8)$$

Expanding the RHS we have

$$\begin{aligned} \frac{\partial}{\partial t} \nabla \times (\varepsilon (\vec{e}_x E_x + \vec{e}_z E_z)) &= \frac{\partial}{\partial t} \left[\vec{e}_x \frac{\partial(\varepsilon E_z)}{\partial y} + \vec{e}_y \left(\frac{\partial(\varepsilon E_x)}{\partial z} - \frac{\partial(\varepsilon E_z)}{\partial x} \right) + \vec{e}_z \left(-\frac{\partial(\varepsilon E_x)}{\partial y} \right) \right] \\ &= \frac{\partial}{\partial t} \left\{ \varepsilon \left[\vec{e}_x \frac{\partial E_z}{\partial y} + \vec{e}_y \left(\frac{\partial E_x}{\partial z} - \frac{\partial E_z}{\partial x} \right) + \vec{e}_z \left(-\frac{\partial E_x}{\partial y} \right) \right] + \vec{e}_y E_x \frac{\partial \varepsilon}{\partial z} \right\} \\ &= \frac{\partial}{\partial t} \left[\varepsilon \nabla \times \vec{E} + \vec{e}_y E_x \frac{\partial \varepsilon}{\partial z} \right] \\ &= \frac{\partial}{\partial t} \varepsilon \nabla \times \vec{E} + \vec{e}_y \frac{\partial}{\partial t} \left(E_x \frac{\partial \varepsilon}{\partial z} \right) \end{aligned} \quad (3.9)$$

If we substitute $\nabla \times \vec{E} = -\mu \frac{\partial \vec{H}}{\partial t}$ and $E_x = -\frac{1}{j\omega\varepsilon} \frac{\partial H_y}{\partial z}$ into the last equation, we get

$$\begin{aligned} \frac{\partial}{\partial t} \varepsilon \nabla \times \vec{E} + \vec{e}_y \frac{\partial}{\partial t} \left(E_x \frac{\partial \varepsilon}{\partial z} \right) &= \frac{\partial}{\partial t} \varepsilon \left(-\frac{\partial \mu \vec{H}}{\partial t} \right) + \vec{e}_y \frac{\partial}{\partial t} \left[\frac{\partial \varepsilon}{\partial z} \left(-\frac{1}{j\omega\varepsilon} \frac{\partial H_y}{\partial z} \right) \right] \\ &= -\varepsilon \frac{\partial^2 \mu \vec{H}}{\partial t^2} - \vec{e}_y \left[\frac{\partial \varepsilon}{\partial z} \left(\frac{1}{j\omega} \frac{\partial H_y}{\partial z} \right) \right]. \end{aligned} \quad (3.10)$$

So we have

$$\nabla \times \nabla \times \vec{H} = -\varepsilon \frac{\partial^2 \mu \vec{H}}{\partial t^2} - \vec{e}_y \left[\frac{\partial \varepsilon}{\partial z} \left(\frac{1}{j\omega} \frac{\partial H_y}{\partial z} \right) \right]. \quad (3.11)$$

We can write this equation in phasor form by replacing the time derivatives by $j\omega$, and

$$\nabla \times \nabla \times \vec{H} = \omega^2 \mu \varepsilon \vec{H} - \vec{e}_y \frac{\partial \varepsilon}{\partial z} \frac{\partial H_y}{\partial z}. \quad (3.12)$$

Thus the Helmholtz equations are as shown

$$\nabla^2 \vec{H} = \nabla(\nabla \times \vec{H}) - \nabla \times \nabla \times \vec{H} = -\nabla \times \nabla \times \vec{H}. \quad (3.13)$$

The magnetic field equation in y -direction (H_y) of the TM mode satisfies the Helmholtz equation as shown below.

$$\left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + \omega^2 \mu \varepsilon \right\} H_y = \frac{1}{\varepsilon} \frac{\partial \varepsilon}{\partial z} \frac{\partial H_y}{\partial z} \quad (3.14)$$

and

$$E_z = \frac{1}{j\omega\varepsilon} \frac{\partial H_y}{\partial x}, \quad E_x = -\frac{1}{j\omega\varepsilon} \frac{\partial H_y}{\partial z}. \quad (3.15)$$

3.3.5 Perturbation expressions and assumptions

Asymptotic expressions can be extracted from the periodic structures which are equivalent to corresponding perturbation parameters. The typical parameters include permittivity (ε), permeability (μ), the electric field (E_x and E_z) and the magnetic field (H_y). A composite expansion is obtained by expanding these parameters in different orders. The structure varies periodically in z direction and has both its permittivity and permeability modulated as given

by the below equations.

$$\begin{aligned}
 L_L(z) &= \delta_L^0 L_0(z_0, z_1, z_2, \dots) + \delta_L^1 L_1(z_0, z_1, z_2, \dots) + \delta_L^2 L_2(z_0, z_1, z_2, \dots) + \dots \\
 &= L_0[1 + \delta_L \eta_{L1} \cos(Kz + \theta_1) + \delta_L^2 \eta_{L2} \cos(2Kz + \theta_2) + \dots] \\
 C_L(z) &= \delta_C^0 C_0(z_0, z_1, z_2, \dots) + \delta_C^1 C_1(z_0, z_1, z_2, \dots) + \delta_C^2 C_2(z_0, z_1, z_2, \dots) + \dots \\
 &= C_0[1 + \delta_C \eta_{C1} \cos(Kz + \theta_1) + \delta_C^2 \eta_{C2} \cos(2Kz + \theta_2) + \dots] \\
 \varepsilon(z) &= \delta_L^0 \varepsilon_0(z_0, z_1, z_2, \dots) + \delta_L^1 \varepsilon_1(z_0, z_1, z_2, \dots) + \delta_L^2 \varepsilon_2(z_0, z_1, z_2, \dots) + \dots \\
 &= \varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_0 d} + \frac{\delta_L \eta_{L1} \cos(Kz + \theta_1)}{\omega^2 L_0 d} + \frac{\delta_L^2 \eta_{L2} \cos(2Kz + \theta_2)}{\omega^2 L_0 d} + \dots \\
 \mu(z) &= \delta_C^0 \mu_0(z_0, z_1, z_2, \dots) + \delta_C^1 \mu_1(z_0, z_1, z_2, \dots) + \delta_C^2 \mu_2(z_0, z_1, z_2, \dots) + \dots \\
 &= \mu_0 - \frac{1}{\omega^2 C_0 d} + \frac{\delta_C \eta_{C1} \cos(Kz + \theta_1)}{\omega^2 C_0 d} + \frac{\delta_C^2 \eta_{C2} \cos(2Kz + \theta_2)}{\omega^2 C_0 d} + \dots
 \end{aligned} \tag{3.16}$$

It can be seen that the variation of μ has been included in the analysis for the proposed periodic structure. The y component of the magnetic field H_y can be expanded as shown in Eqn 3.17

$$\begin{aligned}
 H_y(x, z) &= \delta^0 H_{y0}(x, z_0, z_1, z_2, \dots) + \delta^1 H_{y1}(x, z_0, z_1, z_2, \dots) \\
 &\quad + \delta^2 H_{y2}(x, z_0, z_1, z_2, \dots) + \dots
 \end{aligned} \tag{3.17}$$

Space scales are introduced in the analysis as shown.

$$z_0 = z, \quad z_1 = \delta z \quad z_2 = \delta^2 z, \dots, z_n = \delta^n z. \tag{3.18}$$

From the analysis performed in [98], it is established that the procedure can be carried out until the second order to obtain the desired results. The derivation gets more complex as the order increases. Also, the necessary information like dispersion characteristics, radiation efficiency and radiation angle, can be calculated just from the second order. Hence higher order

modes are not taken into consideration for the analysis. The space scales are assumed only until the second order.

$$z_0 = z, \quad z_1 = \delta z \quad z_2 = \delta^2 z. \quad (3.19)$$

The space derivative transform is given as

$$\frac{\partial}{\partial z} = \frac{\partial z_0}{\partial z} \frac{\partial}{\partial z_0} + \frac{\partial z_1}{\partial z} \frac{\partial}{\partial z_1} + \frac{\partial z_2}{\partial z} \frac{\partial}{\partial z_2} = \frac{\partial}{\partial z_0} + \delta \frac{\partial}{\partial z_1} + \delta^2 \frac{\partial}{\partial z_2}. \quad (3.20)$$

Since it is assumed that $\partial/\partial z_1 \ll \partial/\partial z_2$, the partial derivative term of z_1 becomes zero as $\partial/\partial z_1 \simeq 0$ [98]

$$\begin{cases} \frac{\partial}{\partial z} = \frac{\partial}{\partial z_0} + \delta^2 \frac{\partial}{\partial z_2}, \\ \frac{\partial^2}{\partial z^2} = \frac{\partial^2}{\partial z_0^2} + 2\delta^2 \frac{\partial^2}{\partial z_0 \partial z_2}, \end{cases} \quad (3.21)$$

and the expression for H_y is given by

$$\begin{cases} H_y(x, z) = H_{y0}(x, z_0, z_2) + \delta H_{y1}(x, z_0, z_2) + \delta^2 H_{y2}(x, z_0, z_2), \\ \frac{\partial H_y}{\partial z} = \frac{\partial H_{y0}}{\partial z_0} + \delta \frac{\partial H_{y1}}{\partial z_0} + \delta^2 \left(\frac{\partial H_{y0}}{\partial z_2} + \frac{\partial H_{y2}}{\partial z_0} \right). \end{cases} \quad (3.22)$$

3.4 Statement of the problem

In the previous section, the perturbation expressions were provided for permittivity (ε), permeability (μ), the electric field (E_x and E_z) and the magnetic field (H_y). We now substitute the obtained ε and μ equations in the Helmholtz equations as derived in 3.15. We assume $\delta_L = \delta_C = \delta$.

Substituting 3.16 and (3.22) into (3.15), the differential equation is obtained as

$$\begin{aligned}
 & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z_0^2} + 2\delta^2 \frac{\partial^2}{\partial z_0 \partial z_2} + \omega^2 [\mu_0 - \frac{1}{\omega^2 C_0 d} + \frac{1}{\omega^2 C_0 d} [\delta_c \eta_{C1} \cos(Kz_0 + \theta_1) + \delta_c^2 \eta_{C2} \cos(Kz_0 + \theta_2)]] \right. \\
 & \quad \left. [\varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_0 d}] + \delta \eta_1 \frac{\cos(Kz_0 + \theta_1)}{L_0 d} \right. \\
 & \quad \left. + \delta^2 \eta_2 \frac{\cos(2Kz_0 + \theta_2)}{L_0 d} \right\} [H_{y0} + \delta H_{y1} + \delta^2 H_{y2}] \\
 & = \left[\frac{1}{\varepsilon} \left(\frac{\partial \varepsilon}{\partial z} \right) \right] \left[\frac{\partial H_{y0}}{\partial z_0} + \delta \frac{\partial H_{y1}}{\partial z_0} + \delta^2 \left(\frac{\partial H_{y0}}{\partial z_2} + \frac{\partial H_{y2}}{\partial z_0} \right) \right],
 \end{aligned} \tag{3.23}$$

We can simplify the term $\frac{1}{\varepsilon} \left(\frac{\partial \varepsilon}{\partial z} \right)$ in Eqn 3.25 as shown.

$$\begin{aligned}
 \frac{1}{\varepsilon} \left(\frac{\partial \varepsilon}{\partial z} \right) &= - \frac{\delta \eta_{L1} \frac{K \sin(Kz_0 + \theta_1)}{\omega^2 L_0 d} + \delta^2 \eta_2 \frac{2K \sin(2Kz_0 + \theta_2)}{\omega^2 L_0 d}}{(\varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_0 d}) + \delta \eta_{L1} \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d} + \delta^2 \eta_{L2} \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d}} \\
 &= \frac{\delta \eta_{L1} K \sin(Kz_0 + \theta_1) + \delta^2 \eta_{L2} 2K \sin(2Kz_0 + \theta_2)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1) [1 + \delta \eta_{L1} \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1} + \delta^2 \eta_{L2} \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1}]} \\
 &\approx \frac{\delta \eta_{L1} K \sin(Kz_0 + \theta_1) + \delta^2 \eta_{L2} 2K \sin(2Kz_0 + \theta_2)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1)} \\
 &\quad \left[1 - \delta \eta_{L1} \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1} - \delta^2 \eta_{L2} \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1} \right].
 \end{aligned} \tag{3.24}$$

Substituting back the simplified form(Eqn 3.24) back in Eqn 3.25, we get

$$\begin{aligned}
 & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z_0^2} + 2\delta^2 \frac{\partial^2}{\partial z_0 \partial z_2} + \omega^2 \left[\mu_0 - \frac{1}{\omega^2 C_0 d} + \frac{1}{\omega^2 C_0 d} [\delta_c \eta_{C1} \cos(Kz_0 + \theta_1) \right. \right. \\
 & \quad \left. \left. + \delta_c^2 \eta_{C2} \cos(Kz_0 + \theta_2)] \right] \right. \\
 & \quad \left. \left[\varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_0 d} \right] + \delta \eta_1 \frac{\cos(Kz_0 + \theta_1)}{L_0 d} \right. \\
 & \quad \left. + \delta^2 \eta_2 \frac{\cos(2Kz_0 + \theta_2)}{L_0 d} \right\} [H_{y0} + \delta H_{y1} + \delta^2 H_{y2}] \\
 & = \frac{\delta \eta_{L1} K \sin(Kz_0 + \theta_1) + \delta^2 \eta_{L2} 2K \sin(2Kz_0 + \theta_2)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1)} \\
 & \quad \left[1 - \delta \eta_{L1} \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1} - \delta^2 \eta_{L2} \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1} \right] \\
 & \quad \left[\frac{\partial H_{y0}}{\partial z_0} + \delta \frac{\partial H_{y1}}{\partial z_0} + \delta^2 \left(\frac{\partial H_{y0}}{\partial z_2} + \frac{\partial H_{y2}}{\partial z_0} \right) \right]
 \end{aligned} \tag{3.25}$$

This is the final form of the expression which will be used to separate out the δ terms based on its order. Depending on the order of δ , the coefficients can be grouped to obtain problem statement for different orders.

3.4.1 Zeroth order problem

Grouping the δ^0 terms from eqn 3.25, we obtain the zeroth order of the perturbation problem.

$$\begin{aligned}
 \varepsilon_{s0} &= \varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_0 d} = \frac{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1}{\omega^2 L_0 d} \\
 \mu_{s0} &= \mu_0 - \frac{1}{\omega^2 C_0 d} = \frac{\omega^2 C_0 d \mu_0 - 1}{\omega^2 C_0 d}
 \end{aligned} \tag{3.26}$$

If we substitute 3.26 in 3.25, we get

$$\begin{aligned}
 & \text{Zeroth order } O(\delta^0) : \\
 & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z_0^2} + \omega^2 \mu_{s0} \varepsilon_{s0} \right\} H_{y0} = 0,
 \end{aligned} \tag{3.27}$$

The solution of zeroth order will provide the dispersion characteristics of the structure.

We need to apply the boundary conditions to Eqn 3.27 in order to obtain the solutions. The boundary conditions for the zeroth order are given in Eqn 3.28 where subscripts s and a represent substrate and air respectively.

Boundary conditions – Zeroth order $O(\delta^0)$:

$$\begin{cases} \frac{\partial H_{y0s}}{\partial x} = 0 & x = 0 \\ H_{y0s} = H_{y0a} \Rightarrow \frac{1}{\varepsilon_{s0}} \frac{\partial}{\partial x} H_{y0s} = \frac{1}{\varepsilon_0} \frac{\partial}{\partial x} H_{y0a} & x = s \end{cases} \quad (3.28)$$

3.4.2 First order and second order problem

If we reorganize the LHS of Eqn 3.25 and rearrange the δ^1 and δ^2 terms, we get equation in the form shown in Eqn 3.29

$$\begin{aligned} & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z_0^2} + 2\delta^2 \frac{\partial^2}{\partial z_0 \partial z_2} + \omega^2 [(\mu_{s0}\varepsilon_{s0}) + \delta\mu_{s0}(\frac{\eta_{L1}\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d}) \right. \\ & \quad + \delta\varepsilon_{s0}(\frac{\eta_{C1}\cos(Kz_0 + \theta_1)}{\omega^2 C_0 d}) + \delta^2\mu_{s0}(\frac{\eta_{L2}\cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d}) \\ & \quad \left. + \delta^2\varepsilon_{s0}(\frac{\eta_{C2}\cos(2Kz_0 + \theta_2)}{\omega^2 C_0 d}) \right. \\ & \quad \left. + \delta^2(\frac{\eta_{C1}\cos(Kz_0 + \theta_1)}{\omega^2 C_0 d})(\frac{\eta_{L1}\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d}) \right\} [H_{y0} + \delta H_{y1} + \delta^2 H_{y2}] \\ & = \frac{\delta\eta_{L1}K\sin(Kz_0 + \theta_1) + \delta^2\eta_{L2}2K\sin(2Kz_0 + \theta_2)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1)} \\ & \quad [1 - \delta\eta_{L1} \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1} - \delta^2\eta_{L2} \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1}] \\ & \quad [\frac{\partial H_{y0}}{\partial z_0} + \delta \frac{\partial H_{y1}}{\partial z_0} + \delta^2 (\frac{\partial H_{y0}}{\partial z_2} + \frac{\partial H_{y2}}{\partial z_0})] \end{aligned} \quad (3.29)$$

Based on the above equation, we define the first order problem as shown in Eqn 3.30

First order $O(\delta^1)$:

$$\begin{aligned}
 & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z_0^2} + \omega^2 \mu_{s0} \varepsilon_{s0} \right\} H_{y1} + \left(\frac{\mu_{s0} \eta_{L1} \cos(Kz_0 + \theta_1)}{L_0 d} \right) H_{y0} \\
 & + \varepsilon_{s0} \left(\frac{\eta_{C1} \cos(Kz_0 + \theta_1)}{C_0 d} \right) H_{y0} \\
 & = - \frac{\eta_{L1} K \sin(Kz_0 + \theta_1)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1)} \frac{\partial H_{y0}}{\partial z_0},
 \end{aligned} \tag{3.30}$$

The boundary conditions for the first order problem is given as

Boundary conditions – First order $O(\delta^1)$:

$$\left\{ \begin{array}{ll} \frac{\partial H_{y1s}}{\partial x} = 0 & (x = 0) \\ H_{y1s} = H_{y1a} \Rightarrow & \\ \frac{1}{\varepsilon_{s0}} \frac{\partial}{\partial x} H_{y1s} - \frac{\omega^2 L_0 d \eta_{L1} \cos(Kz_0 + \theta_1)}{(2\varepsilon_0 \varepsilon_{eff} \omega^2 L_0 d - 1)^2} \frac{\partial}{\partial x} H_{y0s} & \\ = \frac{1}{\varepsilon_0} \frac{\partial}{\partial x} H_{y1a} & (x = s) \end{array} \right. \tag{3.31}$$

By grouping the δ^2 terms of Eqn 3.29 we get the second order problem which is defined by Eqn 3.32

Second order $O(\delta^2)$:

$$\begin{aligned}
 & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z_0^2} + \omega^2 \mu \varepsilon_{s0} \right\} H_{y2} \\
 &= -2 \frac{\partial^2}{\partial z_0 \partial z_2} H_{y0} + \left(\frac{\mu_{s0} \eta_{L2} \cos(2Kz_0 + \theta_2)}{L_0 d} \right) H_{y0} - \left(\frac{\varepsilon_{s0} \eta_{C2} \cos(2Kz_0 + \theta_2)}{C_0 d} \right) H_{y0} \\
 & \quad - \frac{\eta_{L1} \eta_{C1} K \cos(Kz_0 + \theta_1) \cos(Kz_0 + \theta_1)}{\omega^2 L_0 C_0 d^2} H_{y0} \\
 & \quad - \frac{\eta_{L1} \eta_{C1} K \cos(Kz_0 + \theta_1) \sin(Kz_0 + \theta_1)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff})^2} \frac{\partial H_{y0}}{\partial z_0} \\
 & \quad - \frac{2\eta_{L2} K \sin(2Kz_0 + \theta_2)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1)} \frac{\partial H_{y0}}{\partial z_0} + \left(\frac{\mu_{s0} \eta_{L1} \cos(2Kz_0 + \theta_1)}{L_0 d} \right) H_{y1} \\
 & \quad - \left(\frac{\varepsilon_{s0} \eta_{C1} \cos(Kz_0 + \theta_1)}{C_0 d} \right) H_{y1} - \left(\frac{\eta_{L1} K \sin(Kz_0 + \theta_1)}{(\omega^2 L_0 d \varepsilon_0 \varepsilon_{eff} - 1)} \right) H_{y1}
 \end{aligned} \tag{3.32}$$

The boundary conditions for the second order equations are given as(Eqn 3.33)

Boundary Conditions – Second order $O(\delta^2)$:

$$\left\{ \begin{array}{l} \frac{\partial H_{y2s}}{\partial x} = 0 \\ H_{y2s} = H_{y2a} \Rightarrow \\ \frac{1}{\varepsilon_{s0}} \frac{\partial}{\partial x} H_{y2s} - \frac{\omega^2 L_0 d \eta_{L1} \cos(Kz_0 + \theta_1)}{(2\varepsilon_0 \varepsilon_{eff} \omega^2 L_0 d - 1)^2} \frac{\partial}{\partial x} H_{y1s} = \frac{1}{\varepsilon_0} \frac{\partial}{\partial x} H_{y2a} \end{array} \right. \quad (x = 0), \quad (x = s), \tag{3.33}$$

For each order of δ^n , the boundary conditions are obtained when the fields H_y and E_z are continuous at $x=s$ and the field $E_z=0$ at $x=0$.

