Study of periodic and optical controlled structures for microwave circuits

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Study of Periodic and Optical controlled Structures for Microwave Circuits

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Summary

Periodic structures have found wide applications in modern communication systems such as filters, power combiners and radar systems. Nowadays, periodic structures play an important role in wireless communication systems due to their attractive features such as compact size, ease of fabrication and high efficiency. However, with the rapid progress of technology, higher requirements have risen which require the devices have more compact size, lighter weight and multi-function to accommodate various services. The sizes and the weight of periodic structures can be reduced by applying novel split ring patterned photonic bandgap/electromagnetic (PBG/EBG) structures to one dimensional microstrip line. These bandgap structures realized on semiconductor material could be exploited for optically controlled behavior. These kinds of periodic structure and their applications are the main subjects of study in this thesis.

Periodic aperture patterns etched on the ground plane of a microstrip line have been extensively investigated by many researchers. However, recently few research works have been reported on split ring pattern photonic bandgap structures and optical controlled photonic bandgap structures. In this thesis, a thorough study has been conducted on the scattering and circuit perspectives, radiation efficiency

and mutual coupling of these periodic structures have been estimated.

Numerous kinds of PBG/EBG patterns have been introduced and investigated, including split ring pattern PBG/EBG structures. A microstrip line loaded with these PBG/EBG structures has been designed and the stop band characteristics of all these structures have been discussed. Especially for split ring resonant loaded microstrip line, the transmission loss is as high as 80 dB. Also the dimension of split ring resonator loaded microstrip line is much smaller than conventional square pattern PBG/EBG loaded microstrip lines, which greatly reduced the size and weight of the device. After this, the combined defected ground plane structure (DGS) and split ring resonator (SRR) structure has been discussed. By applying split ring resonator into DGS structures, the spurious response in the stop band is efficiently suppressed compared to those of the conventional DGS structures. The stopband characteristic of DGS-SRR structure is significantly improved with about 50% of 10dB bandwidth increase using this novel structure, and the dimension of the structure is 25% reduced compared to conventional one.

To combine the optical controlled semiconductor and PBG/EBG loaded microstrip line, the plasma-induced PBG/EBG loaded microstrip lines have been introduced and the performance of different kinds of PBG/EBG patterns has been investigated with rigorous theoretical treatments. The square pattern, ring pattern and split ring pattern PBG/EBG structures with and without plasma induced by laser illumination have been analyzed. Scattering parameters have been given out and tunability by optical control of the structure has also been described.

Perturbation method with multiple scales has been used to carry out a rigorous electromagnetic analysis on these optical controlled periodical structures. For the structure with fixed surface corrugation and variable dielectric corrugation, the characteristics of three-mode coupling have been investigated. The results have been drawn out for the reflection coefficients and the transmission coefficient of all three modes. The possibility of fabricating a tunable rejection filter enhanced by optical excitation has also been stated. Besides the three mode matching behavior, the leaky wave characteristics on this photo induced double grating silicon slab have been analyzed rigorously by singular perturbation method based on multiple space scales. The leakage coefficient and the radiation efficiency have been analyzed and computed numerically and these radiation characteristics have been discussed as a function of plasma density due to laser intensity. The radiation efficiency is enhanced by the optical excitation and this gives possibility to control the radiation behavior optically.

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List of Acronyms

ABBREVIATIONS

FULL EXPRESSIONS

1-D, 2-D, 3-D

One dimensional, two dimensional, three dimensional

ADS

Advanced design system

BW

Bandwidth

CPW

Coplanar waveguide

CSRR

Complimentary split ring resonator

dB

Decibel

DGS

Defected groundplane structure

DNG

Double negative

EBG

Electromagnetic banggap

ECM

Equivalent circuit model

EM

Electromagnetic

EMC

Electromagnetic compatibility

FDTD

Finite-difference time domain

FEM

Finite-element method

HFSS

High frequency structure simulator

LH

Left handed

MMIC	Monolithic microwave integrated circuit
MM	Mode-matching
MOM	Method of moments
NIR	Negative index of refraction
OEIC	optoelectronic integrated circuits
PBG	Photonic banggap
PCB	Printed circuit board
RF	Radio frequency
SRR	Split ring resonator
TE	Transverse electric
TM	Transverse magnetic
TEM	Transverse electric and magnetic
UWB	Ultra wide band

Chapter 1

Introduction

1.1 Motivation

Periodic structures are very common and useful structures at microwave frequencies, and photonic bandgap (PBG) structure is one of those kinds of periodic structures. They are multi-dimensions and/or multi-layered design with periodic structures that effectively prevent electromagnetic wave propagations in a certain band of frequencies. If the PBG structures are incorporated in the conventional microwave circuits, they can exhibit passband and stopband characteristics in the desired frequency domain. Waves are allowed to propagate during the passband, and get rejected in the stopband region. In electromagnetic (EM) applications, The term electromagnetic band gap (EBG) has been adapted by the microwave community to make a more general term than the term PBG. Due to the passband characteristic of EBG, it is often used as slow wave medium, which reduces the size of electronic circuits. And the wide stopband of EBG can be applied to suppress spurious transmission and leakage in guided structures [1]–[6].

1.1 Motivation

EBG structures are widely used in microwave devices and applications due to their passband and stopband characteristics. The EBG assisted transmission line can provide excellent improvement in transmission coefficient both in low and high frequencies with wider rejection bandwidth [6]–[8]. For antennas, with a shorted microstrip patch with an EBG ground plane showed a more than 3dB improvement in the gain and a significant improvement in the reduction of cross-polarization levels compared to conventional ground plane [9]. Also the EBG structure has found its potential application in amplifiers, filters, power combiners and power dividers, phased array antenna system, electromagnetic compatibility (EMC) measurement and integrated circuits. The structures have been widely applied to enhance the performance of microwave devices, by suppressing some of the spurious responses.

But most of the conventional EBG structures analyzed are fixed periodic structures and the resonant frequency is determined by the period of the EBG structure. This may cause the structure to be too large in certain frequency ranges and the sub-wavelength designs are recently getting more attentions in various applications.

Also dynamic tuning capabilities have not been attempted in EBG structures in early days, but in recent times electronic tuning has been attempted. Also the uniform illumination on the semiconductor slab for beam switching/scanning have been investigated in some of the previous reported works [10]. But for most of the EBG structures, they have certain passband and stopband and these passband and stopband characteristic cannot be changed. There are few research works focused

1.2 Objectives

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toward the on-off or tunable characteristic of device design with EBG structures. And for periodic structures there is not enough work exploiting rigorous analytical work on EBG structures, most designs are simulated using commercially available software tools. To get a more deeper understanding on the wave propagation, analytical research and study on propagation characteristics need to be done.

From the above discussions, it is understood that although tremendous research works have been done on periodic structures, there are still plenty of room for making improvements, especially on one dimensional (1-D) split ring patterned EBG structures and optical controlled EBG structures and their applications. This thesis aims to develop new techniques to analyze these problems, which is the motivation of this Ph.D thesis.

1.2 Objectives

In responding to the aforementioned problems, the goal of this thesis is to investigate split ring patterned EBG structures and design optical controlled EBG structures for modern wireless communications. Since optical control has a feather of dynamic variation of permittivity, this could be exploited in filtering certain wireless channels with tunable characteristics. To fulfil the goal, the following tasks have been envisaged.

- (i) The study is begun with a comprehensive literature review of the structures, applications and analyzing methods of EBG structures. This exercise creates a strong foundation and background for further exploration.
- (ii) It is noted that split ring resonators (SRR) have many advantages and unique

1.3 Major Contributions of the Thesis

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characteristics compared to the conventional pattern EBG structures. To design different structures and devices with these special split ring resonators is very meaningful for microwave devices application.

- (iii) Another objective of the thesis work is to combine the EBG structures in an optical controlled semiconductor environment, that is, to design devices and models which are to be optically controlled.
- (iv) As there is not a rigorous analytical method relating the incident and reflected waves of this kind of periodic model, the work has been extended to develop an accurate and efficient method for analyzing different kinds of periodic structures including doubly periodic structures, especially optical controlled periodic structures.

The outcomes of the Ph.D study have originated a few fundamental contributions and developments over the existing research works. They are outlined in the next section.