Substituting (Eqn 3.5) and (3.22) into the electric fields as shown in (3.14), the electric field E_z is derived as

$$\begin{aligned}
 E_z &= \frac{1}{j\omega \varepsilon} \frac{\partial H_y}{\partial x} \\
 &= \frac{1}{j\omega} \frac{1}{\varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_{L0} \Lambda_D} + \frac{\delta \eta_1 \cos(Kz_0 + \theta_1)}{\omega^2 L_{L0} \Lambda_D} + \frac{\delta^2 \eta_2 \cos(2Kz_0 + \theta_2)}{\omega^2 L_{L0} \Lambda_D}} \frac{\partial}{\partial x} (H_{y0} + \delta H_{y1} + \delta^2 H_{y2}).
 \end{aligned} \tag{3.34}$$

Simplifying it further to obtain final expression for E_z

$$\begin{aligned}
 & \frac{1}{\varepsilon_0 \varepsilon_{eff} - \frac{1}{\omega^2 L_{L0} d} + \frac{\delta \eta_1 \cos(Kz_0 + \theta_1)}{\omega^2 L_{L0} d} + \frac{\delta^2 \eta_2 \cos(2Kz_0 + \theta_2)}{\omega^2 L_{L0} d}} \\
 &= \frac{1}{\varepsilon_0 \varepsilon_{eff} - \omega^2 L_{L0} d} \frac{1}{1 + \delta \eta_1 \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_{L0} d \varepsilon_0 \varepsilon_{eff} - 1} + \delta^2 \eta_2 \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_{L0} d \varepsilon_0 \varepsilon_{eff} - 1}} \\
 &\approx \frac{1}{\varepsilon_0 \varepsilon_{eff} - \omega^2 L_{L0} d} \left[1 - \delta \eta_1 \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_{L0} d \varepsilon_0 \varepsilon_{eff} - 1} - \delta^2 \eta_2 \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_{L0} d \varepsilon_0 \varepsilon_{eff} - 1} \right],
 \end{aligned} \tag{3.35}$$

and then,

$$\begin{aligned}
 E_z = \frac{1}{j\omega} \frac{1}{\varepsilon_0 \varepsilon_{eff} - \omega^2 L_{L0} d} & \left[1 - \delta \eta_1 \frac{\cos(Kz_0 + \theta_1)}{\omega^2 L_{L0} d \varepsilon_0 \varepsilon_{eff} - 1} - \delta^2 \eta_2 \frac{\cos(2Kz_0 + \theta_2)}{\omega^2 L_{L0} d \varepsilon_0 \varepsilon_{eff} - 1} \right] \\
 & \frac{\partial}{\partial x} (H_{y0} + \delta H_{y1} + \delta^2 H_{y2}).
 \end{aligned} \tag{3.36}$$

3.5 Solutions for each order problems

3.5.1 Zeroth-order problem

The dispersion results of the DP CRLH TL can be obtained from the solution of the zeroth order problem.

In the substrate medium, the zeroth-order component H_{y0} satisfies the Eqn 3.27 which defines the zeroth order problem. At the interface between the air and substrate ($x = s$), the fields H_{y0} and E_{z0} are continuous. At the ground plane, the field $E_{z0} = 0$ when ($x = 0$). The boundary conditions are obtained as (3.28), where H_{y0s} and H_{y0a} are the fields in the substrate and freespace, respectively.

If we assume the solution in the product form $H_{y0} = f(x)g(z_0)$, and substitute H_{y0} into (Eqn 3.27), we get

$$\frac{f''(x)}{f(x)} + \frac{g''(z_0)}{g(z_0)} + \omega^2 \mu_{s0} \varepsilon_{s0} = 0. \tag{3.37}$$

After substituting $f(x)$ and $g(z_0)$ for H_{y0} and dividing the equations by $f(x)$ and $g(z_0)$, the term $\frac{f''(x)}{f(x)}$ remains as a function of only x and the term $\frac{g''(z_0)}{g(z_0)}$ is dependent only on z_0 .

1) In the case of substrate, the propagation constants in different directions are assumed as

$$\begin{cases} \frac{f''(x)}{f(x)} = -k_{xs}^2, \\ \frac{g''(z_0)}{g(z_0)} = -k_z^2, \end{cases} \quad (3.38)$$

The solutions of $f(x)$ and $g(z_0)$ in the substrate are predicted based on particular form expression given as

$$\begin{cases} f_s(x) = A\cos(k_{xs}x) + B\sin(k_{xs}x) & 0 < x < s \\ g(z_0) = e^{-jk_z z_0}. & x > s \end{cases} \quad (3.39)$$

2) In the case of air, the propagation constants in different directions are assumed as

$$\begin{cases} \frac{f''(x)}{f(x)} = -k_{xa}^2, \\ \frac{g''(z_0)}{g(z_0)} = -k_z^2. \end{cases} \quad (3.40)$$

Solutions of $f(x)$ and $g(z_0)$ in air are supposed to be

$$\begin{cases} f_a(x) = C\cos(k_{xa}x) + D\sin(k_{xa}x) & x > s \\ g(z_0) = e^{-jk_z z_0}. & 0 < x < s \end{cases} \quad (3.41)$$

Figure 3.7: Wave numbers of DP CRLH TLs

The k_{xs} and k_{xa} are the complex wave numbers along x -direction in substrate and air, respectively, and k_z is the complex wave number along the z -direction as shown in Figure 3.7. These wave numbers satisfy

$$\left\{ \begin{array}{ll} \text{In freespace} \\ k_{xa}^2 + k_z^2 = \omega^2 \mu_0 \epsilon_0 & x > s \\ \\ \text{In substrate} \\ k_x s^2 + k_z^2 = \omega^2 \mu \epsilon_{s0} & 0 < x < s \end{array} \right. \quad (3.42)$$

and they are complex numbers. They can be written in terms of phase constant and attenuation

constants as

$$\begin{cases} k_{xa} = k'_{xa} + jk''_{xa} = \beta_{xa} - j\alpha_{xa}, \\ k_{xs} = k'_s + jk''_s = \beta_{xs} - j\alpha_{xs}, \\ k_z = k'_z + jk''_z = \beta_z - j\alpha_z, \end{cases} \quad (3.43)$$

where β_{xa} , β_{xs} and β_z are the corresponding phase constants and α_{xa} , α_{xs} and α_z are the corresponding attenuation constants.

Now we proceed to evaluate the constants A, B, C and D.

1. Based on the boundary conditions as shown in (Eqn 3.28), at $x = 0$, $E_{z0s} = 0$, and then

$\frac{\partial H_{y0s}}{\partial x} = 0$ (substrate), hence

$$\begin{aligned} \frac{\partial H_{y0s}}{\partial x} &= \frac{\partial f_s(x)g(z_0)}{\partial x} \Big|_{x=0} \\ &= \frac{\partial}{\partial x} ([A \cos(k_{xs}x) + B \sin(k_{xs}x)] e^{-jk_z z_0}) \Big|_{x=0} = B k_{xs} e^{-jk_z z_0} = 0. \end{aligned} \quad (3.44)$$

Therefore, $B = 0$.

2. At $x = \infty$, $f_a(x)$ should have limit value as $f_a(x)|_{x=\infty} < \infty$ (air). Therefore

$$\begin{aligned} f_a(x)|_{x=\infty} &= C \cos(k_{xa}x)|_{x=\infty} + D \sin(k_{xa}x)|_{x=\infty} \\ &= C \frac{e^{jk_{xa}x} + e^{-jk_{xa}x}}{2} \Big|_{x=\infty} + D \frac{e^{jk_{xa}x} - e^{-jk_{xa}x}}{2} \Big|_{x=\infty}. \end{aligned} \quad (3.45)$$

Because $\alpha_{xa} > 0$, hence

$$\begin{cases} e^{jk_{xa}x}|_{x=\infty} = e^{(\alpha_{xa} + j\beta_{xa})x}|_{x=\infty} = \infty, \\ e^{-jk_{xa}x}|_{x=\infty} = e^{(-\alpha_{xa} - j\beta_{xa})x}|_{x=\infty} < \infty. \end{cases} \quad (3.46)$$

Therefore, based on the boundary condition, (Eqn 3.45) is assumed as

$$f_a(x) = C \frac{e^{-jk_{xa}x}}{2} - D \frac{e^{-jk_{xa}x}}{2} \doteq E e^{-jk_{xa}x}. \quad (3.47)$$

3. At $x = s$, $H_{y0s} = H_{y0a}$, so that,

$$A \cos(k_{xs}s) e^{-jk_z z_0} \big|_{x=s} = E e^{-jk_{xa}s} e^{-jk_z z_0} \big|_{x=s}. \quad (3.48)$$

Hence, the parameters A and E are

$$\begin{cases} A = \frac{1}{\cos(k_{xs}s)}, \\ E = e^{jk_{xa}s}. \end{cases} \quad (3.49)$$

Based on the differential equations (Eqn 3.27) and the boundary conditions (Eqn 3.28), the fields satisfy the below equations

$$\begin{cases} H_{y0a} = N_g a_g e^{-jk_{xa}(x-s)} e^{-jk_z z_0} & x > s, \\ H_{y0s} = N_g a_g \frac{\cos(k_{xs}s)}{\cos(k_{xs}s)} e^{-jk_z z_0} & 0 < x < s. \end{cases} \quad (3.50)$$

The K_{xa} , k_{xs} and k_z are defined from (3.42) and (3.43).

By applying the boundary condition that the E field (E_z) is continuous on the air-substrate interface $x = s$ (Eqn 3.28) to the fields in the substrate and air with (Eqn 3.50), the solution satisfies the below relation after partially differentiating it with respect to x .

$$\begin{cases} \frac{\partial}{\partial x} H_{y0s} \big|_{x=s} = -N_g a_g k_{xs} \tan(k_{xs}s) e^{-jk_z z_0}, \\ \frac{\partial}{\partial x} H_{y0a} \big|_{x=s} = -j N_g a_g k_{xa} e^{-jk_z z_0}. \end{cases} \quad (3.51)$$

Hence, it is obtained that

$$N_g a_g k_{xs} \tan(k_{xs}s) e^{-jk_z z_0} = j N_g a_g k_{xa} e^{-jk_z z_0}. \quad (3.52)$$

The dispersion relation can be obtained from (Eqn 3.51) and satisfies

$$k_{xs} \tan(k_{xs}s) = \frac{\varepsilon_{s0}}{\varepsilon_0} j k_{xa}. \quad (3.53)$$

The field is a wave of varying amplitude a_g and the unknown normalization constant N_g in (Eqn 3.50) is chosen by considering the power carrier by the guided wave in the z -direction as $|a_g|^2$ over a unit width in the y -direction, where a_g is a guided power along the TL. In the passband, the attenuation constant of the fundamental harmonic in the z -direction (α_z) is zero. The Poynting vector of $|a_g|^2$ toward the z -direction is

$$\begin{aligned} P_z &= \frac{1}{2} \text{Re} \iint (E \times H^*) \cdot e_z dx dy = \frac{1}{2} \text{Re} \iint (E_x H_y^*) dx dy \\ &= \frac{1}{2} \text{Re} \iint \left(-\frac{1}{j\omega\varepsilon} \frac{\partial H_y}{\partial z} H_y^* \right) dx dy \\ &= \frac{1}{2} \text{Re} \int_{y=0}^{y=1} \int_{x=0}^{x=s} \left(-\frac{1}{j\omega\varepsilon} \frac{\partial H_{y0s}}{\partial z} H_{y0s}^* \right) dx dy \\ &\quad + \frac{1}{2} \text{Re} \int_{y=0}^{y=1} \int_{x=s}^{x=\infty} \left(-\frac{1}{j\omega\varepsilon} \frac{\partial H_{y0a}}{\partial z} H_{y0a}^* \right) dx dy. \end{aligned} \quad (3.54)$$

Substituting (Eqn 3.50) and (Eqn 3.43) into (Eqn 3.54), the Poynting vector is obtained as

$$P_z = \frac{\beta_z N_g^2 \alpha_g^2}{4\omega\varepsilon_{s0}\alpha_{xa}\beta_{xs}^2} \left[\frac{\varepsilon_{s0}}{\varepsilon_0} (\alpha_{xa}^2 + \beta_{xs}^2) + s\alpha_{xa}(\beta_{xs}^2 + \alpha_{xa}^2 \frac{\varepsilon_{s0}^2}{\varepsilon_0^2}) \right]. \quad (3.55)$$

Hence, the normalization constant N_g in the passband is obtained as

$$|N_g|^2 = \frac{4\omega\varepsilon_{s0}\alpha_{xa}\beta_{xs}^2}{\beta_z \left[\frac{\varepsilon_{s0}}{\varepsilon_0} (\alpha_{xa}^2 + \beta_{xs}^2) + s\alpha_{xa}(\beta_{xs}^2 + \alpha_{xa}^2 \frac{\varepsilon_{s0}^2}{\varepsilon_0^2}) \right]}. \quad (3.56)$$

3.5.2 First-order problem

Many higher order Floquet modes are generated when the permittivity and the permeability of a medium are harmonically varied. We now consider the solution to first order Floquet

mode along with perturbation orders of δ^1 and δ^2 in first and second order analyses. The corresponding differential equations and boundary conditions are given in (Eqn 3.30) and ((Eqn 3.31), respectively.

Based on the differential equations and boundary conditions, the solutions (H_{y1}) of (3.30) for the first-order are given in terms of general solution and particular integrals. We assume the following solutions in air and substrate mediums.

$$\begin{cases} H_{y1a} = F_{1a}e^{-\alpha_{1a}(x-s)}e^{-j(k_z+K)z_0} + N_re^{j\theta_1}(a_ie^{jk_{-1a}(x-s)} + b_ie^{-jk_{-1a}(x-s)})e^{-j(k_z-K)z_0}, \\ H_{y1s} = \sum_{i=\pm 1} e^{-j(k_z \pm K)z_0} \left[\frac{F_{\pm 1s} \cos(k_{\pm 1s}x)}{\cos(k_{\pm 1s}s)} + \frac{G_{\pm 1s} \cos(k_{xs}x)}{\cos(k_{xs}s)} \right], \end{cases} \quad (3.57)$$

where

$$G_{\pm 1s} = \frac{N_g a_g e^{j\theta_1}}{2(2Kk_z \pm K^2)} \left(\frac{k_z \eta_{L1}}{\varepsilon_0 \varepsilon_{eff} \omega^2 L_0 d - 1} - \frac{\eta_{L1} \mu_{s0}}{L_0 d} - \frac{\eta_{C1} \varepsilon_{s0}}{L_0 d} \right), \quad (3.58)$$

and

In air :

$$\begin{cases} k_{-1a}^2 = \omega^2 \mu_0 \varepsilon_0 - (k_z - K)^2, \\ \alpha_{1a}^2 = -\omega^2 \mu_0 \varepsilon_0 + (k_z + K)^2. \end{cases} \quad (3.59)$$

In the substrate :

$$\begin{cases} k_{-1s}^2 = \omega^2 \mu_{s0} \varepsilon_{s0} - (k_z - K)^2, \\ k_{1s}^2 = \omega^2 \mu_{s0} \varepsilon_{s0} - (k_z + K)^2. \end{cases}$$

The first-order solutions contain $(k_z - K)$ and $(k_z + K)$ terms, which are the wave numbers in the z -direction. For the leaky wave phenomenon to occur atleast one of the scattered Floquet modes must be a fast wave. Solving for the particular integral equations (Eqn 3.57) will give

further expressions for H_{y1}

$$\begin{aligned}
 H_{y1a} &= F_{1a} e^{-\alpha_{1a}(x-s)} e^{-j(k_z+K)z_0} + N_r e^{j\theta_1} (a_i e^{jk_{-1a}(x-s)} \\
 &\quad + b_i e^{-jk_{-1a}(x-s)}) e^{-j(k_z-K)z_0} \\
 H_{y1s} &= \sum_{i=\pm 1} e^{-j(k_z \pm K)z_0} \left[\frac{F_{\pm is} \cos(k_{\pm is} x)}{\cos(k_{\pm is} s)} + \frac{G_{\pm is} \cos(k_{xs} x)}{\cos(k_{xs} s)} \right] \\
 &\quad + \frac{N_g a_g e^{j\theta_1}}{2(2kk_z + k_{xs}^2)} \left[e^{j(k_z-K)z_0} \left(\frac{-\mu_{s0}\eta_{L1}}{L_0 d} - \frac{-\varepsilon_{s0}\eta_{C1}}{C_0 d} + \frac{k_z \eta_{L1}}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} \right) \right] \\
 &\quad + \frac{N_g a_g e^{-j\theta_1}}{2(2kk_z + k_{xs}^2)} \left[e^{-j(k_z-K)z_0} \left(\frac{\mu_{s0}\eta_{L1}}{L_0 d} + \frac{-\varepsilon_{s0}\eta_{C1}}{C_0 d} + \frac{k_z \eta_{L1}}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} \right) \right]
 \end{aligned} \tag{3.60}$$

The normalization factor N_r is evaluated by considering the power flow in the $-x$ and $+x$ direction as incident $|a_i|^2$ and radiated $|b_i|^2$ powers. The Poynting vector in x -direction is given as

$$\begin{aligned}
 P_x &= \frac{1}{2} Re \int_{z=0}^{z=1} \int_{y=0}^{y=1} (E \times H^*) \cdot e_x dy dz = \frac{1}{2} Re \int_{z=0}^{z=1} \int_{y=0}^{y=1} (E_z H_y^*) dy dz \\
 &= \frac{1}{2} Re \int_{z=0}^{z=1} \int_{y=0}^{y=1} \left(\frac{1}{j\omega\varepsilon_0} \frac{\partial H_y}{\partial x} H_y^* \right) dy dz = \frac{1}{j2\omega\varepsilon_0} \frac{\partial H_y}{\partial x} H_y^*.
 \end{aligned} \tag{3.61}$$

Substituting (Eqn 3.57) and (Eqn 3.59) into (Eqn 3.61), the Poynting vector is obtained as

$$P_x = \frac{k_{-1a}}{2\omega\varepsilon_0} N_r^2 (|a_i|^2 - |b_i|^2). \tag{3.62}$$

By normalizing the incident wave and radiated wave, the normalization factor N_r is evaluated as

$$N_r = \sqrt{\frac{2\omega\varepsilon_0}{k_{-1a}}}. \tag{3.63}$$

Figure 3.8: Coupling interaction between incident wave a_i , guided wave a_g and radiated wave b_i

The power flow in the $-x$ and $+x$ -directions is given by the incident power $|a_i|^2$ and radiated power $|b_i|^2$, respectively. The relation between the radiated wave b_i , incident wave a_i and guided wave a_g (as shown in Figure 3.8) with in-phase condition $e^{-j(k_z - k)z_0}$ of $\cos(Kx)$ can be derived as [98]

$$b_i = C_{rg}a_g + C_{rr}a_i, \quad (3.64)$$

where C_{rg} is the coupling coefficient and C_{rr} is the reflection coefficient.

By solving the Eqn 3.60 further and applying the boundary conditions, we obtain the final expression which is used to group the a_g and the a_i terms.

$$\begin{aligned}
 & \frac{1}{\varepsilon_{s0}} \{ AN_r(a_i + b_i)e^{j\theta_1} - \frac{N_g a_g e^{j\theta_1}}{2(2kk_z + k_{xs}^2)} \left[\frac{-\mu_{s0}\eta_{L1}}{L_0 d} - \frac{-\varepsilon_{s0}\eta_{C1}}{C_0 d} + \frac{k_z \eta_{L1}}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} \right] \\
 & + \frac{N_g a_g e^{-j\theta_1}}{2(2kk_z + k_{xs}^2)} \left[\frac{\mu_{s0}\eta_{L1}}{L_0 d} + \frac{-\varepsilon_{s0}\eta_{C1}}{C_0 d} + \frac{k_z \eta_{L1}}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} (-k_{xs} \tan k_{xs} s) \right] \\
 & - a_g N_g \tan k_{xs} s \frac{\omega^2 L_0 d \eta_{L1} \cos(Kz_0 + \theta_1)}{2 \cos(Kz_0) (\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1)^2} \\
 & = j \frac{-k_{-1a}}{\varepsilon_0} N_r e^{j\theta_1} (a_i - b_i)
 \end{aligned} \tag{3.65}$$

The coefficients of a_g and a_i are represented as C_{rg} and C_{rr} respectively. The electric fields E_z and E_x can be obtained from Maxwell's equations. Substituting (3.57) into the boundary conditions of the first-order, F_{-1s} , F_{1s} are obtained and the coupling coefficient C_{rg} , the reflection coefficient C_{rr} are obtained as

$$\begin{aligned}
 C_{rg} = & \frac{\frac{N_g e^{j\theta_1}}{2(2kk_z - k_{xs}^2)} \left[\frac{\mu_{s0}\eta_{L1}}{L_0 d} + \frac{-\varepsilon_{s0}\eta_{C1}}{C_0 d} + \frac{k_z \eta_{L1}}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} \right] (-k_{1s} \tan k_{1s} s - k_{xs} \tan k_{xs} s) \\
 & - N_g \tan k_{xs} s \frac{\omega^2 L_0 d \eta_{L1} \cos(Kz_0 + \theta_1)}{2 \cos(Kz_0) (\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1)^2}}{N_r e^{j\theta_1} \left[\frac{(-k_{1s} \tan k_{1s} s)}{\varepsilon_{s0}} + \frac{k_{-1a}}{\varepsilon_0} \right]} \\
 C_{rr} = & - \frac{\varepsilon_0 k_{-1s} \tan(k_{-1s} s) + j k_{-1s} \varepsilon_{s0}}{\varepsilon_0 k_{-1s} \tan(k_{-1s} s) - j k_{-1s} \varepsilon_{s0}}
 \end{aligned} \tag{3.66}$$

3.5.3 Second-order problem

In order to determine the guided mode amplitude and the phase change of the periodic structure, we need to solve the second order analysis. The inhomogeneous second-order problem has a solution if a certain solvability condition is satisfied. This solvability condition is the amplitude transport equation governing the nature of interactions between the zeroth, first, and second-order fields.

The corresponding differential equations and boundary conditions are given in Eqn 3.32 and Eqn 3.33, respectively. Based on the differential equations and boundary conditions, the solutions (H_{y2}) of Eqn 3.32 for the second-order are obtained in the form

$$\left\{ \begin{array}{l} \text{In air} \\ H_{y2a} = \phi_a(x)e^{-jk_z z_0} \quad x > s, \\ \text{In substrate} \\ H_{y2s} = \phi_s(x)e^{-jk_z z_0} \quad 0 < x < s. \end{array} \right. \quad (3.67)$$

Substituting (Eqn 3.67) into (Eqn 3.32), and solving the differential equations for $\phi_a(x)$ and $\phi_s(x)$, the solutions are obtained as

$$\left\{ \begin{array}{l} \phi_a(x) = A_1 e^{-jk_{xa}(x-s)} - \frac{jk_z}{jk_{xa}} N_g \frac{da_g}{dz_2} x e^{-jk_{xa}(x-s)}, \\ \phi_s(x) = B_1 \frac{\cos(k_{xs}x)}{\cos(k_{xs}s)} + C_1 \frac{\sin(k_{xs}x)}{\cos(k_{xs}s)} + \frac{jk_z N_g}{k_{xs} \tan(k_{xs}s)} \frac{da_g}{dz_2} \sin(k_{xs}x). \end{array} \right. \quad (3.68)$$

Substituting (Eqn 3.67) into the boundary conditions for the second-order at $x = 0$, yields $C_1 = 0$. At $x = s$, followed by the elimination of the arbitrary constants, the amplitude transport equation is obtained as

$$\frac{da_g}{dz_2} = C_{gg}a_g + C_{gr}a_i, \quad (3.69)$$

In the LWA problem, the incident wave a_i is not considered, since it is based on the scattering of guided waves into a radiated wave due to the presence of a nonuniformity along the waveguide and hence the relations reduce to

$$b_i = C_{rg}a_g \quad \text{and} \quad \frac{da_g}{dz_2} = C_{gg}a_g. \quad (3.70)$$

The second order analysis involves elaborate expansions, thus making the analysis manually unsolvable beyond this order. By solving for the H_{y2} term we obtain 9 sets of particular integrals. Evaluating these integrals individually will provide the necessary expressions to obtain coefficient of a_g (Eqn 3.71).

$$\begin{aligned}
 C_{gg} &= \frac{N}{D} \\
 N &= \left[\frac{\left(\frac{k_z \eta_{L1} k}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} - \frac{\mu_{s0} \eta_{L1}}{L_0 d} - \frac{-\varepsilon_{s0} \eta_{C1}}{C_0 d} \right)}{2(2k k_z - k_{xs}^2)} \right. \\
 &\quad \left. + \frac{\left(\frac{k_z \eta_{L1} k}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} + \frac{\mu_{s0} \eta_{L1}}{L_0 d} + \frac{-\varepsilon_{s0} \eta_{C1}}{C_0 d} \right)}{2(2k k_z + k_{xs}^2)} \right] \\
 &\quad \left[\frac{\mu_{s0} \eta_{L1} N_g}{8 L_0 d k_{xs}} \right] \left[stank_{xs} s - \frac{tank_{xs} s + sk_{xs}}{k_{xs} tank_{xs} s} \right] \\
 &\quad - \frac{k - k_z}{2k_z - k} \left(\frac{-k_z \eta_{L1} k}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} - \frac{\mu_{s0} \eta_{L1}}{L_0 d} + \frac{\varepsilon_{s0} \eta_{C1}}{C_0 d} \right) \\
 &\quad + \frac{k - k_z}{2k_z + k} \left(\frac{k_z \eta_{L1} k}{\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1} + \frac{\mu_{s0} \eta_{L1}}{L_0 d} - \frac{\varepsilon_{s0} \eta_{C1}}{C_0 d} \right) \\
 &\quad \left[\frac{\eta_{L1} N_g}{8 k_{xs} (\omega^2 \varepsilon_0 \varepsilon_{eff} L_0 d - 1)} \right] \left[stank_{xs} s - \frac{tank_{xs} s + sk_{xs}}{k_{xs} tank_{xs} s} \right] \\
 &\quad + \frac{\eta_{L1} \eta_{C1} N_g stank_{xs} s}{4 k_{xs} \omega^2 L_0 C_0 d^2} \\
 D &= \left\{ \frac{jk_z N_g}{k_{xs}} + \frac{\varepsilon_{s0}}{k_{xs} tank_{xs} s} \left[\frac{jk_z N_g}{\varepsilon_0 \alpha_i} + \frac{jk_z N_g}{\varepsilon_{s0} k_{xs}} (tank_{xs} s + sk_{xs}) - \frac{jk_z N_g s}{\varepsilon_0} \right] \right\}
 \end{aligned} \tag{3.71}$$

Equation (3.69) constitutes a pair of canonical equations relating the guided wave, the incident wave, and the radiated wave. $C_{gg} = C_{ggr} + jC_{ggi}$ is the extinction coefficient. Its real part C_{ggr} gives the leakage coefficient and the imaginary part C_{ggi} determines the exact radiation angle for the optimum radiation efficiency.

In a finite length L_g of the permittivity modulated TL, the radiation efficiency Q_0 and radiation angle θ_r of the LW radiating from the TL with double periodically loaded inductances

and capacitances can be obtained from

$$\begin{cases} Q_0 = \frac{\int_0^{L_g} |b_i|^2 dz_2}{|a_g|_{z_2=0}^2} = 1 - e^{2C_{ggr}L_g}, \\ \theta_r = \tan^{-1}\left(\frac{k_{-1a}}{\beta - K - C_{ggi}}\right), \end{cases} \quad (3.72)$$

where L_g is the total length of the periodic TL. The radiation efficiency is defined as the ratio of the total power radiated from the modulated region to the guided wave power incident at $z_2 = 0$ [98].

3.6 Numerical results

It is assumed that the thickness of the TL is = 1.6 mm, the average loaded inductance $L_0 = 5.05$ nH, the loaded capacitance $C_0 = 0.5$ pF, periodicity of the loaded elements is $d = 10$ mm, relative permittivity of the substrate $\varepsilon_r = 4.9$, and it is considered as non-magnetic. We can calculate the phase constant of the fundamental space harmonic, only when we assume the periodic permittivity modulation applied is same as the unperturbed phase constant. When $k_{xs} \neq 0$, we need to consider quasistatic TM mode solution.

Dispersion diagram aids a better understanding of left Handed and the right Handed regions of the given structure. Until the zeroth order, the analysis remains similar to only one parameter modulation. In Figure 3.9 we can see the dispersion results of the proposed double periodic TL.

We can identify three main regions: (a) New Leaky wave region, (b) Left Handed Passband (c) Right Handed Passband. The frequencies at which these properties are displayed are $f_1 = 1.023$ GHz, $f_2 = 2.467$ GHz and $f_3 = 4.567$ GHz respectively. The curves at f_1 , f_2 and f_3 are inside the airline boundary ($\beta = \omega\sqrt{\mu_0\varepsilon_0}$). In addition to the fact that the proposed antenna radiates both backward and forward at frequencies f_2 and f_3 respectively, it can ra-

Figure 3.9: Dispersion characteristics of the double periodic TL based on singular perturbation procedure

diates a forward wave at $f_{leaky} = f_1$ which is below the LH passband. The bandwidth of the stopband is comparatively wider than in the case of single periodic TL[1]. In both RH and LH passband, the phase constant of the double periodic TL approaches that of single periodic structure. These results are further verified against the measured frequency response of the double periodic structure.

From the singular perturbation method, we also obtain the radiation efficiency Q_0 . We have obtained the radiation efficiency at around frequencies f_1 and f_2 . We have four parameters which play an important role in deciding the radiation efficiency viz. η_{L1} , η_{C1} , η_{L2} and η_{C2} . The values of the modulation index have been suitably chosen so as to give better efficiency.

Figure 3.10: Radiation efficiency as a function of antenna length for different modulation indices

The two combinations of values chosen are

case 1: $\eta_{L1} = 0.1$, $\eta_{C1} = 0.2$, $\eta_{L2} = 0.2$ $\eta_{C2} = 0.1$

case 2: $\eta_{L1} = 0.3$, $\eta_{C1} = 0.3$, $\eta_{L2} = 0.2$ $\eta_{C2} = 0.2$.

The radiation efficiency Q_0 is numerically estimated around the new right-handed leaky-wave region ($f_{leaky} = f_1$) and the LH leaky-wave region ($f = f_2$) as a function of TL length Z_g as shown in Figure 3.10. It can be seen that the radiation efficiency for the new leaky-wave right-handed region is smaller than the radiation behavior in the LH region. It may be developed for large η values from 0.1 upto 0.3 beyond which it exceeds the limit of singular perturbation procedure [12]. It is noted that the radiation efficiency at $f_{leaky} = f_1$ is less than

that at $f = f_2$ and longer transmission lines are needed to get better radiation efficiency. In order to correlate the measured efficiency of the fabricated prototype with the numerically obtained efficiency, the values are compared at the antenna length $Z_g = 0.79\text{mm}$. It was found that the measured results provided an efficiency of -0.1dB for the frequency $f_2 = 2.467\text{ GHz}$ which seem to match the theoretically predicted value. However, for $f_1 = 1.023\text{ GHz}$, the efficiency was found to be around -6dB which happens to be higher than that predicted by the spm analysis. This could be owing to the assumed values of the modulation indices.

3.7 Fabricated prototype and measurement results

The structure has been realized in a double periodic manner with both inductance and the capacitance modulated along the propagation axis. The structure has been implemented on a microstrip line. The periodicity for the given structure (d) is 10mm with eight unit cells. The dimensions of the structure follow the single perturbation analysis and they can be referred from Figure 3.5. The width of the microstrip line is 3.7mm and the thickness is 1.6mm . The proposed structure is implemented on a FR-4 substrate with relative dielectric constant $\epsilon_r = 4.9$ with an overall length of 79mm .

Figure 3.11: Fabricated proposed double periodic TL structure

Figure 3.11 shows the fabricated prototype of the proposed double periodic CRLH TL. It can be seen that the inductance that has been realized as a stub with a via, and the capacitance which has been realized as gap between interdigitated fingers which is as shown in Figure 3.2, have both been varied in length in a double periodic manner. We can see the measured scattering parameters of the TL in Figure 3.12. Comparing the measured frequency response with the numerically obtained dispersion results, we find that the S-parameters of the double periodic TL which as predicted by the SPM exhibits a new leaky wave phenomenon. The LH region, stopband and RH region of operation can be demarcated based on the theoretically obtained dispersion results.

Figure 3.12: Measured frequency responses of TLs

The experimentally verified [111] frequency response also indicates that no leaky wave property in the LH passband can be found in the single periodic CRLH TL. This highlights the fact that double periodic TL has an advantage over single periodic TL in terms of LH leaky wave radiation in the given frequency range.

The far-field radiation patterns of the proposed antenna designed using microstrip line double periodically loaded with both series interdigital capacitances and shunt inductances are measured at $f_1 = 1.023$ GHz (new leaky wave, forward radiation), $f_2 = 2.467$ GHz (LH, backward radiation) and $f_3 = 4.567$ GHz (conventional RH, forward radiation) as shown in Figure 3.13. The SPM provides the approximate radiation angle for the structure as predicted from the second order solution. In this scenario, we can see that the theoretically calculated radiation angle at $f_1 = 1.023$ GHz is found to be 18° , $f_2 = 2.467$ GHz is -16° (indicating LH radiation) and at $f_3 = 4.567$ GHz it is 21° . Both the simulated and measured results have been shown and except for the gain values, they seem to be in good agreement.

Figure 3.13: Simulated and measured radiation pattern of the DP CRLH TLs at (a) $f_1 = 1.023$ GHz, (b) $f_2 = 2.467$ GHz and (c) $f_3 = 4.567$ GHz

3.8 Summary

In this chapter, we have proposed an analysis for a double periodic TL structure with both inductance and capacitance loaded. Single perturbation method has been used to analyze the CRLH TL structure. Numerical Expressions have been obtained which aid in obtaining dispersion characteristics, propagation constant and radiation characteristics. We have included the modulation of the permeability which has increased the degree of freedom of the analysis. The radiation efficiency has been plotted as a function of antennas length to illustrate the effect of modulating indices on the antenna performance. It has been obtained at the three frequencies of interest. We also observe a novel leaky wave region with narrow bandwidth in the desired frequency band. The experimental validation of the numerical results have also been provided.

Chapter 4

Composite right/left handed based substrate integrated waveguide leaky wave antenna with periodic slots

In the previous chapter, the CRLH LWA based on microstrip was discussed. This chapter presents the implementation of CRLH LWA on substrate integrated waveguide and half mode substrate integrated waveguide (HMSIW) platforms. We discuss the design of v shaped slot leaky wave antenna implemented on a SIW platform which radiates in both backward and forward directions. The design procedures of the SIW and the slot are discussed and the measured frequency responses are presented.

4.1 Problem definition

There has always been a need for wide bandwidth antennas with high gain values. Also, the necessity for antennas with good performance characteristics at higher frequency is in demand. This chapter aims at designing compact leaky wave antennas that can operate at

higher frequencies with wide 10-dB bandwidth, have good gain values and provide broadside radiation in addition to backfire to endfire beam scanning with frequency variation .

4.2 Introduction

A new technology for implementing microwave planar components known as substrate integrated waveguide (SIW) was proposed by Hirokawa [12]. At higher frequencies, the existing platforms like microstrip and coplanar waveguide face loss related issues. The capability to handle high power and high value of quality factor combined with advantages like simplicity, light weight and low cost at higher frequencies, give SIW an edge over the existing platforms. Presently, SIWs are becoming an inevitable choice for the implementation of high frequency integrated circuits. Basically SIWs are planar implementation of conventional rectangular waveguides. A typical SIW structure consists of two parallel metal plates separated by a dielectric substrate, with the bottom plate being the ground plane. Along the broader side of the metal plates, there are two rows of via holes or also called conducting cylinders, which are parallel to each other. These two rows of via holes connect the two metal plates through the substrate, thus confining the electromagnetic energy within the structure. They also connect the surface currents which maintain guided wave propagation in the structure [112].

SIWs offer a favorable platform for realizing LWAs. The LWAs designed on SIWs are very compact, while retaining other antenna characteristics intact. SIWs that have been implemented as CRLH transmission lines have been used as leaky wave antennas that were capable of steering the main beam from backward to forward direction whenever frequency was varied [14]. There is a constant effort to improve the characteristics of these antennas. Previous works have indicated that LWAs based on CRLH SIW concept can achieve high gain levels while providing a beam scanning from broadside to endfire [113, 114, 115, 116].

In the next sections we explain the design aspects of the proposed structure, its equivalent

circuit, before proceeding to discuss the results obtained through full wave simulations.

4.3 Design steps

The initial procedure for designing any SIW based LWA begins with the design of basic SIW structure. The choice of dimensions is made based on the desired frequency range of operation. The slots etched on the surface act as radiating elements. The design steps include [87]

1) The SIW dimensions are chosen depending on the frequency of operation. The important design parameters are

- Via hole diameter(d)
- Spacing between the via holes (s)
- Effective width between the two rows of via holes (w) where w is given by

$$w = W[SIW] - 1.08 \frac{d^2}{s} + 0.1 \frac{d^2}{W[SIW]} \quad (4.1)$$

where $W[SIW]$ is the initial overall width of the SIW.

These dimensions need to satisfy the constraints of

$$d < \frac{\lambda_g}{5} \quad \text{and} \quad s \leq 4d$$

2) *Choice of substrate:* The substrate is chosen based on the desired application of the structure.

For radiating applications, substrate with lower permittivity values is chosen so that the losses are minimized. For our case, we have chosen RT duroid substrate with $\epsilon_r=2.2$.

3) *Choice of slot shape:* It was shown in [117] that a slot etched on the top surface of the SIW contributes to series capacitance generating LH behaviour. From the basics of CRLH

TL concepts, we know that different slot shapes translate into different capacitance values which in turn affect the resonating frequency of the structure.

- 4) *Number of radiating elements*: The choice of length of LWA is made based on the number of radiating elements. They control the amount of power that propagates and leaks before reaching the terminating port. This is chosen based on the optimum values of gain that can be achieved by the structure.
- 5) *Feed design*: The SIW structure can have different feed types like gap coupled, stub feed or tapered feed. The tapered microstrip feeding has the advantages of being less lossy, capacity to cover the entire bandwidth [118].

These design aspects decide the performance characteristics of SIW structures when applied for leaky wave applications.

4.4 CRLH SIW LWA

In this section, we propose a SIW LWA based on CRLH TL concept. V shaped slots have been etched on the top surface of the structure. Its radiation characteristics have been studied.

4.4.1 SIW design

The initial design of the proposed structure begins with the design of basic SIW. The parameter values are chosen based on the desired frequency band and the substrate. The dimensions of the SIW are chosen such that the proposed structure operates between 31 GHz to 46 GHz. We first compute the width between the two rows of via holes (w) based on the design equations provided in [4] which decides the operating frequency range of the SIW. The diameter (d) of the via is 0.77mm and the spacing (s) between them is 1.45mm. These values satisfy the criteria

of the ratios $s/d < 3$ and $d/w < 1/5$ such that the width between the two rows of via holes to be accurate for given substrate with permittivity of 2.2 and height of 0.508mm.

Figure 4.1: SIW with microstrip taper design

Another important design aspect includes the integration of SIW with other platforms. In this design, we provide a tapered microstrip transition [118]. It is widely used and has quite a few advantages of being less lossy, capacity to cover the entire bandwidth. The process of matching becomes simplified. Basically, the taper transforms quasi TEM mode of the microstrip line into dominant TE_{10} mode of the rectangular waveguide. The taper length is generally chosen to be a multiple of quarter wavelength, so that the return loss is minimal. The taper length depends on the difference between the width of the SIW and the microstrip line. In this design, since the difference is not too large, we chose an optimum taper length (t_1) of 2.8mm. The width of the microstrip feed line (t_2) which matches 50Ω is found to be 1.55mm (Figure 4.1)

4.4.2 Equivalent circuit

Figure 4.2: Equivalent circuit of a single unit cell of the proposed CRLH SIW LWA

Figure 4.2 shows the equivalent circuit of the proposed structure which consists of RH and LH components. The v shaped slot has been introduced into the model as C_L . L_L represents the inductance generated by the via-walls. Together they correspond to the LH contribution. Right-handed (RH) behaviour is owing to the distributed shunt capacitance C_R and L_R , which are contributed by the distributed series inductance from the waveguide and the parasitic inductance from the fingers.

Varying the waveguide width will result in different values of shunt inductance L_L which would also lead to the change of the waveguide cutoff frequency. By changing the slot width and length, the value of the series capacitor C_L can be adjusted conveniently, which will determine the position of LH region. It has been studied in Dong et al [95] that when C_L is large, the LH band will move to a lower frequency thus generating a band gap between the LH and RH ranges.

Figure 4.3: Dispersion diagram of CRLH SIW LWA

Figure 4.3 presents the dispersion diagram for the CRLH SIW unit cell with parameter values given as $L_L=0.1\text{nH}$, $C_L=0.9\text{pF}$, $L_R=0.25\text{nH}$, $C_R=0.65\text{pF}$. The period p is small compared to the wavelength. To achieve balanced condition, the frequencies of the series and shunt elements have to be equal. The LH to RH transition happens at this frequency. In the proposed structure, by adjusting the slot width to 0.68mm , it was found that the balanced condition could be achieved.

Two methods can be adopted to calculate the dispersion curve for a single unit cell. One is based on the eigen-mode simulation by applying periodic boundary conditions. The plot of phase constant with frequency (dispersion diagram) of the unit cell is obtained by the Eigen mode solver of commercial CST MWS software. The phase is varied between 0° to 180° in the parameter sweep. The mesh was set to be tetrahedral and the number of modes was selected to be 2. The simulated results indicate that the transition from LH to RH region occurs at around

33.8 GHz. The airline cuts the higher frequency at around 46 GHz and the LH frequency around 31.8 GHz. It can be seen that LH region is exhibited at frequency range 31 GHz to 33.8 GHz. And the RH region is between 34 GHz and 46.5 GHz. The Eigen-mode simulation method is more accurate but time consuming [95].

The other method is based on the S-parameters from driven mode simulation using equation 4.2.

$$\cosh(\gamma) = (1/p)[(1 - |S_{11}|^2 + |S_{21}|^2) + (Z_{01}/Z_{02})(1 - |S_{11}|^2 + |S_{21}|^2)]/4S_{21} \quad (4.2)$$

The propagation constant γ is given as $\gamma = \alpha + j\beta$, where α is the attenuation constant and β is the phase constant. Z_{01} and Z_{02} are the impedances at port 1 and port 2 respectively, which are set to be 50Ω . To give a comparison, Figure 4.3 plots the dispersion curves of a balanced CRLH SIW unit cell using both of the two approaches. A good agreement is observed between the two curves.

4.4.3 Slot design

For the SIW structure to operate as a leaky wave antenna, one of the methods is to etch slots on the top patch surface. The shape, position and the orientation of the slot decides the radiation pattern. We propose inverted v shaped slot to enable leaky wave radiation. This slot can be viewed as a combination of two oppositely inclined slots. It is known that when a slot is inclined at an angle φ to the axis of the SIW, the power it radiates is proportional to $\sin 2\varphi$ [117].

Figure 4.4: Slot dimensions.

In our case, the location of the slot disturbs the current density in the x direction. The y-components are also perturbed but they get cancelled out. Since the slots are oppositely inclined to each other, the final radiation pattern will have a narrow beam, which can be exploited to achieve an electronic pivoting of antenna beam as a function of change of transmission signal frequency. In the proposed design the slot dimensions are chosen such that length of the slot (L) is approximately equal to $\lambda/2$ and the slot width (w_1) is less than $\lambda/10$. This ensures optimum radiation from the slots. The two inclined slots make an angle of 90° between them. The distance between the slots is optimized and found to be around 2.8mm. The horizontal total slot length (L_1) is 2.6mm (Figure 4.4).