1.3 Major Contributions of the Thesis

In the Ph.D research project on EBG structures and their applications, six major contributions have been made. Some of them have been published in international journals and conferences. The major contributions of the thesis are listed as follows: (Please see author's publications on page 107)

- (i) A thorough literature review on the previous research works on PBG/EBG structures have been carried out, through which it is concluded that PBG/EBG structures are playing important roles in the modern communication systems.

1.3 Major Contributions of the Thesis

From all the review, it is concluded that one dimensional PBG/EBG patterns have many advantages compared to conventional two dimensional (2-D) or three dimensional (3-D) periodic structures.

- (ii) The characteristics of numerous kinds of EBG patterns have been investigated, including split ring pattern EBG structures. A microstrip line loaded with different EBG structures have been designed and the stop band characteristics of all the structures have been qualitatively discussed. Especially for split ring resonant loaded microstrip line, the insertion loss is as high as 80 dB . Also the dimension of split ring resonator loaded microstrip line is much shorter in length than conventional square pattern EBG loaded microstrip lines.
- (iii) The combined defected ground plane structure (DGS) and split ring resonator (DGS-SRR) structure has been discussed. By applying split ring resonator into DGS structures, it can efficiently suppress the spurious response in the stopband compared to the conventional DGS structures. The 10dB bandwidth is increased by 50% and the size of the structure is greatly reduced using this novel structure.
- (iv) The dispersion characteristics of PBG/EBG structures have been discussed to get a clear understanding of the wave propagation in the structures. The scattering parameters of single cell split ring structure are obtained in complex value, and the wave propagation constant of the structure have been computed. The attenuation constant and the phase constant have been discussed to have an insight into the wave propagations in the periodic structures.

1.3 Major Contributions of the Thesis

- (v) Optical controlled semiconductor, microstrip lines with different kinds of EBG patterns and the plasma-induced EBG loaded microstrip line have been discussed in the subsequent chapter. The square pattern, ring pattern and split ring pattern EBG structures with and without plasma induced by laser illumination have been analyzed. First, the characteristics of the semiconductor illuminated by laser have been investigated and the results were given out indicating how the dielectric constant of the semiconductor were changing with the plasma density of the illuminating laser. Then microstrip line loaded with different EBG patterns have been designed and simulated in Ansoft HFSS, with and without laser illumination respectively. With the presence of the split ring resonator, it can achieve 80 *dB* insertion loss at 18 *GHz* without illumination (OFF status), and around 1 *dB* insertion loss when it is illuminated with the plasma density up to $10^{22}/m^3$ (ON status). It shows the possibility to implement an optical controlled switch using the split ring pattern EBG loaded microstrip line. At 18 *GHz*, when the microstrip line is not illuminated by the laser from the bottom, the transmission coefficient (S_{21}) is down to -80 *dB*, the switch is off; and when it is illuminated, the S_{21} is less than 1 *dB* and the switch is on. This kind of switch can achieve high isolation in microwave circuits since the controlled signal is optical and the speed of the switch can be of pico-second sampling.
- (vi) The characteristics of three-mode coupling in a dielectric slab waveguide having doubly periodic corrugations have been investigated. The three modes indicates the the fundamental mode, first higher-order mode and the first order backward reflection mode in the waveguide. The perturbation method

1.4 Organization of the Thesis

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with multiple scales was used in this research. The coupled-mode equations governing the first order Bragg interactions of the three propagating transverse electric (TE) modes have been given out. For the structure with fixed surface corrugation and variable dielectric corrugation, the results are given out for the reflection coefficients and the transmission coefficient of all three modes. It gives out the possibility of realizing a tunable rejection filter enhanced by optical excitation.

- (vii) The leaky wave characteristics on a photo induced double grating silicon slab rigorously by singular perturbation method based on multiple space scales have been analyzed. The leakage coefficient and the radiation efficiency are given out numerically and these radiation characteristics are investigated as a function of optically induced plasma density. The optical excitation enhances considerably the radiation efficiency and also gives flexibility in controlling the radiation behavior.

In some of the above studies, experiments have not been done due to non availability of high resistivity semiconductor samples and the multiple optical sources. However to verify the theoretical results, alternative simulation tools/special cases of early reported results have been used.

1.4 Organization of the Thesis

The remaining part of the thesis is organized as follows. Chapter 2 presents a literature review of the periodic structures. Firstly the previous research works on

1.4 Organization of the Thesis

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conventional periodic structures are investigated. Design and theoretical studies on electromagnetic bandgap structures are described. Then the existing research on novel split ring resonators in recent years are introduced. Coupled mode analysis and the perturbation method which are used to analyze the periodic grating structures are also introduced.

The studies are followed with analyzing different types of periodic structures. In Chapter 3, the analysis on complementary split ring pattern EBG etched on the ground plane of a microstrip line is proposed and its application in bandstop filters are also introduced. All the structures are simulated in Advanced design system (ADS) 2005a. After that, the structures are fabricated and measured using a vector network analyzer N5230A. Good agreement is observed for the scattering parameters between measured data and those calculated by our method. Also the novel DGS with SRR structures are presented in this chapter. The design and simulation of the defected ground plane structure with split ring resonator pattern are presented. The size of the split ring and the period of the structure are also analyzed using ADS 2005a. Fabrications are carried out for both traditional DGS structure and DGS-SRR structure. Also the results are examined using network analyzer to validate the analyzed results.

Chapter 4 investigates EBG structures etched on the ground plane of microstrip lines with silicon being the substrate. Both square pattern and SRR pattern are used in the research and they are etched on the ground plane of the strip line. The dielectric constant of the silicon changes when it is excited by laser of certain wavelength due to the introduction of free carrier. Simulations are performed with

1.4 Organization of the Thesis

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different laser intensities on the exposed ground plane apertures. The possibility of design and fabricating tunable on-off bandstop filters are proposed.

The characteristics of the three-mode coupling in a dielectric slab waveguide having doubly periodic corrugations are studied in Chapter 5. The structures discussed have surface grating in the upper side and permittivity grating in the lower side, so there are modes coupled to the original mode through the upper and lower corrugations respectively. The perturbation procedure is employed to analyze this structure, where the multiple scales and the boundary perturbations are applied. The coupled-mode equations governing the first order Bragg interactions of the three propagating TE modes are derived and the possibility of realizing tunable reflection filters is discussed in the chapter.

The principle of coupled mode analysis and perturbation method are introduced in Chapter 6. These two methods are used to analyze a periodic double grating structure. This structure is designed in a silicon slab. The upper side of the structure is slightly grating periodically and the other side is periodically stimulated by laser with different intensities to make permittivity modulation. The structure is analyzed by perturbation method with multiple space scales and the leakage coefficients of the structure with different laser intensities are described.

Finally, Chapter 7 concludes all these works and gives recommendations for further research.

Chapter 2

Literature Review of EBG Structures

2.1 History and Applications of PBG/EBG Structures

Periodic structures have been under study since many decades [11]–[18]. The wave propagation characteristics of periodic structures were extensively studied during these years [18]–[23]. Two special properties of those structures were revealed by the researchers:

- The eigen modes of periodic structures consisted of an infinite number of space harmonics and the phase velocities of these harmonics varied from zero to infinite.
- Propagating waves could only be supported in certain frequency bands, that is, there was certain passband and stopband in the frequency domain.

2.1 History and Applications of PBG/EBG Structures

Due to these unique characteristics, periodic structures found applications in a variety of fields like filters [24, 25], antennas [26, 27] and grating couplers [28, 29], etc.

There are many different forms of periodic structures, electromagnetic bandgap structure is one of them. Wave propagation in periodic structures has been an important and interesting subject to the electromagnetic society for many years. In the past one decade, attention has been focused on the special properties of wave propagation in periodic structures, which is called photonic bandgap structures, or electromagnetic bandgap structures. Terminology of photonic bandgap, which originated from the photonics field, has been introduced into the microwave field[21].