4.4.4 Simulated design

Figure 4.5 shows the proposed CRLH based SIW LWA. It comprises of 13 unit cells. It was found that increasing the number of unit cells further reduced the gain values considerably. The structure has been designed on Rogers 5880 substrate with ϵ_r values of 2.2 and height of 0.508mm. The total length of the leaky wave antenna (L_{SIW}) is 74.6mm and the width (W_{SIW}) is 7.2mm. The proposed structure has been designed and simulated using the transient solver

of commercially available full wave simulation tool CST MWS. The dispersion characteristics were obtained by running the Eigen mode solver of the CST MWS with two modes. The performance of the structure will be discussed in the next section.

Figure 4.5: Overall proposed SIW LWA (All dimensions in mm).

Figure 4.6: Current distribution of the proposed LWA at different frequencies.

Figure 4.6 shows the current distribution of the proposed SIW structure at LH region (32 GHz) and RH region (36 GHz and 44 GHz) of operation.

4.4.5 Results

The structure has been fabricated using the standard PCB process and the S parameter measurements have been carried out. Figure 4.7 shows the fabricated photo of the structure. The via holes have been fabricated by completely filling the hole area with the copper metal so as to connect the top and bottom layers through the substrate.

Figure 4.7: Fabricated prototype of proposed CRLH SIW LWA

Figure 4.8 and Figure 4.9 show the measured and simulated S-parameters of this leaky-wave antenna, which are in good agreement. Owing to the frequency limitations of the network analyzer, the measurements have been carried out only until 43.5 GHz. A satisfactory return loss above 10 dB in the band of interest (from 31.2 GHz to 48 GHz) is achieved. The insertion loss is almost below -10 dB, which indicates good leakage radiation. The curve also shows that in the LH region, the radiation is more effective compared with that in the RH region. The discrepancy of S_{21} between the simulation and measurement is due to the increase of reflection, the loss from the SMA connectors and probably the increased conductor loss.

Figure 4.8: S_{11} of the proposed structure

Figure 4.9: S_{21} of the proposed structure

The far field radiation pattern of the proposed leaky wave antenna has been simulated at different frequencies. The antenna radiates in backward direction from 31 GHz to around 34 GHz, and forward from 34.4 GHz to around 48 GHz. It can be seen that as the frequency increases, there is scanning from the backfire to endfire direction. There is radiation at the broadside in the transition frequency. The frequency dependence of the beam can be viewed clearly from the Figure 4.10.

Figure 4.10: Simulated 3-D radiation patterns at different frequencies of the proposed structure.

Figure 4.10 shows the simulated 3-D radiation patterns. It is seen that, when the frequency is increased, the main beam moves from the backfire towards the endfire direction. At the transition frequency, the radiation goes exactly to the broadside.

Figure 4.11: Simulated normalized radiation patterns of the proposed structure in the y-z plane at 32.5 GHz (a) x-z plane at 32.5 GHz (b) in LH region. The y-z plane at 33.8 GHz (c) x-z plane at 33.8 GHz (d) at transition from LH to RH.

Figure 4.12: Simulated normalized radiation patterns of the proposed structure in the y-z plane at 43 GHz (e) x-z plane at 43 GHz (f). The y-z plane at 46.5 GHz (g) x-z plane at 46.5 GHz (h) RH region.

The normalized two dimension radiation patterns for the proposed SIW leaky wave antenna at four different frequencies have also been shown in Figure 4.11 and Figure 4.12. In the LH region, the beam scans from -32° starting at frequency 31.5 GHz until the frequency is increased to 33.8 GHz in the y-z plane. In the RH region, the beam scanning takes place from frequency 34 GHz to around 46.5 GHz upto 61° .

A continuous scanning can be observed from the pattern in these frequency ranges. We observe a fan shaped beam in the x-z plane of the scanning range. In the LH region the

antenna radiates backward and the beam is broader in comparison to the RH region which is the common behavior observed in most of CRLH LWAs.

Figure 4.13: Simulated antenna gain of the proposed structure

As seen from Fig 4.13 the realized gain values in the LH region are comparatively lower ranging around 6-8dB while in the RH region it varies around 13dB.

Figure 4.14: Simulated antenna efficiency of the proposed structure

Fig 4.14 shows the radiation efficiency for the proposed structure. The radiation efficiency

is obtained from the S-parameters using the relation given by equation 4.3

$$\eta = (1 - |S_{11}|^2 - |S_{21}|^2). \quad (4.3)$$

The average efficiency for the scanning range is given to be 81%. The results obtained from simulation will be verified with the fabrication of the designed antenna.

4.5 CRLH HMSIW LWA

In this section, we propose an extension of the structure proposed in the previous section. The LWA is implemented on a SIW operating at its half mode. When a structure is adapted to its half mode counterpart, it is reduced into almost half of its original size. This happens when we draw a fictitious line along the magnetic wall, thus dividing the original structure into two parts. However, in our case, we have implemented the V slot entirely on the HMSIW owing to its asymmetric shape.

4.5.1 HMSIW configuration

The HMSIW configuration is obtained by drawing the fictitious line along transmission direction of the SIW structure dividing it into two halves and adjusting the feed. The width of the structure is reduced to almost half of its SIW counterpart. The taper feed has been retained from the SIW LWA design on one side of the structure such that 50Ω impedance matching is ensured.

4.5.2 Equivalent circuit

Figure 4.15: Equivalent circuit of a single unit cell of the proposed CRLH HMSIW LWA

Figure 4.15 presents the equivalent circuit model for the original HMSIW TLs without the slots, which is similar to the traditional rectangular waveguide and is symmetrical. The metal surface and the ground can be modeled as a two-wire TL with distributed series inductance and distributed shunt capacitance which are associated with the permeability and permittivity of the substrate, respectively. It should be noted that the input impedance of a short circuited stub (via-walls) after a piece of TL appears as inductive.

The V shaped slot has been introduced into the model as C_L . L_L represents the inductance generated by the via-walls. Together they correspond to the LH contribution. Right-handed (RH) behaviour is owing to the distributed shunt capacitance C_R and L_R , which are contributed by the distributed series inductance from the waveguide and the parasitic inductance from the fingers.

Varying the waveguide width will result in different values of shunt inductance L_L which would also lead to the change of the waveguide cutoff frequency. By changing the slot width and length, the value of the series capacitor C_L can be adjusted conveniently, which will determine the position of LH region. It has been studied in Dong et al [95] that when C_L is large, the LH band will move to a lower frequency thus generating a band gap between the LH and RH ranges.

Figure 4.16: Dispersion diagram of CRLH HMSIW LWA

Figure 4.16 presents the dispersion diagram for the CRLH SIW unit cell with parameter values given as $L_L=0.4\text{nH}$, $C_L=0.2\text{pF}$, $L_R=0.15\text{nH}$, $C_R=0.46\text{pF}$. The period p is small com-

pared to the wavelength. It must be noted that in case of HMSIW there exists a bandgap between the transition of LH and RH regions. This is owing to the unbalanced condition of the proposed HMSIW TL. The structure provides a scanning from backward (LH) to forward (RH) directions.

The dispersion results have been obtained based on the S-parameters from driven mode simulation using equation 4.2.

The results indicate that the LH radiating region is from 32.4 GHz to 35.2 GHz and RH region occurs between 38.3 GHz to 48.4 GHz. The airline demarcates the guided and the radiating regions. It can be observed from the plot that there exists a distinct bandgap between 35.2 GHz to 38.3 GHz.

4.5.3 Simulated design

Figure 4.17 shows the proposed CRLH based HMSIW LWA. It comprises of 13 unit cells. It was found that increasing the number of unit cells further reduced the gain values considerably. The structure has been designed on Rogers 5880 substrate with ϵ_r values of 2.2 and height of 0.508mm. The total length of the leaky wave antenna (L_{SIW}) is 74.6mm and the width (W_{SIW}) is 4.6mm. The proposed structure has been designed and simulated using the transient solver of commercially available full wave simulation tool CST MWS.

The slot design for the structure has been retained from SIW LWA discussed in the previous section. However since the shape of the slot cannot be bisected by retaining symmetry, it has been entirely implemented on the HMSIW structure.

Figure 4.17: Overall proposed HMSIW LWA.

Figure 4.18: Current distribution of the proposed HMSIW LWA at 33 GHz.

Figure 4.19: Current distribution of the proposed HMSIW LWA at 42 GHz.

Fig 4.18 (LH region) and Fig 4.19 (RH region) show the current distribution of the proposed structure at two different frequencies, one in the LH region (33 GHz) and other in the RH region (42 GHz).

4.5.4 Results

The structure has been fabricated and the S parameter measurements have been carried out. Fig 4.20 shows the fabricated photo of the structure.

Figure 4.20: Fabricated prototype of proposed CRLH HMSIW LWA

Figure 4.21 and 4.22 show the measured and simulated S-parameters of this leaky-wave antenna, which are in good agreement. Owing to the frequency limitations of the network analyzer, the measurements have been carried out only until 43.5 GHz. A satisfactory return loss above 10 dB in the band of interest (from 32.4 GHz to 35.2 GHz and 38.3 GHz to 48.4 GHz) is achieved. The insertion loss is almost below -10 dB, which indicates good leakage radiation. The curve also shows that in the LH region, the radiation is more effective compared with that in the RH region. The discrepancy of S_{21} between the simulation and measurement is due to the increase of reflection, the loss from the SMA connectors and probably the increased conductor loss.

Figure 4.21: S_{11} of the proposed structure

Figure 4.22: S_{21} of the proposed structure

The far field radiation pattern of the proposed leaky wave antenna has been simulated at different frequencies. The antenna radiates in backward direction from 32.4 GHz to around 35.2 GHz, and forward from 38.3 GHz to around 48 GHz. It can be seen that as the frequency increases, there is scanning from the backfire to endfire direction. There is no radiation at the broadside due to the stopband. The frequency dependence of the beam can be viewed clearly from the Fig 4.23.

Figure 4.23: Simulated 3-D radiation patterns at different frequencies of the proposed structure.

Figure 4.23 shows the simulated 3-D radiation patterns. It is seen that, when the frequency is increased, the main beam moves from the backfire towards the endfire direction.

Figure 4.24: Simulated normalized radiation patterns of the proposed structure. (a) y-z plane at 33.2 GHz, (b) x-z plane at 33.2 GHz, (c) y-z plane at 34.5 GHz (d) x-z plane at 34.5 GHz of LH region

Figure 4.25: Simulated normalized radiation patterns of the proposed structure. (a)y-z plane at 40.5 GHz, (b) x-z plane at 40.5 GHz,(c)y-z plane at 46 GHz (d) x-z plane at 46 GHz of LH region

The normalized two dimension radiation patterns for the proposed SIW leaky wave antenna at two different frequencies have been shown in Fig 4.24. and Fig 4.25. In the LH region, the beam scans from -44° starting at frequency 32.4 GHz until the frequency is increased to 35.2 GHz. There is no broadside radiation. In the RH region, the beam scanning takes place from frequency 38.3 GHz to around 48GHz going upto 72° . A continuous scanning can be observed from the pattern except near the broadside in these frequency ranges. We observe a fan shaped beam in the x-z plane of the scanning range. In the LH region the antenna radiates backward and the beam is broader in comparison to the RH region which is the common behavior observed in most of CRLH LWAs.

Figure 4.26: Simulated antenna gain and efficiency of the proposed structure

As seen from Fig 4.26 the realized gain values in the LH region are comparatively lower ranging around 6-8dB while in the RH region it varies around 13dB. The radiation efficiency is obtained from the S-parameters using the relation given by equation 4.3

The average efficiency for the scanning range is given to be 89%. The results obtained from simulation will be verified with the fabrication of the designed antenna. The efficiency observed is much better than the SIW LWA since the HMSIW has lower dielectric and conductor losses.

4.6 Comparison studies

This section compares the performances of different LWAs implemented on SIW platforms.

4.6.1 Comparison of CRLH SIW with and without V shaped slots

To understand the impact of V shaped slots on the SIW, the frequency responses of SIW with no slots and SIW with V shaped slots are compared as shown in Figure 4.27 and Figure 4.28. It can be clearly seen that the transmission characteristics have been altered due to the capacitive slots etched on the structure. The LH components contributed by L_L and C_L give an additional backward wave passband which is below the cut off frequency of the original SIW.

Figure 4.27: Comparison of S_{11} of SIW with no slots and SIW with V shaped slots

Figure 4.28: Comparison of S_{21} of SIW with no slots and SIW with V shaped slots

4.6.2 Comparison of CRLH SIW and CRLH HMSIW with V shaped slots

The S-parameters of the proposed CRLH SIW and HMSIW with V slots have been plotted in Figure 4.29 and Figure 4.30. It can be clearly seen that there exists a stopband in case of CRLH HMSIW with V shaped slots between 35.2 GHz and 38.3 GHz unlike in CRLH SIW LWA with V shaped slots. In case of the latter, the transition from LH region to RH region occurs at 33.8 GHz which is smooth. Owing to the unbalanced condition, the HMSIW structure cannot radiate in the broadside direction.

Figure 4.29: Comparison of S_{11} of SIW and HMSIW with V shaped slots

Figure 4.30: Comparison of S_{21} of SIW and HMSIW with V shaped slots

4.6.3 Performance comparison of proposed structure with other reported structures

In order to validate the advantages of the proposed structures, a performance comparison has been made with few other CRLH SIW LWAs reported in literature by Dong et al.[95] and Jan Machac et al. [119]. Table 4.1 summarizes the comparison. The proposed structures have the advantages of higher frequency, wide bandwidth and compact size.

Parameter	Proposed CRLH SIW LWA with V shaped slots	Proposed CRLH HMSIW LWA with V shaped slots	Dong et al	Jan Machac et al.
Frequency of operation (GHz)	30-50	30-50	8-14	8-20
Frequency scanning angles (deg)	93°	116°	88°	60°
Simulated avg peak gain (dB)	14	13	12.5	15
Simulated peak efficiency(%)	83	89	82	81

Table 4.1: Performance characteristics of CRLH based SIW LWAs

4.7 Summary

This chapter presented a LWA implemented on a SIW platform based on CRLH TL concept. Two antennas were proposed based on SIW and HMSIW technology. The CRLH SIW LWA had V shaped slots etched on its top surface. The SIW design aspects were discussed and its equivalent circuit is provided. The initial measurements were performed with available SMA connectors. The measurement can be performed with 2.4 mm (upto 50GHz) and 1.85 mm (upto 60GHz) connectors for the proposed high frequency structures. The dispersion characteristics have been obtained and the LH and the RH frequency ranges are identified. The transition from LH to RH region occurs at around 33.8 GHz. The frequency responses and the radiation patterns are simulated and the beam angle scans from -32° in the LH region to 61° in the RH region. The structure has a wide bandwidth and is very compact.

A CRLH LWA was proposed on HMSIW platform with V shaped slots. This antenna is unbalanced and hence does not support broadside radiation. Its performance is almost comparable with its CRLH SIW LWA counterpart. Comparisons have been made with similar CRLH SIW LWAs reported in literature. It is seen that the proposed structures have the advantages of higher frequency, wide bandwidth and compact size. Owing to the limited measurement facilities, the radiation pattern measurement for the proposed antenna is yet to be made.

Chapter 5

Substrate integrated waveguide based resonator antennas

Having seen the advantages of the SIW platform in the previous chapter, this chapter discusses the substrate integrated resonator antenna which is designed to work at octa mode. Section 5.2 provides the necessary background of the SIW square resonator and its subsequent extension and adaptation to obtain the proposed octa mode SIW. It discusses the structure details, field distributions and resonant frequency modes of half mode, quarter mode and octa mode SIW resonators. Section 5.3 gives the steps involved in the design of the proposed antenna structure. It elaborates on the design procedure starting from the choice of substrate to obtaining a desired frequency of resonance of the proposed resonator. Section 5.4 focuses on the implementation of a quarter mode substrate integrated waveguide antenna whose design has been extended to obtain the proposed octa mode resonator antenna. The simulated and the measured results are discussed. The proposed octa mode resonator antenna that operates at dominant resonant frequency of 2.4 GHz is discussed. The variation of various design parameters and its impact on the structure have been presented. The last Section concludes with the performance comparison of the proposed structure with its counterparts.

5.1 Problem definition

There is always a demand for compact planar antennas which have competent performance characteristics. Keeping this necessity in mind, our objective was to design a resonator antenna that operates at 2.4 GHz for wireless LAN applications. SIW platform was chosen to implement this antenna given its benefits over other transmission lines like microstrip and CPW. The performance characteristics of these resonator antennas were investigated so as to meet the desired requirements. The SIW resonator antennas were designed to operate at a usable wireless frequency while having improved radiation properties like increased gain and efficiency.

5.2 SIW based cavity resonators

The cavity resonator can be implemented on a planar platform in the form of substrate integrated waveguide (SIW). SIW has metallic posts which are equivalent to side metallic walls of a conventional waveguide (Figure 5.1). The field and frequency expressions for the conventional rectangular waveguide resonator hold good for the SIW resonator too.

Figure 5.1: SIW equivalent of cavity resonator

5.2.1 SIW

We begin the discussion of the SIW resonator with the design of the square wave resonator. A square SIW resonator is obtained by setting the same value of width (W) and length (L) of a rectangular cavity resonator. The length and the width of the SIW are kept the same to simplify its adaptation to its miniaturized structures. In this way the symmetry can be retained. In this case, the TE_{101} mode will remain the same as square SIW resonator and the modal components can be expressed as shown in Eqn 5.1 [13], where H_m represents the amplitude of the modal components.

$$\begin{aligned} E_y &= -j\frac{\omega\mu L}{\pi}H_m\sin(\frac{\pi}{L}x)\sin(\frac{\pi}{L}z) \\ H_x &= -H_m\sin(\frac{\pi}{L}x)\cos(\frac{\pi}{L}z) \\ H_z &= H_m\cos(\frac{\pi}{L}x)\sin(\frac{\pi}{L}z) \\ E_x &= E_z = H_y = 0 \end{aligned} \tag{5.1}$$

As discussed before, SIW is a planar realization of the conventional rectangular cavity and its design equations have been provided by Deslandes and K Wu [87]. These equations ensure that the equivalent dimensions for SIW are the same as that of the rectangular waveguide. The design equations establish constraints on the diameter d of via holes and the spacing between them denoted by s . The constraints are

$$d < \frac{\lambda_g}{5} \quad \text{and} \quad s \leq 4d$$

In our case, we have chosen the diameter of via holes to be 1.4mm and the spacing between them given by 2.8mm. These values of d and s seem to cause less radiation losses and it is the closest the SIW cavity can get to act as a rectangular waveguide resonator.

In order to have better confinement of the EM fields in the cavity we have employed two rows

of via holes. They have been aligned optimally, thus ensuring maximum retention of the EM field within the cavity. It was observed that for the proposed SIW structures this arrangement gave slightly better results in comparison to single row of via holes.

To have a symmetry in the SIW that will facilitate further reduction, we have maintained same length and width of the SIW, thus designing a square SIW resonator. The initial configuration and simulated design can be seen in Figure 5.2.

Figure 5.2: SIW square resonator (a) Layout (b) Simulated design

It can be seen that the size of this initial SIW is $35.42 \times 35.42 \text{ mm}^2$. This length serves as a reference for the further designs of HMSIW, QMSIW and OMSIW structures. This resonator has been designed on a FR4 substrate with permittivity of 4.3 and thickness h of 3.2mm. The relation between the dimensions of the SIW resonator structure and frequency of resonance is determined using the relation given by Eqn 5.2

Parameter	Values (mm)
GL_{SIW}	90
GW_{SIW}	100
L	32.42
g_1	53.17
g_2	47.03
g_3	12.7
f_1	19.8
m_1	1.8
m_2	2.9
m_3	2.35
m_4	1.79
m_5	18.38

Table 5.1: Dimensions of SIW

$$f_{mnp}[SIW] = \frac{1}{2\pi\sqrt{\mu\epsilon}} \sqrt{\left(\frac{m\pi}{L_e}\right)^2 + \left(\frac{n\pi}{h}\right)^2 + \left(\frac{p\pi}{L_e}\right)^2} \quad (5.2)$$

Where μ and ϵ represent the permeability and permittivity of the substrate respectively. However, since the ratio of SIW width to height is quite large, only TE_{m0p} modes can resonate in the structure with $n=0$. Thus reducing the frequency expression to

$$f_{101}[SIW] = \frac{c}{2\sqrt{\mu_r\epsilon_r}} \frac{\sqrt{2}}{L_e} \quad (5.3)$$

We can write the expression for effective length L_e given in 5.2 as [87]

$$L_e[SIW] = L - 1.08 \frac{d^2}{s} + 0.1 \frac{d^2}{L} \quad (5.4)$$

L_e physically represent the effective length of the SIW between the rows of via holes, i.e. it is the effective space available for the EM field on the structure. The diameter of the via hole is represented by d and s represents the spacing between adjacent vias.

Since m and p values can be interchanged, the first 3 modes of resonant frequencies for

this structure have been calculated. Using the expression for the resonant frequency given by equation 5.2, initial 3 modes have been tabulated as shown in Table 5.2.

$\begin{matrix} \backslash \\ m \end{matrix} \begin{matrix} p \end{matrix}$	1	2	3
1	3.38	6.48	12.2
2		9.56	14.365
3			16.424

Table 5.2: Resonant frequencies in GHz for the square SIW resonator

Their corresponding electric field distributions are plotted in Figure 5.3. Figure 5.3(a) shows the electric field distribution at 3.38 GHz with TE_{101} mode. The distribution of TE_{102} mode at 6.48 GHz is shown in Figure 5.3(b). Figure 5.3(c) and Figure 5.3(d) show the electric field distribution of the TE_{202} modes at 9.56 GHz and 10.85 GHz, respectively and they are similar, which indicates that the square SIW resonator works at TE_{202} mode at both frequencies.