Although original research mainly focused on optical regime [30], as shown in Fig. 2.1, PBG structures are readily scaleable and applicable to a wide range of frequencies, including microwave and millimeter wave band [31]. Similarly to the energy bandgap concept in solid-state electronic materials, PBG structures, or EBG structures are periodic lattices which can provide effective and flexible control of the propagation of electromagnetic waves within a particular frequency range and along specific or all directions other than conventional guiding and/or filtering structures [32]. PBG/EBG structures are photonic analogous to semiconductor due to the fact that the electromagnetic waves are impeded in PBG/EBG perturbed substrates as photons are impeded in the semiconductor. Within these structures, the lack of electromagnetic propagation modes is referred to as a EBG, analogous to the energy bandgaps seen in the semiconductors. The concept of

2.1 History and Applications of PBG/EBG Structures

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photonic bandgap was first introduced by Yablonovitch and John in 1987 for semiconductor lasers and for photonic applications [33, 34]. The novel physical property of photonic band structures intrinsically originates from non-free-space electromagnetic dispersion relations and corresponding spatial field distributions generated in the periodic structures. The discovery and interest in photonics had grown explosively. An extended and exciting application for photonic crystals was seen in microstrip technology. There had been great interest and extensive effort in developing novel periodic structures for planar microwave circuits and antennas. An improved version of PBG/EBG lattice for microstrip-based application was PBG/EBG ground plane, so that no drilling through the substrate was required. Thus the fabrication process was greatly simplified.

The first comprehensive investigation of synthesized dielectric materials that possess distinctive stopbands for microstrip lines was reported by Yongxi Qian, Vesna Radisic and Tatsuo Itoh in 1997 [2]. Four types of PBG/EBG structures were utilized as substrates for microstrip based circuits as shown in Fig.2.2. They were:

- (a) Square lattice, square hole;
- (b) Triangular lattice, square hole;
- (c) Honeycomb lattice, square hole;
- (d) Honeycomb lattice, circular hole.

2.1 History and Applications of PBG/EBG Structures

(a)

(b)

Figure 2.1: The outlines of photonic crystal structures with (a) square and (b) hexagonal lattices. The white circles represent air holes in a dielectric background. The absence of an air hole in the central site forms the guiding defect [1].

2.1 History and Applications of PBG/EBG Structures

Figure 2.2: Several PBG (EBG) structures for microstrip circuits (a) Square lattice, square hole and (b) Triangular lattice, square hole and (c) Honeycomb lattice, square hole and (d) Honeycomb lattice, circular hole [2]

2.1 History and Applications of PBG/EBG Structures

Figure 2.3: Three-dimensional view of the proposed PBG (EBG) structure. The circular lattice circles are etched in the ground plane of a microstrip line [3].

The PBG/EBG holes were drilled through the dielectric substrate, and a conductive tape was applied in the ground plane of the microstrip line. There are various ways to construct PBG/EBG structures. They may consist of periodic perforations of multi-layered substrate, or they may be achieved through stacking 2-D or 3-D periodic metallic or dielectric structures [35, 36]. The PBG/EBG structure in [35] was made of resonant loop circuits periodically embedded in a dielectric host medium.

One and two dimensional periodic structures for electromagnetic waves had been studied since the early days of microwave radar and had been developed over the past 50 or more years. A new improved version of two-dimensional (2-D) electromagnetic bandgap structures for microstrip lines was then proposed by Radisic in 1998 [3], in which a periodic 2-D pattern consisting of circles was

2.1 History and Applications of PBG/EBG Structures

Figure 2.4: Photograph of a 2-D EBG microstrip reflector and a 1-D EBG microstrip reflector [4].

etched in the ground plane of microstrip line as in Fig. 2.3. No drilling through the substrate was needed, thus the fabrication process was greatly simplified. The experimental results of the proposed structure showed deeper and wider stopbands than previous designs using the dielectric hole approach.

Soon after this, it was discovered that due to the high confinement of the fields around the conductor strip in the microstrip line, it was possible to use a one-dimensional periodic pattern, obtaining similar behavior as the 2-D structures, which reduced the transverse dimension of the device [4]. The layout of this 1-D structure was, on the other hand, notable more compact than the layout of its equivalent 2-D structure, making it a more efficient choice for implementations as in Fig. 2.4. Etched holes with different periods, serial cascading as well as topologies used in 1-D EBG microstrip line structures had been shown to offer good performance. Similar 1-D EBG microstrip line structures in Fig. 2.4 were chosen to minimize the size of the structures in the study of this project.

2.1 History and Applications of PBG/EBG Structures

Over the last few years, there had been a lot of research works on EBG structures [2], [32], [36]–[46]. Most of the activities were aimed at the performance evaluations and applications of the EBG structures. The EBG structures can inhibit signal propagation in certain frequency bands and directions. It can be characterized in frequency domain, which consists of several passbands and stopbands. During the passband characteristic of EBG, waves can pass through freely and it is often used as slow wave medium, which reduces the size of the electronic circuits. While the wide stopband of EBG can be applied to suppress spurious transmission and leakage in guided structures.

Because of the wide-bandwidth of the propagation prevention band gap, EBGs found their applications in various microwave and millimeter wave devices. In [37], a filter designed with microstrip line loaded with U-shape EBGs was presented and achieved an excellent stopband performance and a high selectivity in a compact physical size. In [38], bandstop filters designed on coplanar waveguide (CPW) with unloaded and loaded EBG structures were discussed. The ripples in the passband region of filters could also be suppressed when EBGs were applied in those designs [39]. Less than 0.29 dB ripples in the adjacent passbands was achieved in this design. The PBGs/EBGs had been used in antenna design too [40]–[43]. The application of EBGs in antenna design could obtain a high gain with a very thin structure [44, 45]. In [36], the EBG material was designed in a waveguide slot antenna. While embedded in the exterior fringe of the slot array, the EBGs could effectively suppress the side and backward radiations and in the same time increase the gain of the slots antenna array. The high directivity antenna was also investigated with the design of EBG structures. In [52], the directivity

2.2 Metamaterials and Split Ring Resonators and Their Applications

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of the tapered slot antenna was increased by 240% by applying EBG structures. This dramatic improvement was achieved by micromachining periodic holes into the high permittivity substrate of the antenna.

In the past, there has been considerable work on the topologies of EBG structures. Despite of the normal square cells, circular cells and ring cells, other novel structures were proposed in [6]. Non-uniform EBG structures were also reported. Binomially distributed EBG structures were used to suppress the unwanted spurious transmission in a bandpass filter [7]. A 10-element EBG array with a tapered amplitude according to Chebyshev coefficients were introduced in [8]. The EBG assisted transmission line in the paper could provide excellent improvement in transmission coefficient both in low and high frequencies with wider rejection bandwidth.

It is obvious that more potential applications of PBGs/EBGs may be found in compact and high Q filters, antenna gain enhancement, miniaturization of antenna dimensions and waveguide structures, etc.

2.2 Metamaterials and Split Ring Resonators and Their Applications

Besides the EBGs, there has been a renewed interest in using fabricated structures to develop composite material that mimic known material response or that qualitatively have new, physically realizable response functions that do not occur, or

2.2 Metamaterials and Split Ring Resonators and Their Applications

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may not be readily available in nature. The history of artificial materials seems to trace back to 1898 when Jagadis Chunder Bose proposed his work on the rotation of the plane of polarization by man-made twisted structures [53], that is, artificial chiral structures. After this, various man-made materials were investigated by the researchers in the first half of the twentieth century. In the 1950s and 1960s, there were researches on artificial dielectrics for light-weight microwave antenna lenses, then in the 1980s and 1990s, there were resurrected interest in artificial chiral materials, especially in the applications for microwave radar absorber. In the past 10 years, novel artificial materials were proposed, including double negative (DNG) materials, i.e., artificial materials with simultaneous, effective negative real permittivity and permeability properties; negative index of refraction (NIR) materials; electromagnetic band gap structured material; and complex surfaces such as high impedance ground planes. The artificially fabricated inhomogeneities embedded in host media or connected to or embedded on host surfaces can generate new response functions of these metamaterials.

Metamaterials now have been described in several ways, such as left-handed materials, backward wave materials, negative index of refraction materials, etc. It is noted that while there has been much debate and controversy associated with whether these properties and their applications are real and realizable, it would appear that in the general physics and engineering literature we were beginning to see some consensus on favor of the validity and realization of many of the early claims. The generalized constitutive relations for metamaterials consisting of a 3-D array of inclusions of arbitrary shape were derived by Ishimaru in [54]. One possible realization of metamaterial was also introduced in his paper. In [55],

2.2 Metamaterials and Split Ring Resonators and Their Applications

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the electromagnetic properties of a pair of layered negative-permittivity-only and negative-permeability-only material were revealed by Alu. There were different possibilities to realize metamaterial. A composite medium realized by an array of spherical particles embedded in a background matrix could yield an effective negative permeability and permittivity [56]. The densely loaded L-C anisotropic grid over ground had been constructed, tested and then simulated to show negative refraction [57]. A periodic loaded transmission line that simultaneously exhibited a negative index of refraction and a negative group was introduced in [58]. Other achievement of double negative metamaterial was formed by the split ring resonator of magnetic elements [59, 60].

In some research, it was found that some properties of the EBG materials when used at a wavelength that does not belong to the band gap. Several effects such as negative refraction or control of the emission had been illustrated and fully understood using simple theoretical tools based on the dispersion relation of the Bloch modes in infinite EBG materials, and the continuity of the tangential component of the Bloch wave vector. It must be reminded that the effects that had shown were due to a collective behavior of the whole EBG materials and not to a local modification of the structure as in the case for example of a micro cavity made by creating a defect in a EBG material. In that sense, the EBG materials could be considered as a metamaterial as the left-handed material were, which raise the possibility of alternating the left-handed materials with EBG materials for the optical wavelength [61].

As an alternative to EBGs, another way to achieve the rejection of frequency

2.2 Metamaterials and Split Ring Resonators and Their Applications

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bands in planar transmission lines was by incorporating split ring resonator (SRR). First proposed by Pendry [62], SRRs were metallic rings with split on one side. As it was demonstrated in [62], SRR was proposed as a realization of negative effective magnetic permeability and it was popular as a physical realization of an left handed (LH) medium. By producing periodic structures using series of capacitive or inductive elements in either three dimensions or two dimensions, the negative permittivity / permeability could be realized [63]. When excited by a properly polarized radiation (magnetic field parallel to ring axis), SRRs were able to inhibit signal propagation in the vicinity of the resonant frequency. It could be considered as an externally driven LC circuit with a resonant frequency that could be easily tuned by varying device dimensions [64]. Also the complimentary split ring resonator (CSRR) was proposed in [65], they consisted of two concentric metallic rings with split on opposite sides. According to the concepts of duality and complementarity, it exhibited dual characteristics of SRR structures. It was important to note that the resonance was mainly a property of the individual cells, not due to the characteristics of the arrays. Therefore the period of the SRR structure could be very small compared with the traditional EBG structures. As CSRRs were sub-wavelength structures, their dimensions were electrically small at the resonant frequency and it was only one tenth of the guided wavelength or even less. In [66], a two-dimensional CSRR etched on the ground plane of a microstrip line was introduced and a band stop filter in the X-band was designed. Since the electromagnetic field of the strip line was concentrated near the strip, one-dimensional structure right beneath the strip line would have almost the same band stop characteristics. So in this work, 1-D topology was used to make it more compact. The purpose of this study was to discuss the performance of the CSRR

2.2 Metamaterials and Split Ring Resonators and Their Applications

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structure etched in the ground plane of the microstrip line, and to compare the simulation and measurement data of scattering parameters and also qualitatively discuss the dispersion characteristics of this structure.

The propagation characteristics of SRR and CSRR were discussed widely soon after they were proposed. The behavior of the SRRs and CSRRs were very similar to those of the EBGs and the defected ground structure (DGS). They all had bandstop and bandpass in the desired frequency band. In [67], the attenuation and phase constant of a microstrip line with CSRR etched ground plane was discussed and also the parameter study showed that the resonant frequency of this sub-wavelength structure was independent with the periodicity of the design. Besides this, to discuss the similar attenuation behavior and different attenuation level between the SRR and CSRR structure, the results in [68] pointed out that the propagation characteristics of the second attenuation pole of the CSRR was contributed by the second ring, compared to SRR structure. It was put forward that the two attenuation poles were both produced by CSRR structures and these two poles could be tuned considering two rings with different perimeters of etched open-loop ring resonator in CSRR structures.

Due to these unique characteristics of SRRs and CSRRs, they were widely used in microwave devices design, especially in band-reject filter designs [65, 66], [68]–[72]. By applying the SRRs and CSRRs into microstrip lines and coplanar waveguides, the designs showed compact, high Q bandstop filter with wide and deeper stop band. In [73], the comparisons of SRRs and CSRRs based band reject filters were investigated. With the same dimensions, SRR and CSRR based

2.2 Metamaterials and Split Ring Resonators and Their Applications

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band-reject filters were designed and simulated. Although the two filters were almost working at the same reject frequency, the attenuation and bandwidth of the filters were quite different. The design loaded with CSRRs exhibited much wider bandwidth and much deeper rejection level, which was the result of the second attenuation pole at a high frequency. In [72] a combination of DGS and SRR were presented by keeping SRR near the signal line of microstrip. By adjusting the coupling between DGS and SRR, a dual stopband filter has been achieved. The two resonant frequencies were controlled by the relative position of DGS and SRR pattern. The frequencies were also controlled by the dimension of SRR pattern.

The equivalent circuit models of the SRR and CSRR model were discussed by many researchers in order to understand the circuits more, as indicated in [72], the equivalent circuit model for SRR and DGS combined loaded microstrip line were presented. This RLC resonator could give out the two resonant frequencies of the model. Also the equivalent circuit model could tell how the positions and dimensions of the SRR and DGS pattern could change the resonant frequencies. In [74], the SRR loaded coplanar waveguide was discussed and the equivalent circuit model (ECM) was given. It is noticed for SRR and CSRR loaded microstrip line and coplanar waveguide, the ECM could mostly be present as the LC circuits with the same resonant frequency.

2.3 Methods of Analysis

2.3 Methods of Analysis

Among the last few years, researchers were trying to apply the numerical analysis method to study the PBGs (EBGs) and the metamaterials. Among which, the most notable one was plane wave expansion method (PEM) [75, 76], but this method was only applicable for uniform infinite extended crystals. The scattering matrix method was put forward to solve finite periodic structures. Although this method was even applicable for non-periodic structures, it could only deal with 2-D models [77]. To solve 3-D EBG problems, the finite-difference time-domain (FDTD) and the finite element method (FEM) were applied widely [35], [78]–[82]. In [82], the FEM was combined with domain decomposition algorithm to solve EBGs structures. Based on the finite element approximation and a non-overlapping domain decomposition method, this algorithm could solve time-harmonic electromagnetic fields arising in three dimensional, finite-size photonic band gap and electromagnetic band gap structures.

Besides the analysis methods mentioned, there were lots of other methods such as boundary value methods, perturbation techniques, eigen system methods, and the coupled mode approach. Among them, the singular perturbation method based on multiple space scales was chosen here. One of the advantages of this method is that it is purely analytical and quite accurate if the structural corrugation is less than 10% and permittivity corrugation is between 5% and 20% [83]–[86]. To solve a structure using perturbation method, one should first get the expansion of the fields on multiple space scales, then get the differential functions satisfying the fields for each order, also the boundary condition of each order problem. From the

2.4 Conclusion

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first order solution, the relations between the radiated wave, the guided wave and the incident wave are derived, while the second order solution yield the amplitude transport equation. Using these results, the radiation efficiency and the exact radiation angle can be calculated. Comparing with FDTD and FEM method, this singular perturbation method could predict the radiation angle and radiation efficiency reasonably accurate by solving the first order and second order problems.

2.4 Conclusion

The review of the periodic structures has been carried out in this chapter. The advantages and applications of the PBGs/EBGs have been discussed, and different analysis methods for periodic structures have been compared and perturbation method has been chosen to carry out the study in this thesis.

Chapter 3

Investigation of Split Ring Resonator EBG Structures

3.1 Introduction

Since the concept of electromagnetic bandgap was first introduced into the microwave field, there has been great interest and extensive effort in developing novel periodic structures for their applications to planar microwave circuits and antennas. An improved version of EBG lattice for microstrip-based application was EBG ground plane, so that no drilling through the substrate was required. Thus the fabrication process was greatly simplified.