Figure 5.3: Electric field distribution at different resonant modes of SIW resonator (a) TE_{101} at 3.38 GHz (b) TE_{102} mode at 6.48 GHz (c) TE_{202} mode at 9.56 GHz and 10.85 GHz

The structure shown in Figure 5.2 has been simulated and the resonant frequencies almost agree with the theoretically obtained values. Figure 5.4 shows the simulated return loss of the full SIW cavity resonator. It can be seen that there are resonances at 3.38 GHz (TE_{101}), 6.48 GHz (TE_{102}), 9.24 GHz and 10.82 GHz (TE_{202}) along with higher order resonances. The 10-dB impedance bandwidth of the TE_{101} is found to be around 150 MHz.

Figure 5.4: Simulated return loss of the square SIW resonator indicating different resonating modes

5.2.2 Half mode SIW (HMSIW)

In 2006, Wei Hong et al. proposed this half-mode substrate integrated waveguide (HMSIW) concept [120]. HMSIW is implemented by placing an open circuit along the symmetric plane in the transmission direction. This open side keeps the remaining field distributions intact by acting as a perfect magnetic wall. This is possible because the field distribution in the SIW cavity is symmetrical. It has been shown that this HMSIW structure has low loss characteristics

like its original full SIW component [95]. Also, the attenuation is lower compared with a SIW structure. This idea can be extended to SIW resonator too. By bisecting the SIW with a fictitious magnetic wall, each half of SIW becomes an HMSIW structure while nearly preserving the original field distribution. We can observe from the current field distributions of the SIW that, along the centre plane the fields are symmetric along the z direction.

(a) Transition from SIW to HMSIW.

(b) Layout of HMSIW structure.

Figure 5.5: Configuration of HMSIW.

The maximum fields can be found at places on the SIW which are on the plane of symmetry. Since the SIW W/h ratio is quite large, the magnetic field which is normal is zero at the planes of symmetry. Due to this reason, these planes can be considered equivalent to a quasi-magnetic wall. If we draw the bisecting line along AOC of the SIW resonator, then we obtain the HMSIW resonator. The configuration of the HMSIW resonator is given in Figure 5.5. The original length L of the SIW is retained by one side of the HMSIW structure. The dimensions of the HMSIW structure are as provided in Table 5.3

Parameter	Values (mm)
GL_{HMSIW}	80
GW_{HMSIW}	100
L	32.42
g_1	43.17
g_3	47.03
g_4	12.7
f_2	9.82
m_1	1.8
m_2	2.9
m_3	2.35
m_4	1.79
m_5	18.38

Table 5.3: Dimensions of HMSIW

The field distribution of the HMSIW wave shown in 5.6 depicts the symmetrical other half of the original SIW field behaviour.

The fundamental mode in SIW is almost similar to half the quasi TE_{101} mode in the SIW. The y component of the E field can be estimated as a quarter cosine in the HMSIW and it decays rapidly beyond the open side of the resonator. The effective width of the HMSIW is half that of full SIW. The equivalent width of the HMSIW is given by Eqn 5.5

$$L_e[HMSIW] = \frac{L_e[SIW]}{2} + \Delta L \quad (5.5)$$

Figure 5.6: Electric field distribution at dominant resonant frequency of HMSIW resonator

$$\Delta L = h(0.05 + \frac{0.03}{\epsilon_r}) \times \ln(0.79 \frac{L_e[SIW]}{4h^3}) + \frac{52L_e[SIW] - 261}{h^2} + \frac{38}{h} + 2.77 \quad (5.6)$$

Based on this it can be found that the dominant resonant frequency of the HMSIW remains the same as SIW.

$$f_{101}[HMSIW] = \frac{c}{2\sqrt{\mu_r\epsilon_r}} \frac{\sqrt{2}}{L_e[HMSIW]} \quad (5.7)$$

The structure shown in Figure 5.5 has been simulated and the resonant frequencies almost agree with the theoretically obtained values. Figure 5.7 shows the simulated return loss of the full SIW cavity resonator. It can be seen that there are resonances at 3.15 GHz (TE_{101}), 7.12 GHz (TE_{102}), 9.77 GHz and 11.39 GHz (TE_{202}) along with higher order resonances. The 10-dB impedance bandwidth for TE_{101} mode is given to be 260 MHz.

Figure 5.7: Simulated return loss of the square HMSIW resonator indicating different resonating modes

5.2.3 Quarter mode SIW (QMSIW)

In order to facilitate further size reduction, the HMSIW structure can be bisected again along the magnetic wall to obtain a quarter mode SIW which is nearly 75% reduction in size when compared to original structure. It has been observed [101] that the Q value of QMSIW structure is usually lower than their conventional SIW and HMSIW counterparts since the two open edges are not perfect magnetic walls and some amount of leakage radiation may happen. But the QMSIW is capable of obtaining a much stronger coupling coefficient between two neighbour sections with proper coupling means. By drawing the cutting plane perpendicular to side AOC, we obtain a quarter of the original SIW resonator (Figure 5.8). The side OA becomes the primary source of radiation, while side OB is the secondary radiating side. This dissection leaves us with two sides OA and OB each of dimension $\frac{L}{\sqrt{2}}$. Since we already know the value

of L , each open side of QMSIW measures 33.37mm . The via hole side AB is retained as it is from the HMSIW structure. The TE_{101} mode is used as a reference to draw the cutting plane.

(a) Transition from QMSIW to OMSIW.

(b) Layout of QMSIW structure.

Figure 5.8: Configuration of QMSIW.

The quasi magnetic wall at the center plane of the SIW structure results in a HMSIW design(Figure 5.9). Another symmetrical plane exists along the x axis. By cutting a fictitious

Figure 5.9: QMSIW field distribution

line along this plane we get a QMSIW structure. It is not possible to bisect the structure when it operates at TE_{102} mode. The expected modes of the QMSIW are TE_{101} , TE_{202} , TE_{303} etc. TE_{102} mode disappears. QMSIW is exactly quarter section of the full square SIW structure. It has been shown that this QMSIW can be used for microwave filters, antennas and other components application. It was shown in [100] that the effective width and the length of the QMSIW remains the same as that of HMSIW since the m_6 dimension does not change.

Feeding technique is an important design part, which influences the input impedance and characteristics of the antenna. The bandwidth can be widened by the inset microstrip feed [1]. The inset microstrip feed is used to widen the bandwidth. The inset feed design for the proposed structure has been retained from [101] due to its advantages of higher efficiency, wider bandwidth, and more important, it is convenient to use the inset feeding. Hence the expression for the resonant frequency remains the same for the QMSIW structure.

$$L_e[QMSIW] = L_e[HMSIW] \quad (5.8)$$

Parameter	Values (mm)
GL_{QMSIW}	78.2
GW_{QMSIW}	46.4
g_5	18.1
g_6	21.2
g_7	17.2
f_3	7.1
m_1	1.8
m_2	2.9
m_3	2.35
m_4	1.79
m_5	22.38
m_6	32.20

Table 5.4: Dimensions of QMSIW

$$f_{101}[QMSIW] = \frac{c}{2\sqrt{\mu_r\epsilon_r}} \frac{\sqrt{2}}{L_e[QMSIW]} \quad (5.9)$$

The observed resonant frequency [102, 121] indicate that the frequencies of the QMSIW are slightly shifted compared to that of their SIW counterpart. This occurs because of the fringing fields at the two open edges. This is accounted to in the expression by the addition of the ΔW term.

Figure 5.10: Simulated return loss of QMSIW resonator indicating different resonating modes

Figure 5.8(b) shows the simulated layout of the proposed structure. The overall dimension of the structure is $45 \times 68 \text{ mm}^2$. The justification for the choice of ground dimensions is provided at the end of the chapter. Figure 5.9 shows the E field distribution. It can be seen that the wall OA has the maximum signal concentration at TE_{101} mode. Further studies demonstrate that the radiation properties are contributed by side OA. Figure 5.10 shows the simulated return loss of the proposed QMSIW resonator antenna. The antenna resonates at 2.57 GHz with 10-dB return loss bandwidth of 2.49 GHz to 2.67 GHz. It must be noted that there is no resonance for TE_{102} mode. This antenna has been implemented on a FR-4 substrate with thickness 3.2mm.

5.2.4 Octa mode SIW(OMSIW)

If we continue dissecting the QMSIW along the fictitious cutting line (magnetic wall), we obtain an Octamode SIW resonator. In this section, we propose an octamode SIW resonator that operates at 2.42 GHz. In comparison to its SIW resonator counterpart, this structure is reduced to 1/8th of its size. However, the performance characteristics are not lost during the miniaturization. If we draw the cutting line perpendicular to the side AB, then we remain with only half of the QMSIW structure (Figure 5.8). The side OA is retained from the QMSIW while side OB is replaced by side OX. By geometry, we can calculate the length of each side keeping original length of SIW as reference. Thus we obtain $OA = \frac{L}{\sqrt{2}}$ and $OX = \frac{L}{\sqrt{2}}$, thus providing the length of AX to be equal to $L/2$. The isosceles triangle of QMSIW with sides $\frac{L}{\sqrt{2}}$ is transformed into isosceles right angled triangle with sides $L/2$. The TE_{101} mode has been used as a reference to draw this cutting plane.

(a) Transition from QMSIW to OMSIW.

(b) Layout of HMSIW structure.

Figure 5.11: Configuration of OMSIW.

Figure 5.11(b) shows the simulated layout of the proposed structure and its dimensions are given in table 5.5 . The overall dimension of the structure is $38.4 \times 68.2 \text{ mm}^2$. This structure is more compact in comparison to QMSIW while retaining nearly all of its characteristics. The OMSIW structure retains one of the radiating edges from its predecessor QMSIW and hence the original resonant frequency of the QMSIW is almost unchanged. However it has been observed that as the size of the SIW keeps getting halved, the dominant resonant frequency experiences a

slight downward shift. Figure 5.12 shows the electric field distribution of the OMSIW structure at TE_{101} mode.

Figure 5.12: Electric field distribution at dominant resonant frequency of OMSIW resonator

Parameter	Values (mm)
GL_{OMSIW}	60
GW_{OMSIW}	40
g_8	15
g_9	18.5
g_{10}	15.2
f_4	6.1
m_1	1.8
m_2	2.9
m_3	2.35
m_4	1.79
m_5	22.38
m_7	22.20

Table 5.5: Dimensions of OMSIW

Extending the analysis of the QMSIW to OMSIW, we observe that the expression for the resonant frequency remains the same. This is justified by the experimental results where the

dominant frequency of resonance is retained by OMSIW. Hence we can write

$$f_{101}[OMSIW] = \frac{c}{2\sqrt{\mu_r\epsilon_r}} \frac{\sqrt{2}}{L_e[OMSIW]} \quad (5.10)$$

The field components of the OMSIW for the TE_{101} mode are given by

$$\begin{aligned} E_y &= E_0 \sin\left(\frac{\pi}{L}x\right) \sin\left(\frac{\pi}{L}z\right) \\ H_x &= \frac{j\pi E_0}{k\eta L} \sin\left(\frac{\pi}{L}x\right) \cos\left(\frac{\pi}{L}z\right) \\ H_z &= \frac{-j\beta E_0}{k\eta} \cos\left(\frac{\pi}{L}x\right) \sin\left(\frac{\pi}{L}z\right) \\ E_x &= E_z = H_y = 0 \end{aligned} \quad (5.11)$$

Figure 5.13: Simulated return loss of the OMSIW at dominant resonant frequency mode

Figure 5.13 shows the simulated return loss of the proposed OMSIW resonator antenna. The antenna resonates at 2.42 GHz with 10 dB BW of 2.38 GHz to 2.47 GHz. This antenna has been implemented on a FR4 substrate with thickness 3.2mm.

5.2.5 Comparative studies

In order to understand the shift in the resonant frequencies as the size of the SIW structure is reduced, we compare the return loss results of SIW, HMSIW, QMSIW and OMSIW as shown in Figure 5.14.

Figure 5.14: Comparison of return losses of SIW, HMSIW, QMSIW and OMSIW

Parameter	SIW	HMSIW	QMSIW	OMSIW
Resonant frequency (GHz)	3.38	3.15	2.57	2.42
10 dB BW (MHz)	150	260	180	80

Table 5.6: Performance comparison of SIW, HMSIW, QMSIW and OMSIW for TE_{101} mode.

Theoretically, the resonant frequency of the SIW, HMSIW, QMSIW, and OMSIW must be the same, but a slight difference is observed because the feeding position is varied and the magnetic walls are not ideal due to fringing fields. Nevertheless, the decrease in the resonant frequency from the SIW to the OMSIW gives us the benefit of further size reduction. It must be noticed that as the mode is halved, the separation between the resonant frequency modes increases, thus making it difficult to tune the resonant modes to desired frequency bands.

5.3 Design steps

This section provides the procedure and steps for designing SIW resonator antennas for desired frequency band.

5.3.1 Design parameters

The design parameters that affect the performance of the proposed OMSIW structure are

1) Choice of the substrate

- Dielectric constant (ϵ_r)
- Thickness (h)

2) Dimensions of the structure/patch

- Effective length
- Effective width
- Via hole spacing
- Via hole diameter

3) Dimensions of the ground plane

These design parameters have an impact on the frequency response of the proposed structure. The dimensions obtained from these expressions can be suitably verified and optimized using commercial full wave analysis software like CST MWS. Also, the dimensions can be tuned accordingly for the structure to operate at the desired frequencies. Based on these parameters, the design steps can be deduced for the proposed OMSIW structure.

5.3.2 Design procedure

Following steps provide the general procedure for the designing an octa mode SIW resonator.

- 1) Based on the frequency constraints, the initial length and width of the resonator is chosen using the formula given in Eqn 5.12 where L, W and h represent the length and width of the cavity.

$$f_{mnp}[SIW] = \frac{1}{2\pi\sqrt{\mu\epsilon}} \sqrt{\left(\frac{m\pi}{L}\right)^2 + \left(\frac{n\pi}{h}\right)^2 + \left(\frac{p\pi}{W}\right)^2} \quad (5.12)$$

- 2) After obtaining the initial L and W values, the effective length and width are calculated using the formula given by Eqns 5.13 and 5.14

$$L_e[SIW] = L - 1.08\frac{d^2}{s} + 0.1\frac{d^2}{L} \quad (5.13)$$

$$W_e[SIW] = W - 1.08\frac{d^2}{s} + 0.1\frac{d^2}{W} \quad (5.14)$$

- 3) Based on the L_e and W_e , obtain resonating frequencies at different TE and TM modes.
- 4) Obtain field expressions at these frequencies as given by Eqn 5.1.
- 5) The via hole diameter and spacing between them are chosen based on the constraints $s/d \ll 0.05$ and $d/w > 5$
- 6) The choice of substrate is made based on the desired application of the structure. For our case, we chose the substrate taking into consideration that the proposed structure is

expected to radiate. For the ease of availability, FR-4 with $\epsilon_r=4.3$ and $h=3.2\text{mm}$ has been chosen.

- 7) The dimensions of the ground plane need to be optimized so as to not compromise on the gain values yet retaining a compact size.

5.4 QMSIW resonator antenna

In order to implement an octa mode resonator, we start by designing a QMSIW resonator. We only take into account the linearly polarized dominant TE_{101} mode of resonance for radiation. As seen, the size of the quarter mode structure has undergone one fourth reduction when compared to its square SIW resonator counterpart.

5.4.1 Structure details

Until now, we have seen the design details of the QMSIW structure. Now we discuss its implementation on a FR4 substrate with permittivity 4.3. Figure 5.15 shows the prototype of the implemented QMSIW structure. This has been implemented on a FR4 substrate with thickness of 3.2mm.

The feed design has been retained from the QMSIW design proposed in [101] owing to its simplicity in designing. A 50 ohms matching is ensured by adjusting the width of the microstrip feed accordingly. In this case, the optimal feed width for 50Ω impedance matching was found to be $f_3 = 7.1\text{mm}$.

Figure 5.15: Fabricated prototype of QMSIW

The fundamental resonant frequency is contributed by side OA. Any change in dimension to this side will affect the dominant frequency, thus making it the primary source of radiation. The proposed structure has two open side walls denoted by OA and OB and AB is the via hole side wall. The radiation characteristics of this structure are owing to the slot behaviour from these two open side walls.

It was found that the ground dimensions affect the gain and the back lobe radiation. It has been optimized for roughly $50 \times 75 \text{ mm}^2$ for optimum performance.

5.4.2 Results

The structure has been fabricated and the measurements have been carried out. Figure 5.15 shows the fabricated photo of the structure. Figure 5.16 shows the return loss of the proposed OMSIW structure. It can be seen from the S-parameter results that the simulated dominant mode resonance occurs at 2.57 GHz with a 10-dB BW of 180 MHz. It can also be seen that the simulated resonance is at the centre frequency of 2.57 GHz, while the measured frequency of resonance is at 2.49 GHz for TE_{101} mode. This discrepancy of 0.07 GHz could be owing

to fabrication errors. Also, the 10-dB bandwidth for the simulated structure is found between frequency range 2.49-2.67 GHz, thus resulting in the total 10-dB impedance BW of 2.12% while the measured 10-dB BW is found to be 3.01%.

It can be pointed out that the measured scattering parameters are de-embedded at the two specified reference planes [122, 123, 124]. When a model is used for the test fixture, in the measurement system, it is necessary to perform the de-embedding process. The considered de-embedding is port extensions, which mathematically extends the measurement plane towards the device under test.

It is assumed that the fixture has no loss, a linear phase response and constant impedance. Port extension only adds or subtracts phase length from the measured scattering parameters. Proper calibration of the vector network analyzer is required with Open-Short-Load-Through standard to get the accurate measurement results. This calibration process creates a reference plane at the point where the standards are connected.

Figure 5.16: Simulated and measured return loss of QMSIW

Figure 5.17: Copolarized radiation pattern of QMSIW resonator antenna

Figure 5.18: Crosspolarized radiation pattern of QMSIW resonator antenna

The radiation characteristics are simulated and measured to evaluate the proposed structures applicability to antennas. The radiation pattern has been measured at typical frequencies based on the resonant modes. The far field radiation pattern at 2.57 GHz of the proposed OMSIW structure has been measured at $\phi=45^\circ$ cut plane. The orientation of θ and ϕ has been as shown in Figure 5.15 .

Figure 5.17 shows the co-polarized radiation pattern at 2.57 GHz with simulated gain of 6.49dBi and measured gain of 4.5dBi. The cross polarized radiation pattern is given by Figure 5.18 with simulated gain of -7.2dBi and measured gain of -4.4dBi. The main lobe radiates at a direction of $\theta = 55^\circ$ to the broadside direction at TE_{101} mode.

Figure 5.19: Gain of QMSIW resonator antenna

Figure 5.20: Efficiency of QMSIW resonator antenna

Figure 5.19 shows the simulated and measured gain values at the corresponding radiating regions. The simulated gain is 5.84dBi at 2.57 GHz and measured gain of 3.56dBi at 2.51 GHz for the proposed linearly polarized resonator antenna at TE_{101} mode. The radiation efficiency can be estimated based upon the relative power absorbed (RPA) of the element [125, 113, 71, 126]. It can be mentioned that the value of the RPA represents the total power losses of the electromagnetic device under investigation. The losses may include radiation loss, metal loss, dielectric loss, and surface-wave loss. Notably, neither the conductor loss nor the dielectric loss is separated from the leakage power in the RPA calculation. These non-separated losses are the main reason causing the difference between the simulated and the measured results.

Radiation efficiency is plotted in Figure 5.20 and it shows a simulated efficiency of 82.29% and measured efficiency of 75.42%. This dip in the measured values of gain and efficiency in

comparison to simulated values is due to different losses.

5.5 OMSIW resonator antenna

In this section we propose a compact octamode SIW resonator antenna. As seen, the size of the octamode structure has undergone considerable reduction when compared to its square SIW resonator counterpart. Until now, we have seen the design details of the OMSIW structure. Now we discuss its implementation on a FR-4 substrate with permittivity 4.3.

The feed design has been retained from the QMSIW design owing to its advantages. A $50\ \Omega$ matching is ensured by adjusting the width of the microstrip feed accordingly. In this case, the optimal feed width for 50Ω impedance matching was found to be $f_4 = 6.1\text{mm}$. The fundamental resonant frequency is contributed by side OA. Any change in dimensions to this side will affect the dominant frequency, thus making it the primary source of radiation.

The proposed structure has two open side walls denoted by OA and OX and AX is the via hole side wall. The radiation characteristics of this structure are owing to the slot behaviour from these two open side walls.

It must be noted that the TE_{101} mode operates in linearly polarized manner. The higher order modes allow circular polarization characteristics [101]. However since we are interested in only single band of operation for short range communication systems, we retain only the dominant TE_{101} resonant mode.

5.5.1 Proposed structure details

Figure 5.21 shows the simulated design and its fabricated prototype. The structure has been implemented on FR-4 substrate with permittivity of 4.3 and thickness of 3.3mm. As predicted the performance of the structure is better when implemented on the Rogers substrate in com-

parison to FR-4 substrate. But since our aim is to demonstrate the concept, the performance of the reported structure has been slightly compromised by using the FR-4 substrate for the ease of availability of the substrate. A dielectric constant of 4.3, loss tangent of 0.09 have been used for the simulation. The copper vias have the dimension of 1.4 mm diameter and are separated by a spacing of 2.8mm. The size of the entire SIW structure is 80mm x 100mm and when implemented as OCTA mode, it has been reduced to 40mm x 60mm, thus resulting in a total reduction of 72%.

Figure 5.21: Fabricated prototype of OMSIW

5.5.2 Results

In order to experimentally demonstrate the performance of the proposed structure, the OMSIW antenna has been fabricated. Figure 5.21 shows the fabricated prototypes of the OMSIW antenna. Figure 5.22 shows the return loss of the proposed OMSIW structure. It can

be seen that the simulated resonance is at the centre frequency of 2.42 GHz, while the measured frequency of resonance is at 2.48 GHz for TE_{101} mode. This discrepancy of 0.06 GHz could be owing to fabrication errors. Also, the 10-dB impedance bandwidth for the simulated structure is found between frequency range 2.38-2.46 GHz, thus resulting in the total 10dB BW of 2.52% while the measured 10-dB BW is found to be 2.34%. The simulation and the measurement results are almost in agreement.