1-D and 2-D planar configurations were more attractive due to the fact that they were ease of fabrication, less expensive and compatible to MMIC design. A one-dimensional (1-D) structures could be made by alternating the wave impedances, which had already been analyzed and applied to transmission lines and waveguides

3.1 Introduction

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in microwave engineering to demonstrate the stopband and the slow wave characteristics. Two-dimensional (2-D) structures published in the literature for antenna and microstrip line applications consisted of periodical air holes, which were realized either micro-machined or drilled through the substrate [52]. In microstrip technology, EBG structures were obtained by introducing an appropriate periodic pattern drilled or implanted in the substrate or etched in the ground plane. In the first method, periodic implants, which were comparable in size to a wavelength, may be metallic, dielectric, magneto-dielectric, ferromagnetic, ferroelectric, or active. However the second method was much easier to implement and compatible with monolithic technology, and deeper and wider stopbands can be obtained.

As mentioned in Chapter 2, EBG structures found potential application in various microwave devices. Compared with a shorted patch on a conventional ground plane, a shorted microstrip patch of identical dimensions with a EBG ground plane had been designed and showed significant improvements in the gain and the reduction of cross-polarization levels [9]. A novel power amplifier, which incorporated EBG ground plane, was introduced in [87], and shown that the structure not only limited intrinsic second and third harmonics tuning without any filters, but also offered the potential of greatly reducing the size of the amplifier. The EBG structures are playing more and more important roles in enhancing the performance of microwave devices now. It is meaningful to continue with the investigations and studies on EBG structures.

The remaining parts of this chapter is organized as follows. Section 3.2 explains about the design procedure of the microstrip lines with EBGs etched in the

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

ground plane. Numerical results of scattering parameters and circuit parameters for CSRR pattern EBGs in the ground plane are also presented in this section. The parametrical study is also performed in this part. After that, Section 3.3 gives out the design and results on this novel defected groundplane structure (DGS) and SRR combined structures, and finally the conclusions are provided in Section 3.4.

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

3.2.1 Formulation of the Problem

After the traditional square pattern of the EBG structure, various kinds of novel patterns were put forward to achieve better performance in terms of high suppression. It is well known that the complementary split ring structure is obtained by replacing the metal parts of the original split ring structure with apertures, and the apertures with metal plates [88]. In this way, one can implement a CSRR in microstrip line by etching the split ring structure under the strip in the ground plane. Figure 3.1 gives the topology of the rectangular-shaped CSRR and its dimensions. Figure 3.2 is the photograph of the fabricated microstrip line with three CSRRs and the detailed CSRR structure in the ground plane. In Fig. 3.1, if the CSRR is properly excited, the resonant frequency is determined by the dimension of the CSRR.

The ring pattern CSRR schematics have been introduced in Fig. ???. As have been mentioned, the whole SRR can be considered as an LC resonator [64]. The

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

Figure 3.1: (a) Layout of unit cell of CSRR used in the structure; (b) Proposed structure with three cells etched on the ground plane.

equivalent inductance is considered as the inductance of a closed ring, with its length being $2\pi r_0$ (r_0 is the average radius of the structure) and width c . The capacitance is calculated as the series connection of the capacitance of the left and right halves of the SRR, which is given by $\pi r_0 C$, where C is the per unit length capacitance between the rings. So the resonant frequency ω_0 can be expressed as

$$\omega_0 = \sqrt{\frac{2}{\pi r_0 L C}} \quad (3.1)$$

where L is the total inductance of the SRR. By increasing the dimension of the split ring, the resonant frequency will decrease. In my study, the EBG pattern is

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

designed to be square shaped SRR, which has the same analysis procedure as ring shaped SRR. Both patterns have the same performance, for the ease of fabrication, the square pattern SRR is studied here.

(a)

(b)

Figure 3.2: Photograph of the fabricated prototype (a) strip line and (b) ground plane.

In order to get an idea about the wave propagation in structures, dispersion characteristics of the structures were studied [89]. The scattering parameters of

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

Figure 3.3: Equivalent circuit model of split ring resonator.

single cell CSRR structure were obtained in complex value, and the equation (3.2) has been used to compute the wave propagation constant γ of the structure. The real part of γ is the attenuation constant and the imaginary part of γ is the phase constant [90].

$$\gamma = \frac{1}{\Lambda} \cosh^{-1} \frac{[(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}] + (\frac{Z_{01}}{Z_{02}})[(1 - S_{11})(1 + S_{22} + S_{12}S_{21})]}{4S_{21}} \quad (3.2)$$

$$\gamma = \alpha + j\beta \quad (3.3)$$

Here, Z_{01} and Z_{02} are the impedances of port 1 and port 2, respectively. In this study, they are assumed as 50Ω , and Λ is the period of the structure. If T is the wave transmission matrix of the structure, the complex characteristic impedance ($Z_c = Re(Z_c) + Im(Z_c)$) is calculated in terms of the T and scattering (S) parameters as:

$$T = \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} \quad (3.4)$$

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

$$Z_c = \sqrt{\frac{T_{12}}{T_{21}}} = Z_0 \sqrt{\frac{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}} \quad (3.5)$$

3.2.2 Numerical Results

In the structures shown in Fig. 3.1 and Fig. 3.2, the thickness of the microstrip line is 0.635 mm and the width of the strip line is 0.93 mm and is designed for 50Ω characteristic impedance. The material used is RT duriod 6006, its permittivity is 6.15 and its loss tangent is 0.002. The size of the outer square is 2 mm and size of the inner square is 1 mm , the width of the ring is 0.3 mm , and the split width of the ring is 0.3 mm . There are three cells etched out in the ground plane and the distance between them is 7 mm , which is about half wavelength of the microstrip line. In the later part of the work, the independence of the periodicity on the scattering parameters will be discussed.

The structure is simulated in ADS 2003a and the results are given in Fig. 3.4. It is the stop band of the structure in frequency range 12 GHz to 14 GHz and the maximum insertion loss of $|S_{21}|$ is up to 40 dB . The measurement of the fabricated structure is also given in Fig. 3.5. The measured scattering parameters are matching well with the simulated results by ADS Momentum.

For the dispersion characteristics, the obvious stopband can be seen from the data of scattering parameters. The propagation constant of unit cell of the structure is given in Fig. 3.6 and Fig. 3.7.

It can be observed in the results that the normalized attenuation constant α/k_0

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

Figure 3.4: Numerical results of scattering parameters simulated.

has become significant and is nonzero from 12 GHz to 14 GHz , i.e., the band stop of the structure analyzed, while the normalized phase constant β/k_0 phase matches with Bragg frequencies during the stop band region. Thus, the range of frequencies with nonzero attenuation constant is referred to band gap or band stop, because of its wave attenuation or rejection behavior characteristic.

Figure 3.7 describes the extracted complex characteristic impedance with real and imaginary parts, $Re(Z_c)$ and $Im(Z_c)$. When the frequency is in the stop band, as expected the real part of the impedance vanishes and the imaginary part dominates.

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

(a)

(b)

Figure 3.5: Measured and simulated (a) S_{11} and (b) S_{21}

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

(a)

(b)

Figure 3.6: Normalized (a) attenuation constant and (b) phase constant of unit cell

3.2 Design of Microstrip Lines with EBGs Etched on the Ground Plane

Figure 3.7: Characteristic impedance of unit cell.

It is also observed in Fig. 3.6 that in the frequency band near 9 GHz , there is another stop band with a very small attenuation constant value, resulting in the insertion loss of around 1dB. In Fig. 3.7, the behavior of the complex characteristic impedance verifies it. However, there is not an obvious stop band in the scattering parameters. This structure is simulated both in Ansoft high frequency structure simulator (HFSS) 8.0 and ADS 2003a, the former is using finite element method (FEM) and the later by method of moments (MOM) as the simulating method respectively to verify the spurious behavior of the stopband. But the bandstop remained as it is near 9 GHz even with the change in width of the split ring and in the absence of inner split ring. Thus the appearance of the stop band is not

3.3 Design of DGS-SRR Structures

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because of the simulating method or the mutual coupling of the inner split ring and the outer split ring and the width of the split is not influencing on it. So this band stop may be due to some spurious behavior of the split ring structure itself. Although such a band stop exists in the dispersion characteristic, it doesn't affect the scattering parameters much as shown in Fig. 3.4 and Fig. 3.5.

It has also been investigated by changing the orientation of CSRR within three cell topology, and found that no significant change in the transmission characteristics. The reason might be that the CSRR loaded microstrip line behave like a resonant structure and the orientation will not have any dominant effect. In order to investigate the dependence of periodicity, different periods of the CSRR cell structure have been applied and simulated. It is from that as discussed above, the resonance of CSRR dominates over other effects like period and orientation. They are simulated by ADS 2003a and the transmission coefficients are shown in Fig. 3.8. The resonant frequencies are almost the same for the structures with period 5 mm, 6 mm and 7 mm. Therefore, it is sub-wavelength resonator and the size of the structure can be made even smaller, and the ring dimensions dictate the frequency and amount of insertion loss of the level of suppression is contributed by the array dimension.

3.3 Design of DGS-SRR Structures

In the past decades, researchers have paid great attention on the research of all types of filters and focused mainly to improve their performance. Recently, de-

3.3 Design of DGS-SRR Structures

Figure 3.8: Transmission coefficients of 3-cell structures with different periods.

affected ground structure was introduced in designing low pass filters with the objective of suppressing the stopband effectively. A unit cell of DGS is a dumbbell like slot aperture in the ground plane of the microstrip structure, with its equivalent circuit model shown in Fig. 3.9 [91]. As proposed in [92]-[96], the DGS unit can exhibit a stopband characteristic at certain frequencies due to the attenuation pole. The equivalent circuit model for DGS was proposed in [93, 94, 96], which shows the relation between the size of the DGS and its band stop frequency. Several DGS units can be cascaded together to achieve a wider stopband. However, this leads to the size of the structure to be large.

3.3 Design of DGS-SRR Structures

Figure 3.9: (a) Schematic of a unit DGS cell in a microstrip line and (b) its equivalent circuit model.

3.3 Design of DGS-SRR Structures

In order to get a wide bandwidth with the same structure length, the split ring resonator is introduced. As stated in Chapter 2, split ring resonator is another structure in recent times which microwave community is focusing due to its left-handed characteristic. SRR has also been introduced in filter designing [67]. It is shown that for a bandstop filter SRR can greatly reduce the size of the filter while retaining a very high stopband attenuation. This is because the evanescent modes can propagate in SRR structures.

In this section, a novel design combining DGS with SRR is presented. Two low pass filters with only DGS and DGS-SRR are designed and simulated using ADS 2005A. The structures are fabricated and measured to validate the simulated results.

A substrate material of RO 4003 is used which has a dielectric constant of 3.38, a thickness of 60 *mil* and loss tangent of the substrate is 0.02. Therefore, all the DGS and DGS-SRR structures in this work are designed using this material, for the comparison purpose. Figure 3.10(a) shows the conventional DGS on a microstrip line. There are dumbbell like slots created in the ground plane of the 50 Ω microstrip line. By forming an equivalent circuit for unit cell structures, one could analyze the resonant frequency and corresponding dimensions of DGS and SRR structures. Later DGS-SRR structure has been designed, as illustrated in Fig. 3.10(b). The fabricated structures are shown in Fig. 3.11(a) and (b). Compared to normal DGS, some of the DGS slots are replaced by SRR slots and the unit of the new structure consists of one DGS and one SRR, alternatively. The overall length of all these designs is almost the same, while the 10-dB return loss

3.3 Design of DGS-SRR Structures

bandwidth of the DGS-SRR design is much wider than the conventional one.

Figure 3.10: The fabricated structures (a) Conventional DGS slots on a microstrip line; (b) Novel DGS-SRR slots design on a microstrip line.

Agilent ADS 2005a has been used to simulate all the structures. For a unit cell of DGS-SRR case, the simulated results are shown in Fig. 3.12. For DGS

3.3 Design of DGS-SRR Structures

(a)

(b)

Figure 3.11: (a) Conventional DGS slots on a microstrip line; (b) Novel DGS-SRR slots design on a microstrip line.

3.3 Design of DGS-SRR Structures

only design, the pole is at 4.2 GHz . It is obviously from Fig. 3.12 that DGS-SRR design has an additional pole near frequency 5.8 GHz compared with DGS only design.

Figure 3.12: Simulated results of unit DGS only and unit DGS-SRR cell (Dimensions of the unit cell are shown in the inset).

Next, the structural dimensions are optimized to improve the transition region between the passband and stopband and employed eight units in our design. Figure 3.13(a) shows the simulated results of scattering parameters in our designs. It can be observed that spurious response is greatly suppressed in the frequency range of 7 GHz to 10 GHz by introducing SRR slots.

3.4 Conclusion

Both the DGS and the DGS-SRR structures are fabricated to verify our design. Good agreement between the simulated and measured data has been observed. The measured results for both the DGS and DGS-SRR are plotted in Fig. 3.13(b) for comparison. Again it is seen that the DGS-SRR has a wider stopband than DGS, as expected, the stopband of the low pass filter is increased. The results show that combined DGS-SRR design can efficiently suppress the spurious response in the stopband compared to the conventional DGS structures.

Furthermore, the length of DGS-SRR structure is reduced and only three cells are used. The simulation results of transmission coefficients are shown in Fig. 3.14. It is clearly shown in the figure that three-cell structure is sufficient to suppress the spurious response up to $30dB$ and give out a wider stop band. So the dimension of DGS-SRR structure can be reduced by 25%.

3.4 Conclusion

A study on the stop band characteristics of microstrip lines loaded with PBG structures has been carried out in this work.

First part of the chapter is mainly discussed about CSRRs. A very large insertion loss has been obtained in stopband by using CSRRs. Because the bandgap characteristic is the behavior of the CSRR itself, not because of the array of the periodic structure, the dimensions of the structure can be made much smaller than in the case of a microstrip line with a periodic perturbation of square pattern on

3.4 Conclusion

(a)

(b)

Figure 3.13: (a) Simulated results of DGS and DGS-SRR design; (b) Measured results of DGS and DGS-SRR design.

3.4 Conclusion

Figure 3.14: Simulated transmission coefficients of DGS-SRR structures with 3 cells and 4 cells.

3.4 Conclusion

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the ground plane. The attenuation and propagation constants of unit cell of this structure are extracted and a physical behavior of stop band of this CSRR loaded microstrip line has been demonstrated. The high attenuation and suitable band width of the sub-wavelength resonator is useful for rejecting the jamming noise in communication systems.

In the last section of the chapter, the combined DGS-SRR structure is discussed. Combining SRR into DGS structures, results show that this design can efficiently suppress the spurious response in the stopband compared to the conventional DGS structures. The stopband is significantly improved using this novel structure. The stopband around 3 GHz to 10 GHz is very suitable for filter application in ultra wideband (UWB) devices. The whole length of the structure can also be reduced by 25% and make it more compact in size.

Chapter 4

Optical Controlled Periodic Structures

4.1 Introduction

Most of the conventional EBG structures analyzed, as all of the structures discussed in Chapter 3, were fixed periodic structures and dynamic tuning capabilities are very much needed for various applications. Other than fixed periodic structures in microwave fields, there is considerable interest focused on the field of optical control of microwave and millimeter wave devices, circuits and systems, such as phased array antennas, modulators, couplers, switches, generation of millimeter waves and optical probing. This new technology combines the fields of optics and electronics and it is referred to pico-second electronics, and pico-second photoconductivity is the link between these two fields, as explained in [97].

The optical technique for controlling these devices offers unique advantages:

4.1 Introduction

- (i) Since the controlling signal is optical and the controlled signal is electrical, near-perfect isolation can be achieved;
- (ii) Immunity from electromagnetic interference;
- (iii) Fast response;
- (iv) Possibilities for monolithic integration, that is, a mixture of MMIC (monolithic microwave integrated circuits) and OEIC (optoelectronic integrated circuits).

The optical control of microwave and millimeter wave devices can be achieved by directly controlling the passive components such as microstrip lines, image lines, or coplanar waveguides realized on high resistive semiconductor substrate. A light with photon energy that is greater than the semiconductor's band gap energy, induces electron-hole pairs in a semiconductor optically. Thus, one can control the waves in the semiconductor. In this work, this procedure has been used to exploit the optically induced plasma effect on the wave propagation.

In this chapter, the principles of optical controlled semiconductor are introduced in Section 4.2. Then Section 4.3 describes the design and formulation of the on-off filter controlled optically. This filter is designed by microstrip line with substrate silicon. PBG slots are still etched in the ground plane of the strip line. The on-off characteristic of the filter is controlled by laser illumination on the PBG pattern. This structure are simulated with Ansoft's high frequency structure simulator (HFSS) 8.0 and the numerical results are given out in Section 4.4. After that the conclusion is drawn in Section 4.5.