Figure 5.22: Simulated and measured return loss of OMSIW

Figure 5.23: Copolarized radiation pattern of OMSIW resonator antenna

Figure 5.24: Crosspolarized radiation pattern of OMSIW resonator antenna

The radiation characteristics are simulated and measured to evaluate the proposed structures applicability to antennas. The radiation pattern has been measured at typical frequencies based on the resonant modes. The far field radiation patterns at 2.42 GHz of the proposed OMSIW structure have been measured at $\Phi=45^\circ$ cut plane. The orientation of θ and ϕ has been shown in Figure 5.21. Figure 5.23 shows the co-polarized radiation pattern at 2.42 GHz with simulated gain of 5.84dBi and measured gain of 3.56dBi. The cross polarized radiation pattern is given by Figure 5.24 with simulated gain of -3.8dBi and measured gain of -7.2dBi. The main lobe radiates at a direction of $\theta=15^\circ$ to the broadside direction at TE_{101} mode.

Figure 5.25: Gain of OMSIW resonator antenna

Figure 5.26: Efficiency of OMSIW resonator antenna

Figure 5.25 shows the simulated and measured gain values at the corresponding radiating regions. The simulated gain is 6.49dBi at 2.42 GHz and measured gain of 4.5dBi at 2.48 GHz for the proposed linearly polarized resonator antenna at TE_{101} mode. This difference in values between simulated and measured is due to the fabrication errors involved in etching and a few inherent losses of the structure.

Efficiency is plotted in Figure 5.26 and it shows a simulated efficiency of 84.43% and measured efficiency of 72.5 %. This dip in the measured values of gain and efficiency in comparison to simulated values is due to different losses.

5.5.3 Comparitive study

As seen until now, the QMSIW and the OMSIW resonator antennas have been simulated and fabricated. To have a better understanding of the influence of miniaturization, we have compared the performance characteristics of the two antenna prototypes. These results are indicative of the difference in radiating properties of the proposed structure in comparison to the QMSIW resonator antenna. The comparison results have been summarized in Table 5.7.

	QMSIW		OMSIW	
	Simulated	Measured	Simulated	Measured
Resonant Frequency (GHz)	2.57	2.49	2.42	2.48
10-dB BW (MHz)	180	80	80	60
Gain (dBi)	6.49	4.5	5.84	3.56
Efficiency (%)	82.29	75.42	84.43	72.5

Table 5.7: Performance comparison of QMSIW and OMSIW

As discussed before, when we look at the resonant frequencies of these two antennas, we see a downward shift from 2.57 to 2.42 GHz. The 10-dB bandwidth is narrower in OMSIW when compared to QMSIW. The gain and the efficiencies have not experienced a significant change, thus emphasizing the fact that OMSIW provides a performance comparable to QMSIW, while having a more compact size.

5.6 Effect of different parameters

In this section we study the impact of variation of different parameters on the proposed structure. This will provide an insight into the parameters affecting the behaviour of the structure. It will also provide information of the degrees of freedom while designing the structure. The parameters that have been varied are the size of the ground plane and different substrate.

5.6.1 Effect of the variation of ground plane width

In order to study the impact of the ground width on the antenna performance, we simulated five different ground dimensions. The length and width of the overall ground size were varied individually and together to observe the impact of each variation. Table 5.8 provides the behaviour on various ground plane sizes. The dimensions (in mm) of the different variations are provided in Figure 5.27 with values of $g_{11} = 10$, $g_{12} = 20$, $g_{13} = 25$ and $g_{14} = 20.2$. It can be seen that as the width of the ground plane reduces the resonant frequency experiences a slight upward shift from 2.42 to 2.45 GHz. As the width of the ground plane increases the resonant frequency experiences a downward shift.

Ground plane sizes (mm)	Resonant frequency (GHz)	Total Gain (dBi)	Radiation Efficiency (%)
$GL_{OMSIW} \times GW_{OMSIW}$	2.42	6.49	84.43
$GL_{OMSIW} \times GW_1$	2.45	6.21	84.96
$GL_{OMSIW} \times GW_2$	2.42	6.86	84.12
$GL_{OMSIW} \times GW_3$	2.41	6.98	83.86
$GL_1 \times GW_2$	2.37	6.92	82.92
$GL_1 \times GW_3$	2.33	7.01	82.43

Table 5.8: Simulation performances of OMSIW resonator antenna with varied ground plane sizes

Figure 5.27: Proposed OMSIW with varied ground plane sizes (All dimensions in mm)

Figure 5.28: Return loss of OMSIW with varied ground sizes

Figure 5.27 shows the antenna with different ground dimensions. The detailed dimension values can be found in table 5.8. In Figure 5.27(b), the width of the antenna has been reduced compared to the proposed structure. This results in slight increase in the resonant frequency. However, the efficiency is increased by 0.53%. The gain value experiences a reduction. In Figures 5.27(c) and (d), the width is increased which results in increased gain values but a slight decrease in efficiency and resonant frequency. Figures 5.27(e) and (f) represent cases where the length of the ground plane is varied along with the width. Since the side along the feed line is the primary source of radiation, we can see that increase in length causes an improvement in the gain values, while the efficiency is reduced. Figure 5.28 shows the return loss of the OMSIW antenna with varied ground plane sizes. It can be seen that the shift in the resonant frequencies are not very significant.

Since the backlobe radiation is reduced as the ground size increases, the gain values show improvement. However, the radiation efficiency of the proposed structure almost remains the same because the overall power radiated remains unaffected by the width of the ground plane. Since these variations of the performance parameters are not very significant an optimum ground plane size of $40 \times 60 \text{ mm}^2$ has been chosen to implement the proposed OMSIW resonator antenna.

5.6.2 Effect of substrate

As discussed in the design steps, the choice of substrate plays an important role in determining the performance characteristics of the antenna structure[127]. The permittivity and the substrate thickness affect the radiation properties of the antenna. In order to study these effects of the proposed OMSIW resonator antenna, we implemented the structure on Rogers 5880 with permittivity 2.2 and thickness 2.54mm. The dimensions of the structure were suitably modified to obtain a comparable resonance at 2.4 GHz. Table 5.9 provides a comparison of the results of the proposed structure when implemented on these two substrates.

Substrate	Permittivity ε	Loss tangent ($\tan \delta$)	Substrate thickness (mm)	Resonant frequency at TE_{101} mode (GHz)	Simulated Gain (dBi)	Efficiency (%)	10-dB Bandwidth (MHz)
RT duroid 5880	2.2	0.09	2.54	2.44	8.82	91.10	122
FR-4	4.3	0.008	3.20	2.42	6.46	82.42	80

Table 5.9: Simulated comparison results of proposed antenna implemented on RT duroid and FR4 substrates

It can be inferred from the simulation results that the performance of the proposed structure is definitely much better when implemented on RT duroid substrate. It is because the losses associated with the RT duroid substrate are much less when compared to FR4. The dip in the gain values can be seen very clearly. Also, the thickness of the substrate values is lesser for RT duroid. However to maintain low cost and ease of fabrication, FR-4 substrate has been used for the proposed design.

5.7 Summary

In this chapter we have studied the characteristics of a square SIW resonator and its subsequent miniaturization. By effectively drawing the fictitious cutting lines along the magnetic walls, the square SIW resonator has been made to operate at different modes. This concept has been extended to HMSIW, QMSIW and OMSIW. Proceeding with the same idea we propose an octamode SIW resonator that operates at 2.4 GHz. In addition to the frequency of operation, attempt has been made to study the behaviour of the frequency shift as the geometry of the structure gets halved with each mode. The comparison of the frequency modes indicate that as the size of the structure gets reduced, the fundamental frequency experiences a downward shift while the higher order modes shift upwards in the frequency range. The QMSIW structure has been simulated and fabricated. The measured results are in good agreement with the simulated result. At last, the proposed octamode SIW resonator has been implemented on a

FR-4 substrate. The advantage it offers are compactness, usable frequency of operation and wider BW compared to previously reported works. The simulated and the measured results demonstrate that the miniaturized structure has comparable performance characteristics with that of its QMSIW counterpart. The structure finds applications in 2.4 GHz applications. The impact of introducing a slot on the proposed structure has been studied in the next chapter.

Chapter 6

Octamode substrate integrated waveguide antennas with slots

This chapter presents the OMSIW resonator antenna with two different types of slots etched on the top patch surface. Two types of slot structures are discussed. The first one is a C shaped slot which is realized along the feed line side of the OMSIW structure. Initial sections discuss the structure in detail along with its simulated and fabricated results. Different variations of this slot have been studied and the simulated results have been presented. The later part of the chapter focuses on a triangular CSRR slot etched on the OMSIW. The design details of the structure have been discussed while providing the simulated and measured results.

6.1 Problem definition

One may question the need to etch an extra slot on the OMSIW instead of tuning the resonance from the other open side wall to desired frequency band. When we compare the return losses of SIW, HMSIW, QMSIW and OMSIW, we can notice that as the size of the structure reduces, the gap between successive frequencies of resonance increases. In other

words, they drift apart from each other. This establishes a constraint over optimizing the frequencies for desired band of operation. In order to overcome this difficulty in obtaining a new resonant frequency band, we propose to etch a slot on the structure. It will be shown that introduction of the slot provides an extra degree of freedom by generating new frequency band.

6.2 C-slot on OMSIW

In the previous chapter we proposed a OMSIW resonator antenna that operates at 2.42 GHz. Retaining the design of that antenna, in this section, a compact OMSIW antenna is proposed by loading the structure with a C shaped slot. It was observed that the C slot adds an additional frequency of resonance at 5.8 GHz. Thus the proposed structure can be used for Wi-Fi and Bluetooth applications. The proposed antenna has been fabricated and measured. The measured results seem to be in good agreement with the simulated results.

With the objective of introducing another frequency band, we proposed to etch a slot on the structure and study its impact on the frequency response. The slots incise the surface current and arouse the electromagnetic fields [112]. We begin the design of the OMSIW with C slot by first implementing a slit along the radiating edge of the OMSIW structure proposed in the previous chapter. Figure 6.1 shows OMSIW with a slit. This slit provided a resonance at 6.2 GHz at the dominant resonant mode.

However, since we were interested in obtaining a resonance at 5.8 GHz, we proceeded to increase the electrical length of the slit. It is well known that as the electrical length increases the frequency of resonance decreases [128]. Etching a slit along the feed line side of the structure opened the scope for introducing a new frequency band. Since we were interested in tuning the resonant frequency to 5.8 GHz, we modified the slit by increasing its length and thus arriving at a C-shaped slot design. Thus, in physical sense, this translates into stronger currents from the open side wall OA being coupled to the slot. And since the electrical length was increased

on both ends the resonant frequency was lowered. All the dimensions of the structure have been provided in Table 6.1

Figure 6.1: OMSIW with a slit

6.2.1 Structure details

Introduction of a C shaped slot resulted in new resonant frequency at 5.9 GHz. Its dimensions were further tuned to obtain a center frequency of 5.8 GHz, thus making it useful for commercial applications. From the field distributions of OMSIW, we saw that the field is concentrated on the OA side wall. Hence the slot has been etched on the OA side instead of OX. Also, it was observed that the distance of the slot from the radiating edge mildly affects the dominant resonant frequency of the OMSIW structure. In our case, the distance of 2.2mm has been chosen to retain the original resonance at 2.40 GHz.

Parameter	Values (mm)
GL_{OMSIW}	60
GW_{OMSIW}	40
g_8	15
g_9	18.5
g_{10}	15.2
c_1	2.2
c_2	1.1
c_3	10.1
f_5	5.9
m_1	1.8
m_2	2.9
m_3	2.35
m_4	1.79
m_5	22.38
m_7	22.20

Table 6.1: Dimensions of OMSIW with a slit

The C shape slot has been etched on the top surface of the OMSIW antenna. We use the microstrip line feeding which matches the 50Ω impedance. The OMSIW structure has two open ends, through which it can be fed. Simulated results indicate that the location or position of the feed on the other open end doesn't affect the resonant frequency and the field distribution significantly. For the proposed structure, the feed has been retained from its OMSIW design. The inset feed has the advantage of better efficiency and broad BW. However, the width of the feed had to be modified to achieve a 50Ω impedance matching. Simulation results indicated an optimum feed width of $f_5=5.9\text{mm}$ designed as shown in the Figure 6.2. The structure has been implemented on the FR4 substrate with thickness of 3.3mm. The via hole diameter is 1.4mm and the spacing between them is given by 2.8mm.

The C shaped slot has been etched only on the top surface of the patch and not on the ground plane. The slot has not been implemented on the ground plane since it contributes to more noise and increases the radiation losses. There will be a huge dip in the peak gain values at higher frequencies. However if the radiation is desired in two directions, then the slot can be etched on the ground plane too.

Figure 6.2: OMSIW with Cslot layout

Parameter	Values (mm)
GL_{OMSIW}	60
GW_{OMSIW}	40
g_8	15
g_9	18.5
g_{10}	15.2
c_1	2.2
c_2	1.1
c_3	10.1
c_4	4.9
f_5	5.9
m_1	1.8
m_2	2.9
m_3	2.35
m_4	1.79
m_5	22.38
m_7	22.20

Table 6.2: Dimensions of OMSIW with C shaped slot

If we look at the OMSIW structure we find that there are two open side walls and the one

side with via holes. The two open side walls act as radiating edges. By introducing a C-shaped slot along one side wall, we allow the current to be coupled to the slot. The dimensions of the structure are as shown in the Table 6.2.

Figure 6.3 shows the current distribution of the proposed structure. Figure 6.3 (a) shows the current distribution at 2.4 GHz which happens to be from OA side of the OMSIW structure. Figure 6.3 (b) depicts the current distribution at 5.8 GHz. It can be seen that the presence of the slot aids in the current being coupled and thus resulting in additional radiating region.

Figure 6.3: Current distribution of the proposed C-slot resonator antenna at 2.4 GHz (a) and 5.8 GHz (b).

It was found that the ground dimensions affect the gain and the back lobe radiation. It has been optimized for 40mmx60mm for optimum performance.

6.2.2 Results

The proposed structure has been fabricated and the prototype is as shown in the Figure 6.4. This has been implemented on a FR-4 substrate with thickness of 3.2mm.

Figure 6.4: Fabricated prototype of proposed OMSIW structure with C-slot

The full wave simulated and the measured scattering parameter is as shown in Figure 6.5 which represent the frequency response. It can be seen that the simulated and the measured responses are almost in agreement. The S_{11} result indicate two resonances one at 2.4 GHz and another at 5.8 GHz. The 10-dB impedance BW of these resonances are quite narrow for the proposed structure. However they meet the requirement as specified for the Wifi and Bluetooth applications.

It can also be seen from the S-parameter result that the dominant mode resonance occurs at 2.39 GHz with a 10-dB impedance BW of 70 MHz. The simulated resonance is at the centre frequency of 2.398 GHz, while the measured frequency of resonance is at 2.42 GHz for TE_{101} mode. This discrepancy of 0.03 GHz could be owing to fabrication errors. Also, the 10-dB bandwidth for the simulated structure is found between frequency range 2.36-2.43 GHz, thus resulting in the total 10-dB BW of 2.52% while the measured 10-dB BW is found to be 2.24%.

Figure 6.5: Return loss of OMSIW C-slot

The return loss of the antenna has been measured using the Agilent N5230A PNA-L network analyser. The return loss is slightly higher and there is a clear stop band between these two frequencies of operation. The measured resonant frequencies have been slightly shifted upwards due to the effects of manufacturing tolerances.

The radiation characteristics are simulated and measured to evaluate the proposed structures applicability to antennas. The radiation pattern has been measured at typical frequencies based on the resonant modes. The simulated and the measured copolarized and cross polarized radiation pattern have been presented in Figure 6.6 and Figure 6.7 for 2.4 GHz. They seem to be in good agreement.

Figure 6.6: Copolarized radiation pattern of OMSIW resonator antenna at 2.4 GHz

Figure 6.7: Crosspolarized radiation pattern of OMSIW resonator antenna at 2.4 GHz

The co-polarized radiation pattern at 2.39 GHz has simulated gain of 6.56 dBi and measured gain of 4.46dBi. The cross polarized radiation pattern has simulated gain of -4.8dBi and measured gain of -7.2dBi. The main lobe radiates at a direction of $\theta = 5^\circ$ to the broadside direction at TE_{101} mode.

Figure 6.8 shows the co-polarized and cross polarized radiation pattern at 5.8 GHz with simulated gain of 7.83dBi and measured gain of 5.48dBi. The cross polarized radiation pattern is given by Figure 6.9 with simulated gain of -6.8dBi and measured gain of -10.2dBi. The main lobe radiates at a direction of $\theta = 15^\circ$ to the broadside direction. The orientation of theta and phi has been shown in Figure 6.4.

Figure 6.8: Copolarized radiation pattern of OMSIW resonator antenna at 5.8 GHz

Figure 6.9: Crosspolarized radiation pattern of OMSIW resonator antenna at 5.8 GHz

Figure 6.10: Gain of OMSIW C-slot resonator antenna

Figure 6.11: Efficiency of OMSIW C-slot resonator antenna

Figure 6.10 shows the simulated and measured gain values at the corresponding radiating regions. The simulated gain is 6.56 dBi at 2.4 GHz and measured gain of 4.45dBi at 2.4 GHz for the proposed linearly polarized resonator antenna at TE_{101} mode. At 5.8 GHz the simulated gain of the proposed antenna is 7.83dBi and measured gain of 5.48dBi. Efficiency is plotted in Figure 6.11 and it shows a simulated efficiency of 83.37% and measured efficiency of 74.87% for 2.4 GHz and simulated efficiency of 88.57% and measured efficiency of 79.48% for 5.8 GHz. This dip in the measured values of gain and efficiency in comparison to simulated values is due to different losses.

6.2.3 Comparative studies

Figure 6.12 shows the return loss of OMSIW structure with and without slot. It can be clearly seen that the C slot has introduced a new frequency band while still retaining the

original resonance at 2.4 GHz intact. The 10-dB BW at TE_{101} is almost the same.

Figure 6.12: Return loss of OMSIW structure with and without C-slot

6.3 Impact of C-slot orientation

In order to study the impact of coupling between the open side wall and the slot, the C slot was rotated at different angles and simulations were carried out for variation. Figure 6.13 shows the simulated layout of the structure for different angles of C slot. It is ensured that for all angles of rotation, the impedance is matched to 50Ω .

Figure 6.13: Various rotations of C shaped slot on the OMSIW structure

Figure 6.14 and Table 6.3 describe the S_{11} parameters for different angles of rotation. It can be observed that the resonant frequencies undergo an upward shift as the slot is rotated until 180° , since there is a variation in the coupling of the currents as frequency changes.

θ (deg)	Frequency (GHz)
0	5.81
30	6.06
90	6.72
180	5.06
270	5.34

 Table 6.3: Simulated resonant frequency variation for various θ deg

Figure 6.14: Simulated return loss of OMSIW with C-slot rotations

Also, the co polarized patterns remain almost the same as seen in Figure 6.15 and Figure 6.16 for the particular case of $\theta=180^\circ$. However the cross polarized pattern experiences a difference since the orientation of the C slot affects the polarization. Thus it can be concluded that the rotation of the slot on the OMSIW structure affects the resonating frequency while retaining the co polarized radiation performance.

Figure 6.15: Copolarized radiation pattern of OMSIW C-slot resonator antenna at $\theta=180\text{deg}$

Figure 6.16: Crosspolarized radiation pattern of OMSIW C-slot resonator antenna at $\theta=180\text{deg}$

6.4 Triangular CSRR slot on OMSIW

The purpose of introducing a complementary split ring resonator (CSRR) slot was to consider the impact of negative ϵ and negative μ on the OMSIW structure. We chose to implement a

triangular CSRR as a slot structure on the OMSIW. Previously, a square CSRR slot has been implemented in [121]. However, triangular CSRR have the advantage of being more compact and offer better BW in comparison to square CSRR. [129] shows that triangular CSRR offer better values of gain and efficiency in comparison to square CSRR. Taking these aspects into consideration, we designed a triangular CSRR slot on the proposed structure. It might be noted that our proposed structure is compact, has better gain and operates at a usable frequency in comparison to reported previous works in the area.

6.4.1 Structure details

A triangular CSRR has been loaded on the top surface of the proposed OMSIW structure as shown in fig 6.17. In order to retain the integrity of the ground and to reduce the losses contributed by noise and radiation, additional CSRR has not been etched on the ground plane.

The design parameters for the triangular CSRR include

- 1) Width of the ring
- 2) Gap between the split ring
- 3) Overall size of the SRR

It was observed that the width of the ring controls the resonant frequency. The triangular CSRR has the same dimensions for the gap between the SRR and the width of the rings so that it is easier for parameter control. The distance of the CSRR from the open side wall OA has been retained from the C slot structure to ensure maximum coupling of the currents. The dimensions of the structure are given in table 6.4.

Figure 6.17: Simulated layout of triangular CSRR slot OMSIW resonator antenna

Parameter	Values (mm)	Parameter	Values (mm)
GL_{OMSIW}	60	t_1	13.2
GW_{OMSIW}	40	t_2	0.6
g_8	15	t_3	0.7
g_9	18.5	t_4	0.9
g_{10}	15.2	t_5	1.4
c_1	2.2	t_6	1.2
f_6	6.1		
m_1	1.8		
m_2	2.9		
m_3	2.35		
m_4	1.79		
m_5	22.38		
m_7	22.4		

Table 6.4: Dimensions of OMSIW with triangular CSRR slot

Figure 6.18 shows the current distribution of the proposed structure at 1.8 GHz. The impact of the slot can be clearly seen from the field distribution results.

Figure 6.18: Current distribution of the proposed OMSIW resonator antenna with triangular CSRR at 1.8 GHz

6.4.2 Results

The proposed structure has been fabricated and the prototype is as shown in the Figure 6.19. This has been implemented on a FR4 substrate with thickness of 3.2mm.