4.2 Principle of Optical Controlled Semiconductor

4.2 Principle of Optical Controlled Semiconductor

Semiconductors such as germanium and silicon have two important characteristics. First, column IV elements such as germanium and silicon have four electrons in their outermost occupied (valence) band, yet only a moderate activation energy is required to move an electron to the next outermost (conduction) band. As solids, adjacent atoms share electrons, forming covalent bonds. Eight electrons can occupy the valence band around each atom, but the four host atom electrons together with four shared neighboring electrons occupy all of the vacancies. A second important characteristic arising from the covalent bonds is that these elements form regular crystalline solids with atoms so close together that neighboring atoms electron orbits overlap. Thus, an electron, when once imparted with the necessary energy to reach the conduction band, is able to move from atom to atom through the regular crystalline lattice quite freely, thereby providing current flow.

So far, the picture of a semiconductor is one of a crystal in which the electron flow can occur readily, provided that the electrons can be somehow induced to occupy the normally unfilled outer energy (conduction) band about their host atoms. One method of achieving such electron activation is by heating the semiconductor. In the way, the average energy of electrons surrounding the atoms is increasing causing some to be excited to higher energy states in the conduction band. Another means of promoting electrons to the conduction band is by having the photons of light energy absorbed by the crystal. The resulting increase in energy can raise electrons from the valence to the conduction band. Such pho-

4.2 Principle of Optical Controlled Semiconductor

Figure 4.1: Energy band schematic representation for pure silicon showing (intrinsic) creation of an electron hole pair of free carriers [5].

tons simulation of the electrons is also a means of increasing the conductivity of a semiconductor [5].

These two mechanisms results in intrinsic conduction, a diagram for which is shown in Fig. 4.1. The electron excited to the conduction band leaves behind a vacancy that could be occupied by a valence electron from a neighboring atom. Filling this vacancy (called a hole) creates a now vacancy in the neighboring atom. This moving vacancy results in a net current flow similar to that of a positive mobile charge. To distinguish it from the electron flow in the conduction band, it is called hole carrier flow.

Before the discussion on how the permittivity varies with the plasma density, the wavelength of the laser is considered to excite the carriers properly. Fig. 4.2 shows the relationship between absorption coefficient of different materials and laser wavelength. Since the material in this study was silicon, laser wavelength is

4.2 Principle of Optical Controlled Semiconductor

chosen from 800 nm to 850 nm so that the silicon is excited properly. Another potential material for optically excited behavior is GaAs.

Figure 4.2: Absorption coefficient vs. laser wavelength for various semiconducting materials [5]

When the semiconductor is illuminated with the photon energy greater than the bandgap energy between valence band and conduction band of the semiconductor, photons are absorbed; creating electron/hole pairs and resulting thin layer plasma

4.2 Principle of Optical Controlled Semiconductor

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region near the surface of the waveguide. The presence of free carrier/electron-hole plasma results in the modification of conductive and dielectric properties of the semiconductor material according to Drude-Lorentz formula [5]:

$$\varepsilon_p = \varepsilon_s - \sum_{i=e,h} \frac{\omega_{pi}^2}{\omega^2 + \gamma_i^2} (1 + j \frac{\gamma_i}{\omega}) \quad (4.1)$$

Where the subscript i denotes the different kinds of the carriers and, ε_s is the dielectric constant of the host lattice including the contribution of bound charges, γ_i is the collision angular frequency of the carrier, ω_{pi} is the plasma frequency which can be expressed as [5]:

$$\omega_{pi}^2 = \frac{n_i e^2}{\varepsilon_0 m_i^*}, \quad i = e, h \quad (4.2)$$

Here n_e is the electron concentration in the conduction band, n_p is the hole concentration in the valence band, it is assumed that the carrier concentration of both electron and hole are the same.

$$n_e = n_p = N_p \quad (4.3)$$

m_e^* is the effective mass of electrons/holes, e is the electronic charge which equals to $1.6 \times 10^{-19} \text{ C}$ and ε_0 is the free space permittivity, which is $8.854 \times 10^{-12} \text{ F/m}$.

The material parameters for silicon required in the above equations are [5]:

4.2 Principle of Optical Controlled Semiconductor

$$\begin{aligned}
 \varepsilon_s &= 11.8 \\
 m_e^* &= 0.259m_0, \quad m_h^* = 0.38m_0, \quad m_0 = 9.11 \times 10^{-31} \text{Kg} \\
 \gamma_e &= 4.52 \times 10^{12}/\text{sec}, \quad \gamma_h = 7.71 \times 10^{12}/\text{sec} \\
 \omega_{pe} &= 1.226 \times 10^4 n_e, \quad \omega_{ph} = 8.3559 \times 10^3 n_h
 \end{aligned} \tag{4.4}$$

Figure 4.3: Real and imaginary parts of permittivity for different optically induced plasma carrier densities.

The results of the permittivity for different frequencies are shown in Fig. 4.3. The real and imaginary parts of the permittivity as a function of optical induced plasma density n_e have been computed. The estimation of the complex permittivity from Eq. (4.1) shows that the real part of the permittivity ε_{pr} is not affected much by the plasma density up to $10^{21}/\text{m}^3$. But the imaginary part of the permittivity ε_{pi} , i.e. conductivity, is sensitive even for low plasma density, below $10^{18}/\text{m}^3$. Notice that at $n_e = 10^{19}/\text{m}^3$, the imaginary part of the permittivity

4.3 Optical Controlled EBG on Microstrip Lines

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starts to deviate from its original value, and at $n_e = 10^{20}/m^3$, the imaginary part becomes comparable to the real part of the permittivity. It is expected that, at about this density, a detectable change in the phase of the millimeter wave will start taking place. The increase of conductivity with the plasma density will result in the reduction in power of the millimeter wave. If an optical beam is focused on a small region, the power carried by the millimeter wave will be perturbed only at the illuminated spot.

4.3 Optical Controlled EBG on Microstrip Lines

The structures discussed in this chapter are also microstrip lines with EBG structures etched on the ground plane. An optical source will be used to illuminate the microstrip line from underneath. Since the ground plane is etched periodically, so a thin layer of plasma with the same structure of the EBG pattern in the ground plane will be appeared. In Fig. 4.4, it is the schematic of the square pattern EBG microstrip line illuminated with the laser. In this study, since the size of the EBG pattern was small and the thickness of the plasma was very thin, the shape of the plasma assumed to be the same as the EBG pattern in groundplane. The thickness of the plasma is $0.05mm$ after illumination.

To get a high on-off ratio, ring patterned EBG structure and split ring patterned structure have been used due to the high resonance behavior. When using such structure, the area of the plasma is greatly reduced and the energy loss caused by the imaginary part of the plasma is also decreased. Also the split ring resonators

4.3 Optical Controlled EBG on Microstrip Lines

Figure 4.4: (a) The top view and (b) The side view of the square pattern EBG microstrip line illuminated by laser.

4.3 Optical Controlled EBG on Microstrip Lines

exhibit good insertion loss, which could be exploited for switching applications.

The plasma-induced ring patterned EBG structure and the split ring patterned structure are shown in Fig. 4.5 and Fig. 4.6 respectively.

Figure 4.5: (a) The top view and (b) The side view of the ring pattern EBG microstrip line illuminated by laser.

4.3 Optical Controlled EBG on Microstrip Lines

Figure 4.6: (a) The top view and (b) The side view of the split ring pattern EBG microstrip line illuminated by laser.

4.4 Numerical Results

4.4 Numerical Results

Ansoft HFSS 8.0 was used to simulate the structure. When the intensity of the laser is different, the density of the plasma is also different, which will affect the wave propagation in the microstrip line. In Fig. 4.7, scattering parameters are shown for various densities of the plasma such as $10^{18}/m^3$, $10^{20}/m^3$, $10^{22}/m^3$. This result is for the square pattern EBG structure shown in Fig. 4.4.

During the simulation, since the permittivity of the illuminated silicon varied with frequency, the whole frequency band was divided into several sub-bands. In each sub-bands, the permittivity was assumed to be uniform (with real and imaginary parts). This is because that HFSS could not handle frequency dependent permittivity problems. In the simulation, up to 57565 tetrahedra mesh were used and the number of sub-bands used were 7.