The full wave simulated and the measured scattering parameter is as shown in Figure 6.20 which represents the frequency response. It can be seen that the simulated and the measured responses are almost in agreement. The S_{11} simulation result indicate two resonances one at 1.79 GHz and another at 2.66 GHz.

Figure 6.19: Fabricated prototype of proposed OMSIW structure with triangular CSRR slot

It can be seen from the S-parameter result that the first resonance occurs at 1.79 GHz with a 10-dB return loss BW of 80 MHz which is much below the cut off frequency of the OMSIW structure. The dominant mode resonance of the OMSIW resonator antenna gets shifted to 2.66 GHz from its original value of 2.42 GHz. The 10dB BW of this resonant antenna at 2.66 GHz is 130 MHz. The measured return loss values are found at 1.89 GHz and 2.71 GHz. This discrepancy could be owing to fabrication errors.

The return loss of the antenna has been measured using the Agilent N5230A PNA-L network analyser. The return loss is slightly higher and there is a clear stop band between these two frequencies of operation. The measured resonant frequencies have been slightly shifted upwards due to the effects of manufacturing tolerances.

Figure 6.20: Return loss OMSIW triangular CSRR slot

The radiation characteristics are simulated and measured to evaluate the proposed structures applicability to antennas. The radiation pattern has been measured at typical frequencies based on the resonant modes.

The simulated and the measured copolarized and cross polarized radiation pattern have been presented in Figure 6.21 and Figure 6.22 for the first frequency of resonance at 1.79 GHz. They seem to be in good agreement. The co-polarized radiation pattern has simulated gain of 8.1 dBi and measured gain of 6.8 dBi. The cross polarized radiation pattern has simulated gain of -2.98 dBi and measured gain of -9.4 dBi. The main lobe radiates at a direction of $\theta = 5^\circ$ to the broadside direction.

Figure 6.21: Copolarized radiation pattern of OMSIW resonator antenna at 1.8 GHz

Figure 6.22: Crosspolarized radiation pattern of OMSIW resonator antenna at 1.8 GHz

Figure 6.23 shows the co-polarized radiation pattern at 2.66 GHz with simulated gain of 6.42 dBi and measured gain of 5.68 dBi. The cross polarized radiation pattern is given by Figure 6.24 with simulated gain of -6.8dBi and measured gain of -7.2dBi. The main lobe radiates at a direction of $\theta = 30^\circ$ to the broadside direction. The orientation of theta and phi has been shown in Figure 6.19.

Figure 6.23: Copolarized radiation pattern of OMSIW resonator antenna at 2.8 GHz

Figure 6.24: Crosspolarized radiation pattern of OMSIW resonator antenna at 2.8 GHz

Figure 6.25 shows the simulated and measured gain values at the corresponding radiating regions. The simulated gain is 8.1dBi and measured gain of 6.8dBi is found at 1.79 GHz. At 2.66 GHz for the proposed linearly polarized resonator antenna at TE_{101} mode, the simulated gain is 6.42dBi and measured value is 5.68dBi. Efficiency is plotted in Figure 6.26 and it shows a simulated efficiency of 88.63% and measured efficiency of 82.39% at 1.79 GHz. At 2.66 GHz, the simulated efficiency is 81.46% and measured value is 78.52%. This dip in the measured values of gain and efficiency in comparison to simulated values is due to different losses.

Figure 6.25: Gain of OMSIW triangular CSRR slot resonator antenna

Figure 6.26: Efficiency of OMSIW triangular CSRR slot resonator antenna

6.4.3 Comparitive studies

Inorder to highlight the advantages of the proposed triangular CSRR OMSIW resonator antenna, we have compared its performance characteristics with the structure proposed in [121]. Table 6.5 summarizes the comparison.

Parameter	Proposed structure [Triangular CSRR]	Reference structure [Square CSRR] [121]
Resonant frequency (GHz)	1.79	5.07
Measured 10-dB BW (%)	2.42	1.55
Measured Gain (dBi)	6.8	5.9
Measured efficiency(%)	82.4	80.85

Table 6.5: Comparison of performance characteristics of proposed strucuture with reference works

It can be seen that the proposed structure is better than the reported work in terms of gain

and BW. Since the substrate used is FR4, the efficiency is lower, but the proposed structure is much simpler and compact than the square CSRR resonator antenna.

6.5 Summary

This chapter discussed two types of slots etched on the octamode substrate integrated resonator antenna. It was shown that the slots introduce a new frequency band and thus provide an extra degree of freedom to design antennas for specific frequency band. The proposed antennas were fabricated. The measured results were compared with the simulated ones and they seem to be in good agreement.

Chapter 7

Conclusion and Future Work

7.1 Conclusion

In this thesis, we present a few antenna designs for wide/multi-band applications implemented on substrate integrated waveguide technology, while incorporating the concepts of composite right/ left handed metamaterials. Of the most importance in this work, following conclusions can be drawn from this thesis.

A thorough literature review has been carried out on metamaterials and their application to the field of microwaves. Two different types of metamaterials (resonant and non-resonant) along with their pros and cons have been reviewed. Non-resonant type of metamaterials also known as composite right left handed transmission lines have been studied extensively. Their implementation has been discussed. Two prominent ones are microstrip and substrate integrated waveguide (SIW). The SIW platform has been studied in detail to highlight the benefits it offers above other means of implementation. They include high quality factor, higher power handling capabilities, and so on.

Basic concepts of a numerical method called singular perturbation have been discussed in

detail. The flowchart explaining the design steps has been provided. The method has been used to analyze a double periodic LWA CRLH structure loaded with alternate capacitance and inductances. Numerical expressions have been obtained which aid in obtaining dispersion characteristics, propagation constant and radiation characteristics. The modulation of the permeability which has increased the degree of freedom of the analysis has been included. The radiation efficiency has been plotted as a function of antennas length to illustrate the effect of modulating indices on the antenna performance. It has been obtained at the three frequencies of interest. The results obtained from the analysis have been verified by applying it to a double periodic CRLH leaky wave structure. The theoretical and measured results are in good agreement.

A LWA implemented on a SIW platform based on CRLH TL concept was presented. Two antennas were proposed based on SIW and HMSIW technology. The CRLH SIW LWA had V shaped slots etched on its top surface. The SIW design aspects were discussed and its equivalent circuit is provided. The proposed structure was simulated using commercially available simulation tool CST MWS. The dispersion characteristics have been obtained and the LH and the RH frequency ranges are identified. The transition from LH to RH region occurs at around 33.8 GHz. The frequency responses and the radiation patterns are simulated and the beam angle scans from -32° in the LH region to 61° in the RH region. The structure has a wide bandwidth and is very compact. The structure has been fabricated and its frequency response has been measured. The initial measurements indicate good agreement.

Then A CRLH LWA was proposed on HMSIW platform with V shaped slots. This antenna is unbalanced in nature and provides scanning from backward to forward directions. Its performance is almost comparable with its CRLH SIW LWA counterpart. Performance comparisons have been made with similar CRLH SIW LWAs reported in literature. It is seen that the proposed structures have the advantages of higher frequency, wide bandwidth and compact

size.

The characteristics of a square SIW resonator and its subsequent miniaturization have been presented. By effectively drawing the fictitious cutting lines along the magnetic walls, the square SIW resonator has been made to operate at different modes. This concept has been extended to HMSIW, QMSIW and OMSIW models. Proceeding with the same idea we propose an octamode SIW resonator that operates at 2.4 GHz has been proposed. In addition to the frequency of operation, attempt has been made to study the behaviour of the frequency shift as the geometry of the structure gets halved with each mode. The comparison of the frequency modes indicate that as the size of the structure gets reduced, the fundamental frequency experiences a downward shift while the higher order modes shift upwards in the frequency range. The QMSIW structure has been simulated and fabricated. The measured results are found to be in good agreement with the simulated result. The proposed octamode SIW resonator has been implemented on a FR4 substrate. The simulated and the measured results demonstrate that the miniaturized structure has comparable performance characteristics with that of its QMSIW counterpart. The structure finds potential commercial applications at 2.4 GHz.

Two types of slots etched on the octamode substrate integrated resonator antenna were proposed. It was shown that the slots introduce a new frequency band and thus provide an extra degree of freedom to design antennas for specific frequency band. A C-shaped slot was implemented which introduced a new frequency of resonance at 5.8 GHz. This proposed antenna can be used for dual band commercial applications. The next design involved a triangular CSRR slot etched which introduced a new resonant frequency at 1.8 GHz below the original cutoff frequency of 2.4GHz of the OMSIW structure. Both the proposed designs were fabricated and the measured results were compared with the simulated ones and they seem to be in good agreement.

Thus, the work presented provides a theoretical, simulation and experimental validation of

the proposed concepts. These novel antenna design can find applications in high frequency Ka-band communication systems and at the ISM band, the proposed antenna designs suit zigbee, IEEE 802.11n(WiFi), wireless HDMI and bluetooth standard applications.

7.2 Scope for future work

This thesis has presented novel antennas designs. However, like any useful new idea, it offers opportunities for many future investigations on the same or similar structures. Some of these possible studies are suggested as follows:

- 1) Measurement of the radiation characteristics of the CRLH SIW and CRLH HMSIW with V slots LWAs. Owing to the limited measurement facilities at higher frequencies, it was not possible to measure the radiation pattern of the proposed designs.
- 2) The OMSIW with slot concept can be extended to obtain triple band resonance by varying the type of slot structure. Also, the BW of the resonances can be made wider with further tuning.
- 3) The perturbation analysis can be extended to two directional periodicity. This can be achieved by varying the parameters individually along two dimensions.
- 4) Metasurfaces can be introduced beneath the array of CRLH LWA antennas in order to improve the performance characteristics.

List of Publications

Journal Papers

- [1] **M. Mujumdar**, C. Jin, and A. Alphones, "Double periodic composite right/left handed transmission line based leaky wave antenna by singular perturbation method," *Progress In Electromagnetics Research*, vol. 132, 113128, 2012.
- [2] **Manisha M** and A Alphones, "Octamode substrate integrated resonator antenna at 2.4 GHz," *IEEE Antennas and Wireless Propagation Letters* (Accepted for publication).
- [3] **Manisha M** and A Alphones, "Eighth mode substrate integrated waveguide dual band resonator antennas," *IEEE transactions on Antennas and Propagation* (Submitted).
- [4] Cheng Jin, Arokiaswami Alphones and **Manisha M. Dhirendra**, "TM Mode Analysis in a Metamaterial Based Dielectric Waveguide," *Progress In Electromagnetics Research*, vol. 132, 113128, 2012.
- [5] Jin, C., **Mujumdar, M.** and Alphones, A., "Broadband beam scanning circularly-polarised leaky-wave antenna based on modified substrate integrated waveguide," *Electronics Letters*, vol.49, no.5, pp.316,318, February 28 2013.

Conference Papers

- [6] **Mujumdar, M.**, Alphones, A., Jin Cheng, Nasimuddin, "Compact leaky wave antenna with periodical slots on substrate integrated waveguide," *Antennas and Propagation (EuCAP), 2014 8th European Conference on* , vol., no., pp.766,770, 6-11 April 2014.
- [7] **Mujumdar, M.**, Cheng Jin, Alphones, A., "Leaky wave radiation from a Composite Right/Left Handed based coplanar waveguide," *Microwave Conference Proceedings (APMC), 2011 Asia-Pacific* , vol., no., pp.1873,1876, 5-8 Dec. 2011.
- [8] Alphones, A.; **Mujumdar, M.**; Cheng Jin, "Substrate integrated waveguide and its applications to leaky wave antennas," *Microwave Conference Proceedings (APMC), 2013 Asia-Pacific* , vol., no., pp.470,472, 5-8 Nov. 2013.
- [9] **Mujumdar M.**, Alphones, A., Nasimuddin, "Compact Leaky Wave Antenna with Periodical Slots on Half mode Substrate Integrated Waveguide," *Antennas and Propagation Society International Symposium (APSURSI)*, 2015. (Submitted)
- [10] **Mujumdar M.**, Alphones, A. and Jin Cheng," Composite right/left handed metamaterial based leaky wave antennas", *Meta -14, The 5th International Conference on Metamaterials, Photonic Crystals and Plasmonics*, 20-23 May, 2014.

References

- [1] G. Kumar and K. P. Ray, *Broadband microstrip antennas*. Artech House, Inc., 2003.
- [2] V. G. Veselago, “Electrodynamics of substances with simultaneously negative values of sigma and mu,” *Soviet Physics Uspekhi-Ussr*, vol. 10, p. 509, 1968.
- [3] D. R. Smith, J. J. Mock, A. F. Starr, and D. Schurig, “Gradient index metamaterials,” *Physical review E*, vol. 71, p. 036609, 2005.
- [4] J. B. Pendry, “Negative refraction makes a perfect lens,” *Phys. Rev. Lett.*, vol. 85, pp. 3966–3969, 2000.
- [5] C. Caloz and T. Itoh, “Transmission line approach of left-handed (LH) materials and microstrip implementation of an artificial LH transmission line,” *IEEE Trans. Antennas Propagat.*, vol. 52, pp. 1159–1166, 2004.
- [6] D. R. Smith, W. J. Padilla, D. C. Vier, S. C. Nemat-Nasser, and S. Schultz, “Composite medium with simultaneously negative permeability and permittivity,” *Phys. Rev. Lett.*, vol. 84, pp. 4184–4187, 2000.
- [7] R. W. Ziolkowski and E. Heyman, “Wave propagation in media having negative permittivity and permeability,” *Phys. Rev. E*, vol. 64, p. 056625, 2001.
- [8] R. Marques, F. Martin, and M. Sorolla, *Metamaterials with negative parameters: theory, design, and microwave applications*. New York: Wiley, 2008.
- [9] C. Caloz and T. Itoh, *Electromagnetic metamaterials transmission line theory and microwave applications*. New York: Wiley, 2006.
- [10] G. V. Eleftheriades, A. K. Iyer, and P. C. Kremer, “Planar negative refractive index media using periodically L-C loaded transmission lines,” *IEEE Trans. Microw. Theory Tech.*, vol. 50, pp. 2702–2712, 2002.
- [11] C. Caloz and T. Itoh, “Novel microwave devices and structures based on the transmission line approach of meta-materials,” in *Microwave Symposium Digest, IEEE MTT-S International*, vol. 1, 2003, pp. 195 – 198.
- [12] J. Hirokawa and M. Ando, “Single-layer feed waveguide consisting of posts for plane wave excitation in parallel plates,” *IEEE Trans. Antennas Propagat.*, vol. 46, pp. 625 – 630, 1998.

REFERENCES

- [13] D. M. Pozar., *Microwave engineering, third edition.* New York: Wiley, 2004.
- [14] A. Grbic and G. V. Eleftheriades, “A backward-wave antenna based on negative refractive index l-c networks,” in *Antennas and Propagation Society International Symposium, IEEE*, vol. 4, 2002, pp. 340 – 343.
- [15] C. Jin, A. Alphones, and M. Tsutsumi, “Leaky-wave characteristics from double periodic composite right-/left-handed transmission lines,” *Microwaves, Antennas Propagation, IET*, vol. 5, no. 12, pp. 1399–1407, 2011.
- [16] J. B. Pendry, A. J. Holden, W. J. Stewart, and I. Youngs, “Extremely low frequency plasmons in metallic mesostructures,” *Phys. Rev. Lett.*, vol. 76, pp. 4773–4776, 1996.
- [17] J. B. Pendry, A. J. Holden, D. J. Robbins, and W. J. Stewart, “Magnetism from conductors and enhanced nonlinear phenomena,” *IEEE Trans. Microw. Theory Tech.*, vol. 47, pp. 2075–2084, 1999.
- [18] D. R. Smith, D. C. Vier, N. Kroll, and S. Schultz, “Direct calculation of permeability and permittivity for a left-handed metamaterial,” *Appl. Phys. Lett.*, vol. 77, no. 14, pp. 2246–2248, 2000.
- [19] R. A. Shelby, D. R. Smith, and S. Schultz, “Experimental verification of a negative index of refraction,” *Science*, vol. 292, p. 77C79, 2001.
- [20] S. A. Ramakrishna and T. M. Grzegorzczuk, *Physics and applications of negative refractive index materials.* CRC Press, 2009.
- [21] T. J. Cui, D. R. Smith, and R. Liu, Eds., *Metamaterials: theory, design, and applications.* Berlin: Springer-Verlag, 2010.
- [22] D. Schurig, J. J. Mock, B. J. Justice, S. A. Cummer, J. B. Pendry, A. F. Starr, and D. R. Smith, “Metamaterial electromagnetic cloak at microwave frequencies,” *Science*, vol. 314, pp. 977–980, 2006.
- [23] R. E. Collin, *Field theory of guided waves.* New York: IEEE Press, 1990.
- [24] R. E. Collin, *Foundations for microwave engineering.* New York, London: McGraw-Hill, 1992.
- [25] H. Raether, *Excitation of plasmons and interband transitions by electrons.* Berlin: Springer-Verlag, 1980.
- [26] M. S. Sodha and N. C. Shrivastava, *Microwave propagation in ferrimagnetics.* New York: Plenum Press, 1981.
- [27] L. D. Landau, E. M. Lifshitz, and L. P. Pitaevskii, *Electrodynamics of continuous media.* New York: Pergamon, 1984.
- [28] G. V. Eleftheriades and K. G. Balmain, *Negative-refraction metamaterials: fundamental principles and applications.* New York: Wiley, 2005.

REFERENCES

- [29] J. Valentine, S. Zhang, T. Zentgraf, E. Ulin-Avila, D. A. Genov, G. Bartal, and X. Zhang, "Three-dimensional optical metamaterial with a negative refractive index," *Nature*, vol. 455, no. 7211, pp. 376–U32, SEP 18 2008.
- [30] S. H. Lee, C. M. Park, Y. M. Seo, and C. K. Kim, "Reversed Doppler effect in double negative metamaterials," *Phys. Rev. B*, vol. 81, no. 24, p. 241102, JUN 14 2010.
- [31] J. Lu, T. M. Grzegorzczak, Y. Zhang, J. Pacheco, B. I. Wu, J. A. Kong, and M. Chen, "Cerenkov radiation in materials with negative permittivity and permeability," *Optics Express*, vol. 11, pp. 723–734, 2003.
- [32] P. Kolinko and D. R. Smith, "Numerical study of electromagnetic waves interacting with negative index materials," *Optics Express*, vol. 11, no. 7, pp. 640–648, APR 7 2003.
- [33] J. B. Pendry, D. Schurig, and D. R. Smith, "Controlling electromagnetic fields," *Science*, vol. 312, p. 1780C1782, 2006.
- [34] A. A. Houck, J. B. Brock, and I. J. Chuang, "Experimental observations of a left-handed material that obeys snell's law," *Phys. Rev. Lett.*, vol. 90, 2003.
- [35] R. Liu, C. Ji, J. J. Mock, J. Y. Chin, T. J. Cui, and D. R. Smith, "Broadband ground-plane cloak," *Science*, vol. 323, p. 366, 2009.
- [36] H. F. M. and T. J. Cui, "Three-dimensional broadband ground-plane cloak made of metamaterials," *Nature Commun.* 1, vol. 23, 2010.
- [37] R. A. Shelby, D. R. Smith, S. C. Nemat-Nasser, and S. Schultz, "Microwave transmission through a two-dimensional, isotropic, left-handed metamaterial," *Appl. Phys. Lett.*, vol. 78, no. 4, pp. 489–491, 2001.
- [38] P. M. Valanju, R. M. Walser, and A. P. Valanju, "Wave refraction in negative-index media: Always positive and very inhomogeneous," *Phys. Rev. Lett.*, vol. 88, p. 187401, 2002.
- [39] V. G. Veselago, "Formulating fermat's principle for light traveling in negative refraction materials," *PhysicsCUSpekhi*, vol. 45, p. 1097C1099, 2002.
- [40] R. W. Ziolkowski, "Pulsed and CW Gaussian beam interactions with double negative metamaterial slabs," *Optics Express*, vol. 11, pp. 662–681, 2003.
- [41] A. Alu and N. Engheta, "Guided modes in a waveguide filled with a pair of single-negative (sng), double-negative (dng), and/or double-positive (dps) layers," *IEEE Trans. Microw. Theory Tech.*, vol. 52, pp. 199 – 210, 2004.
- [42] A. Lakhtakia, "Positive and negative goos-h anchen shifts and negative phase-velocity mediums (alias left-handed materials)," *Int. J. Electron. Commun.*, vol. 58, pp. 229–231, 2004.
- [43] R. Marques, F. Mesa, J. Martel, and F. Medina, "Comparative analysis of edge- and broadside-coupled split ring resonators for metamaterial design - Theory and experiments," *IEEE Trans. Antennas Propagat.*, vol. 51, pp. 2572–2581, 2003.

REFERENCES

- [44] F. Martin, J. Bonache, F. Falcone, M. Sorolla, and R. Marques, "Split ring resonator-based left-handed coplanar waveguide," *Appl. Phys. Lett.*, vol. 83, no. 22, pp. 4652–4654, 2003.
- [45] J. D. Baena, R. Marques, F. Medina, and J. Martel, "Artificial magnetic metamaterial design by using spiral resonators," *Phys. Rev. B*, vol. 69, 2004.
- [46] F. Falcone, F. Martin, J. Bonache, R. Marques, T. Lopetegi, and M. Sorolla, "Left handed coplanar waveguide band pass filters based on bi-layer split ring resonators," *IEEE Microw. and Wireless Compon. Lett.*, vol. 14, pp. 10–12, 2004.
- [47] F. Falcone, T. Lopetegi, J. D. Baena, R. Marques, F. Martin, and M. Sorolla, "Effective negative- ϵ stopband microstrip lines based on complementary split ring resonators," *IEEE Microw. and Wireless Compon. Lett.*, vol. 14, pp. 280 – 282, 2004.
- [48] J. Bonache, I. Gil, J. Garcia-Garcia, and E. Martin, "Complementary split ring resonators for microstrip diplexer design," *Electron. Lett.*, vol. 41, no. 14, pp. 810–811, 2005.
- [49] J. Bonache, I. Gil, J. Garcia-Garcia, and F. Martin, "Novel microstrip bandpass filters based on complementary split-ring resonators," *IEEE Trans. Microw. Theory Tech.*, vol. 54, pp. 265–271, 2006.
- [50] M. Gil, J. Bonache, I. Gil, J. Garcia-Garcia, and F. Martin, "On the transmission properties of left-handed microstrip lines implemented by complementary split rings resonators," *International Journal of Numerical Modelling: Electronic Networks, Devices and Fields*, vol. 19, pp. 87–103, 2006.
- [51] I. Arnedo, J. Illescas, M. Flores, T. Lopetegi, M. A. G. Laso, F. Falcone, J. Bonache, J. Garcia-Garcia, F. Martin, J. A. Marcotegui, R. Marques, and M. Sorolla, "Forward and backward leaky wave radiation in split-ring-resonator-based metamaterials," *IET Microw., Antennas Propag.*, vol. 1, pp. 65–68, 2007.
- [52] M. Gil, J. Bonache, and F. Martin, "Metamaterial filters: a review," *Metamaterials*, vol. 2, pp. 186–197, 2008.
- [53] J. C. Sureau, "Propagation constants of leaky-wave antenna for near end-fire radiation," *IEEE Trans. Antennas Propagat.*, vol. AP15, pp. 308–309, 1967.
- [54] A. A. Oliner, "A periodic-structure negative-refractive-index medium without resonant elements," in *URSI Digest, IEEE-AP-S USNC/URSI National Radio Science Meeting*, San Antonio, TX, 2002, p. 41.
- [55] C. Caloz, H. Okabe, T. Iwai, and T. Itoh, "Anisotropic pbg surface and its transmission line model," in *URSI Digest, IEEE-AP-S USNC/URSI National Radio Science Meeting*, San Antonio, TX, 2002, p. 224.
- [56] A. F. Harvey, "Periodic and guiding structures at microwave frequencies," *IRE (IEEE) Trans. Microw. Theory Tech.*, vol. 8, pp. 30 –61, 1960.
- [57] C. Caloz and T. Itoh, "A novel mixed conventional microstrip and composite right/left-handed backward-wave directional coupler with broadband and tight coupling characteristics," *IEEE Microw. and Wireless Compon. Lett.*, vol. 14, pp. 31–33, 2004.

REFERENCES

- [58] G. V. Eleftheriades and R. Islam, "Miniaturized microwave components and antennas using negative-refractive-index transmission-line (NRI-TL) metamaterials," *Metamaterials*, vol. 1, pp. 53–61, 2007.
- [59] S. Paulotto, P. Baccarelli, F. Frezza, and D. R. Jackson, "A novel technique for open-stopband suppression in 1-D periodic printed leaky-wave antennas," *IEEE Trans. Antennas Propagat.*, vol. 57, pp. 1894–1906, 2009.
- [60] M. Tsutsumi, "Negative refractive index transmission media and its applications to the microwave circuits," *Journal of IEICE*, vol. 88, pp. 23–27, 2005.
- [61] M. Tsutsumi, C. Jin, and A. Alphones, "Leaky wave phenomenon from double periodic left handed waveguide," in *Asia-Pacific Microwave Conf.*, Singapore, pp. 1238 - 1241, 2009, pp. 1238 – 1241.
- [62] C. Jin, A. Alphones, and L. C. Ong, "Broadband leaky-wave antenna based on composite right/left handed substrate integrated waveguide," *Electron. Lett.*, vol. 46, no. 24, pp. 1584 –1585, 2010.
- [63] B. Y. Tong, "Electronic structure of one dimensional binary alloys," *Phys. Rev. Lett.*, vol. 175, pp. 710–722, 1968.
- [64] J. R. Pierce and N. Wax, "A note on filter-type traveling-wave amplifiers," in *the IRE*, vol. 37, pp. 622-625, 1949.
- [65] E. Bloch, "Über die quantenmechanik derlektronen in kristallgittern," *Z. Phys*, vol. 52, pp. 555–600, 1928.
- [66] S. R. Barone and M. A. Narcowich, "Floquet theory and applications," *Phys. Rev. A*, vol. 15, pp. 1109–1125, 1977.
- [67] C. Kittel, *Introduction to solid state physics*. New York: Wiley, 2005.
- [68] C. Elachi, "Waves in active and passive periodic structures: A review," in *Proceedings of the IEEE*, vol. 64, 1976, pp. 1666– 1698.
- [69] L. Brillouin, *Wave propagation in periodic structures; Electric Filters and Crystal Lattices*. New York, London: McGraw-Hill, 1946.
- [70] A. A. Oliner and D. R. Jackson, *Antenna engineering handbook, 4th edition*, J. L. Volakis, Ed. McGraw-Hill, 2007.
- [71] S. Paulotto, P. Baccarelli, F. Frezza, and D. R. Jackson, "Full-wave modal dispersion analysis and broadside optimization for a class of microstrip CRLH Leaky-wave antennas," *IEEE Trans. Microw. Theory Tech.*, vol. 56, pp. 2826–2837, 2008.
- [72] W. Hong, T. L. Chen, C. Y. Chang, J. W. Sheen, and Y. D. Lin, "Broadband tapered microstrip leaky-wave antenna," *IEEE Trans. Antennas Propagat.*, vol. 51, pp. 1922 – 1928, 2003.
- [73] L. Liu, C. Caloz, and T. Itoh, "Dominant mode (DM) leaky-wave antenna with backfire-to-endfire scanning capability," *Electron. Lett.*, vol. 38, no. 23, pp. 1414–1416, 2000.

REFERENCES

- [74] W. Menzel, "A new travelling wave antenna in microstrip," in *European Microwave Conference (EuMC)*, 1978, pp. 302–306.
- [75] K. C. Gupta, R. Garh, and I. Bahl, *Microstrip lines and slotlines*. Artech House, Inc., 1996.
- [76] A. Sanada, C. Caloz, and T. Itoh, "Zeroth order resonance in composite right/left-handed transmission line resonators," in *Asia-Pacific Microwave Conf.*, vol. 3, 2003, p. 1588C1592.
- [77] S. Lim, C. Caloz, and T. Itoh, "Electronically scanned composite right/left handed microstrip leaky-wave antenna," *IEEE Microw. and Wireless Compon. Lett.*, vol. 14, pp. 277–279, 2004.
- [78] R. Mongia, I. Bahl, and P. Bhartia, *RF and microwave coupled-line circuits*. Artech House, Inc., 1999.
- [79] H. Okabe, C. Caloz, and T. Itoh, "A compact enhanced-bandwidth hybrid ring using an artificial lumped-element left-handed transmission-line section," *IEEE Trans. Microw. Theory Tech.*, vol. 52, pp. 798–804, 2003.
- [80] W. H. Weng, L. S. Jang, and W. Y. Chen, "A sierpinski-based resonator applied for low loss and miniaturized bandpass filters," *Microwave Opt. Technol. Lett.*, vol. 51, pp. 411–413, 2009.
- [81] Y. J. Sung, "Miniaturized defected ground structure using lumped capacitor," *Microwave Opt. Technol. Lett.*, vol. 51, pp. 802–804, 2009.
- [82] Y. Horii, C. Caloz, and T. Itoh, "Super-compact multilayered left-handed transmission line and diplexer application," *IEEE Trans. Microw. Theory Tech.*, vol. 53, pp. 1527–1534, 2005, part 2.
- [83] J. Bonache, G. Posada, G. Carchon, W. D. Raedt, and E. Martin, "Compact ($<0.5\text{mm}^2$) k-band metamaterial bandpass filter in mcm-d technology," *Electron. Lett.*, vol. 43, no. 5, pp. 288–290, 2007.
- [84] F. P. Casares-Miranda, E. Marquez-Segura, P. Otero, and C. Camacho-Penalosa, "Composite right/left-handed transmission line with wire bonded interdigital capacitor," *IEEE Microw. and Wireless Compon. Lett.*, vol. 16, pp. 624–626, 2006.
- [85] H. Liu, R. Knoechel, and K. Schuenemann, "Miniaturized bandstop filter using meander spurline and capacitively loaded stubs," *ETRI Journal*, vol. 29, pp. 614–618, 2007.
- [86] A. W. Lines, G. R. Nicoll, and A. M. Woodward, "Some properties of waveguides with periodic structure," *Journal of IEE*, vol. 97, pp. 207–208, 1950.
- [87] D. Deslandes and K. Wu, "Single-substrate integration technique of planar circuits and waveguide filters," *IEEE Trans. Microw. Theory Tech.*, vol. 51, pp. 593–596, 2003.
- [88] J. X. Chen, W. Hong, Z. C. Hao, H. Li, and K. Wu, "Development of a low cost microwave mixer using a broad-band substrate integrated waveguide (siw) coupler," *IEEE Microw. and Wireless Compon. Lett.*, vol. 16, pp. 84–86, 2006.

REFERENCES

- [89] W. D'Orazio and K. Wu, "Substrate-integrated-waveguide circulators suitable for millimeter-wave integration," *IEEE Trans. Microw. Theory Tech.*, vol. 54, pp. 3675 – 3680, 2006.
- [90] F. Xu, K. Wu, and X. Zhang, "Periodic leaky-wave antenna for millimeter wave applications based on substrate integrated waveguide," *IEEE Trans. Antennas Propagat.*, vol. 58, no. 2, pp. 340 –347, Feb. 2010.
- [91] F. Xu and K. Wu, "Guided-wave and leakage characteristics of substrate integrated waveguide," *IEEE Trans. Microw. Theory Tech.*, vol. 53, pp. 66 – 73, 2005.
- [92] Y. Weitsch and T. F. Eibert, "A left-handed/right-handed leaky-wave antenna derived from slotted rectangular hollow waveguide," in *European Microwave Conference (EuMC)*, Munich, Germany, 2007, pp. 917 –920.
- [93] S. Lim, C. Caloz, and T. Itoh, "Metamaterial-based electronically controlled transmission-line structure as a novel leaky-wave antenna with tunable radiation angle and beamwidth," *IEEE Trans. Microw. Theory Tech.*, vol. 53, pp. 161–173, 2005.
- [94] H. Zhao, T. J. Cui, X. Q. Lin, and H. F. Ma, "The Study of Composite Right/Left Handed Structure in Substrate Integrated Waveguide," *2006 International Symposium on Biophotonics, Nanophotonics and Metamaterials*, pp. 547–549, Oct. 2006.
- [95] Y. Dong and T. Itoh, "Composite right/left-handed substrate integrated waveguide and half mode substrate integrated waveguide leaky-wave structures," *IEEE Trans. Antennas Propag.*, vol. 59, pp. 767 –775, 2011.
- [96] F. Aznar, M. Gil, J. Bonache, and F. Martin, "Revising the equivalent circuit models of resonant-type metamaterial transmission lines," in *Microwave Symposium Digest, IEEE MTT-S International*, Atlanta, GA, 2008, pp. 322–325.
- [97] M. H. Holmes, *Introduction to perturbation methods*. Berlin: Springer-Verlag, 1995.
- [98] A. Alphones and M. Tsutsumi, "Leaky-wave radiation from a periodically photoexcited semiconductor slab waveguide," *IEEE Trans. Microw. Theory Tech.*, vol. 43, pp. 2435–2441, 1995, part 2.
- [99] K. C.-J. L. Leong and T. Itoh, "Design of resonant small antenna using composite right/left-handed transmission line," in *Antennas and Propagation Society International Symposium, IEEE*, vol. 2B, 2005, pp. 218–221.
- [100] M. Rehman, Z. Baharudin, M. Zakariya, M. Khir, and M. Khan, "Microwave bandpass filter using qmsiw," in *RF and Microwave Conference (RFM), 2013 IEEE International*, Dec 2013, pp. 172–175.
- [101] C. Jin, R. Li, A. Alphones, and X. Bao, "Quarter-Mode Substrate Integrated Waveguide and Its Application to Antennas Design," *IEEE Transactions on Antennas and Propagation*, vol. 61, no. 6, pp. 2921–2928, Jun. 2013.
- [102] Z. Zhang, N. Yang, and K. Wu, "5-GHz bandpass filter demonstration using quarter-mode substrate integrated waveguide cavity for wireless systems," *2009 IEEE Radio and Wireless Symposium*, pp. 95–98, Jan. 2009.

REFERENCES

- [103] Z. Guo, K.-S. Chin, W. Che, and C.-C. Chang, "Cross-coupled bandpass filters using QM-SIW cavities and S-shaped slot coupling structures," *Journal of Electromagnetic Waves and Applications*, vol. 27, no. 2, pp. 160–167, Jan. 2013.
- [104] C. Jin, R. Li, and A. Alphones, "Conical beam windmill -shaped array antenna design based on quarter mode substrate integrated waveguide," *Progress In Electromagnetics Research Letters*, vol. 39, no. February, pp. 1–8, 2013.
- [105] Q. Lai and F. Christophe, "Characterization of the Propagation Properties of the Half-Mode Substrate Integrated Waveguide," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 57, no. 8, pp. 1996–2004, 2009.
- [106] Q. H. Lai, K. Z. Q. Hong, Wei, Y. S. Zhang, and K. Wu, "Half-Mode Substrate Integrated Waveguide Transverse Slot Array Antennas," *Antennas and Propagation, IEEE Transactions on*, vol. 57, no. 4, pp. 1064–1072, 2009.
- [107] S. Razavi and M. Neshati, "Development of a low-profile circularly polarized cavity-backed antenna using hmsiw technique," *Antennas and Propagation, IEEE Transactions on*, vol. 61, no. 3, pp. 1041–1047, March 2013.
- [108] T. Kaufmann and C. Fumeaux, "Wearable textile half-mode substrate-integrated cavity antenna using embroidered vias," *Antennas and Wireless Propagation Letters, IEEE*, vol. 12, pp. 805–808, 2013.
- [109] S. Agneessens and H. Rogier, "Compact half diamond dual-band textile hmsiw on-body antenna," *Antennas and Propagation, IEEE Transactions on*, vol. 62, no. 5, pp. 2374–2381, May 2014.
- [110] S. Peng, T. Tamir, and H. L. Bertoni, "Theory of periodic dielect waveguides," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 23, no. 1, pp. 123–133, Jan 1975.
- [111] M. Tsutsumi, J. Cheng, and A. Alphones, "Leaky wave phenomenon from double periodic left handed waveguide," in *Microwave Conference, 2009. APMC 2009. Asia Pacific*, Dec 2009, pp. 1238–1241.
- [112] N. Marcuvitz, *Waveguide handbook*. P. Peregrinus on behalf of the Institution of Electrical Engineers, London, UK, 1986.
- [113] Y.-C. Chen, C.-K. Wu, and C.-K. Tzuang, "Dual-frequency electric-magnetic-electric microstrip leaky-mode antenna of a single fan beam," *IEEE Trans. Microw. Theory Tech.*, vol. 50, pp. 2713 – 2720, 2002.
- [114] A. K. Iyer and G. V. Eleftheriades, "Negative refractive index metamaterials supporting 2-D waves," in *Microwave Symposium Digest, IEEE MTT-S International*, Hamilton, R, Ed., 2002, pp. 1067–1070.
- [115] L. Liu, C. Caloz, and T. Itoh, "Dominant mode leaky-wave antenna with backfire-to-endfire scanning capability," *Electron. Lett.*, vol. 38, no. 23, pp. 1414 – 1416, 2002.
- [116] C. Caloz, C.-C. Chang, and T. Itoh, "Full-wave verification of the fundamental properties of left-handed materials in waveguide configurations," *J. of Appl. Phys.*, vol. 90, pp. 5483 – 5486, 2001.

REFERENCES

- [117] R. Elliott and L. Kurtz, "The design of small slot arrays," *Antennas and Propagation, IEEE Transactions on*, vol. 26, no. 2, pp. 214–219, Mar 1978.
- [118] D. Deslandes and K. Wu, "Integrated microstrip and rectangular waveguide in planar form," *Microwave and Wireless Components Letters, IEEE*, vol. 11, no. 2, pp. 68–70, Feb 2001.
- [119] J. Macha and M. Polivka, "A dual band crlh substrate integrated waveguide leaky wave antenna," in *Microwave Conference (EuMC), 2011 41st European*, Oct 2011, pp. 535–538.
- [120] W. Hong, B. Liu, Y. Wang, Q. Lai, H. Tang, X. X. Yin, Y. D. Dong, Y. Zhang, and K. Wu, "Half mode substrate integrated waveguide: A new guided wave structure for microwave and millimeter wave application," in *Infrared Millimeter Waves and 14th International Conference on Terahertz Electronics, 2006. IRMMW-THz 2006. Joint 31st International Conference on*, Sept 2006, pp. 219–219.
- [121] S. Sam and S. Lim, "Electrically small eighth-mode substrate-integrated waveguide (em-siw) antenna with different resonant frequencies depending on rotation of complementary split ring resonator," *Antennas and Propagation, IEEE Transactions on*, vol. 61, no. 10, pp. 4933–4939, Oct 2013.
- [122] L. Hongwei, L. Joy, H. Mike, and P. Ram, "A novel de-embedding technique for millimeter-wave package characterization," in *ARFTG Conference Digest-Spring, 54th*, vol. 36, dec. 1999, pp. 1–8.
- [123] S. Sercu and L. Martens, "Accurate de-embedding of the contribution of the test boards to the high-frequency characteristics of backplane connectors," in *IEEE 6th Topical Meeting on Electrical Performance of Electronic Packaging*, 1997, pp. 177–180.
- [124] M. Ferndahl, C. Fager, K. Andersson, P. Linner, H.-O. Vicks, and H. Zirath, "A general statistical equivalent-circuit-based de-embedding procedure for high-frequency measurements," *IEEE Trans. Microw. Theory Tech.*, vol. 56, pp. 2692–2700, 2008.
- [125] H. K. Smith and P. E. Mayes, "Log-periodic array of dual-feed microstrip patch antennas," *IEEE Trans. Antennas Propagat.*, vol. 39, pp. 1659–1664, 1991.
- [126] M. Schuhler, R. Wansch, and M. A. Hein, "On strongly truncated leaky-wave antennas based on periodically loaded transmission lines," *IEEE Trans. Antennas Propag.*, vol. 58, pp. 3505–3514, 2010.
- [127] R. Garg, *Microstrip Antenna Design Handbook*, ser. Antennas and Propagation Library. Artech House, 2001.
- [128] L. Brillouin, *Wave propagation in periodic structures*. Dover Publications, Inc., 1953.
- [129] M. Vidyalakshmi, B. Rekha, and P. Rao, "Stopband characteristics of complementary triangular split ring resonator loaded microstrip line," in *Applied Electromagnetics Conference (AEMC), 2011 IEEE*, Dec 2011, pp. 1–4.
- [130] R. E. Collin and F. J. Zucker, *Antenna Theory*, part 2 ed. New York, London: McGraw-Hill, 1969.

REFERENCES

- [131] A. A. Oliner, S. T. Peng, T. Hsu, and A. Sanchez, "Guidance and leakage properties of a class of open dielectric waveguides: part ii—new physical effects," *IEEE Trans. Microw. Theory Tech.*, vol. 29, pp. 855 – 869, 1981.
- [132] S. T. Peng and A. A. Oliner, "Guidance and leakage properties of a class of open dielectric waveguides: part i—mathematical formulations," *IEEE Trans. Microw. Theory Tech.*, vol. 29, pp. 843 – 855, 1981.
- [133] D. R. Jackson and N. Alexopoulos, "Gain enhancement methods for printed circuit antennas," *IEEE Trans. Antennas Propag.*, vol. 33, pp. 976 – 987, 1985.
- [134] W. S. Park and S. R. Seshadri, "Theory of the grating coupler for a grounded-dielectric-slab waveguide," *IET Microw., Antennas Propag.*, vol. 132, pp. 149–156, 1985.
- [135] D. M. Pozar and B. Kaufman, "Increasing the bandwidth of a microstrip antenna by proximity coupling," *Electron. Lett.*, vol. 23, no. 8, p. 369, 1987.
- [136] H. Yang and N. Alexopoulos, "Gain enhancement methods for printed circuit antennas through multiple superstrates," *IEEE Trans. Antennas Propag.*, vol. 35, pp. 860 – 863, 1987.
- [137] S. R. Seshadri, "Search results coupling of guided modes in thin films with surface corrugation," *J. of Appl. Phys.*, vol. 63, pp. R115–R146, 1988.
- [138] R. D. Meade, K. D. Brommer, A. M. Rappe, and J. D. Joannopoulos, "Electromagnetic bloch waves at the surface of a photonic crystal," *Phys. Rev. B*, vol. 44, pp. 10 961–10 964, 1991.
- [139] M. Kasraian, "Double-grating thin-film devices based on second-orderbragg interaction," *J. of Appl. Phys.*, vol. 75, pp. 7639–7652, 1994.
- [140] S. Ramo, J. R. Whinnery, and T. V. Duzer, *Fields and waves in communication electronics, third edition*. New York: Wiley, 1994.
- [141] J. D. Joannopoulos, R. D. Meade, and J. N. Winn, *Photonic crystals*. Princeton University Press, 1995.
- [142] C. Jin and A. Alphones, "Compact interdigital microstrip band pass filter," *Microwave Opt. Technol. Lett.*, vol. 52, pp. 2128–2132, 2010.
- [143] C. Jin, A. Alphones, and O. L. Chuen, "Leaky wave antenna based on composite right/left handed substrate integrated waveguide," in *Asia-Pacific Microwave Conf.*, Yokohama, Japan, 2010, pp. 1997–2000.
- [144] J.-Q. Huang and Q.-X. Chu, "Compact uwb band-pass filter utilizing modified composite right/left handed structure with cross coupling," *Progress In Electromagnetics Research*, vol. 107, pp. 179–186, 2010.