We can see clearly from Fig. 4.7, with the increasing density of the plasma, the scattering parameters are different, the attenuation becomes less and less in the stop band frequency range which is almost -10 dB without illuminating. More energy is lost (changing to other forms of energy) when the density of the plasma is higher, which becomes a main disadvantage of the structure.

In order to verify the results given by HFSS, the structure was also simulated in CST Studio Suite 2006b, which applies FDTD in the simulation. The comparison of results given by HFSS and CST was given in Fig. 4.8. Two extreme cases with no illumination and with high laser intensity induced plasma density $10^{22}/m^3$ were

4.4 Numerical Results

(a)

(b)

Figure 4.7: (a) Parameter $|S_{11}|$ and (b) Parameter $|S_{21}|$ when n_p is $10^{18}/m^3$, $10^{20}/m^3$, $10^{22}/m^3$ with square patterned EBG.

4.4 Numerical Results

(a)

(b)

Figure 4.8: Comparison of simulation results given by HFSS and CST with square patterned EBG (a) no illumination, plasma density N_p is zero and (b) plasma density is $10^{22}/m^3$.

4.5 Conclusion

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studied for square patterned EBG. From the figure, one could see the trends of the curves were almost the same and the values of S parameters were similar.

Figure 4.9 shows the comparison of scattering parameters where there is no laser illumination and the laser intensity is $10^{22}/m^3$ for square pattern EBGs. To improve the structure and reduce the energy loss, a ring pattern structure has been used. In Fig. 4.10, the results of ring pattern EBG was given out. When it was illuminated with optical density $10^{22}/m^3$, S_{21} is around -1 dB and the energy loss is much less than that of a square patterned EBG structure. Comparing with the non-plasma ring patterned EBG structure, the difference of S_{21} between these two is not very large, about 10 dB only.

In order to get a larger difference between these two cases, split ring structure is introduced, which exhibit maximum attenuation in the stopband up to 80 dB . For split ring pattern EBGs, the comparison of scattering parameters where there is no laser illumination and the laser intensity of equivalent plasma density is $10^{22}/m^3$ is shown in Fig. 4.11. Compared with the Fig. 4.10, $|S_{21}|$ in Fig. 4.11 is also around -1 dB when the plasma density is $10^{22}/m^3$ and the energy loss is low in this case also.

4.5 Conclusion

In this chapter, it is discussed about optical controlled semiconductor, microstrip lines with different kinds of EBG patterns and the plasma-induced EBG microstrip

4.5 Conclusion

(a)

(b)

Figure 4.9: Comparison of (a) Parameter $|S_{11}|$ and (b) Parameter $|S_{21}|$ when there is no laser illumination and laser intensity is $10^{22}/m^3$ applying to the square pattern EBG.

4.5 Conclusion

(a)

(b)

Figure 4.10: Comparison of (a) Parameter $|S_{11}|$ and (b) Parameter $|S_{21}|$ when there is no laser illumination and plasma density is $10^{22}/m^3$ applying to the ring pattern EBG.

4.5 Conclusion

(a)

(b)

Figure 4.11: Comparison of (a) Parameter $|S_{11}|$ and (b) Parameter $|S_{21}|$ when there is no laser illumination and plasma density is $10^{22}/\text{m}^3$ applying to the split ring pattern EBG.

4.5 Conclusion

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line. The square pattern, ring pattern and split ring pattern EBG structures with and without plasma induced have been analyzed. For the square pattern EBG case, when it is illuminated, the results show that the energy loss is high, which is due to the large plasma absorbing area, that is, square EBGs on the ground plane. So it is not suitable for realization of microwave circuits with such high loss. For ring pattern and split ring pattern EBG microstrip line, when it is illuminated from the ground plane, the simulated results show that scattering parameter S_{21} comes up to about $1dB$ when the plasma density is up to $10^{22}/m^3$. When the split ring pattern EBG structure is not illuminated by the optical source, the insertion loss at 18 GHz is up to 80 dB , it shows the possibility to fabricate an optical controlled switch using the split ring pattern EBG microstrip line. At 18 GHz , when the microstrip line is not illuminated by the laser from the ground, the S_{21} is up to $-80dB$, the switch is off; and when it is illuminated, the S_{21} is less than 1 dB and the switch is on. This kind of switch can achieve high isolation in microwave circuits since the controlled signal is optical and the speed of the switch can be of pico-second sampling. Although the lack of facilities restrain the work purely theoretically, the results obtained still have a promising potential of high speed and high isolation on-off switches. To validate the results obtained by HFSS, alternative simulation tool CST microwave studio was employed and demonstrated consistency of the results.

Chapter 5

Mode Coupling in the Optically Excited Double Periodic Structures

5.1 Introduction

As has been stated in Chapter 4, considerable interest has been focused on the field of optical control of microwave and millimeter wave devices, circuits and systems [98, 99]. Having known the unique advantages of optical control which are reported in Chapter 4 [10], yet another problem of mode coupling in a semiconductor waveguide has been explored in this chapter.

Several research works have been carried out studying the wave interaction in thin-film dielectric waveguides with periodic discontinuities [100, 101]. The principle of the wave interaction is important to reveal the fundamental properties of

5.1 Introduction

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wave propagation in periodic structures and to the application of optical filters. If the waveguide includes a single periodic structure, there will be modal coupling between the incident guided wave propagating in one direction and the coherently reflected guided waves propagating in the opposite direction when the Bragg condition is satisfied [83], [102, 103]. There is a particular interest in the study of wave interaction in a waveguide with multi-periodic structures. When the study of wave propagation in doubly periodic structures is considered, an approximation is taken and a three-mode coupling are considered when two appropriate Bragg conditions are satisfied. The three modes indicates the fundamental mode, first higher-order mode and the first order backward reflection mode. The three-mode coupling in a thin film dielectric slab waveguide in which the permittivity has a doubly periodic variation in the propagation direction was analyzed by Seshadri [84]. The characteristics of three-mode coupling in the dielectric slab waveguide having doubly periodic surface corrugations were investigated by Yasumoto [86]. In [86] the double grating were both structural and in [84] the double grating were both about permittivity modulation. Some of the integrated circuits realized on silicon substrate have grating configuration for filter application. These devices are for a fixed frequency and if the tuning is necessary, then optically controlled behavior could be exploited. Hence in this chapter, the study has been carried out on a dielectric slab waveguide with corrugations at the surface in the upper side and permittivity grating in the lower side. The structure studied in this chapter combined the surface corrugation and permittivity grating at the same time, which has not been explored before. Also, tunability becomes possible for this structure if we control the laser density. The characteristics of the three-mode coupling in this structure have been investigated.

5.2 Analysis Model

Although lot of methods could deal with this structure, such as boundary value methods, eigen system methods and perturbation techniques, the perturbation method is chosen so that the relations between the radiated wave, the guided wave and the incident wave can be given directly from the first order solution. The perturbation procedure is employed and the multiple scales and the boundary perturbations are applied [83, 104]. The flow chart of perturbation method is shown in Fig. 5.1. The coupled-mode equations governing the first order Bragg interactions of the three propagating TE modes are derived. The equations are solved for the case where the incident guided wave couples to the coherently reflected two guided waves. The reflection and transmission characteristics of this periodic waveguide of finite length with different index of permittivity corrugation are given and the possibility of constructing tunable reflection filters is discussed in this chapter.

5.2 Analysis Model

The geometry considered is a dielectric slab waveguide, with doubly grating on both sides, as illustrated in Fig. 5.2. The regions 1 and 3 in the figure are dielectric substrates with index n_1 and n_3 respectively; and the region 2 is the dielectric slab waveguide that is silicon substrate and the refractive index of the silicon is n_2 . The whole structure is uniform in the y direction. The boundary surfaces of the slab at an average thickness d are slightly corrugated sinusoidally in the z direction; while at the boundary surface $x = 0$, the structure is stimulated with lasers periodically which cause the dielectric constant corrugation.

5.2 Analysis Model

Figure 5.1: Flow chart of perturbation method.

5.2 Analysis Model

Figure 5.2: Schematic of the doubly grating structure.

For the surface corrugation, the grating at $x = d$ can be expressed using the following equation:

$$x = u(z) = d[1 + \delta\eta_u \cos(K_u z + \theta)] \quad (5.1)$$

where K_u and η_u are the upper grating vector and index of the undulation of the upper surface. δ is a dimensionless parameter to identify that the index of undulation is much smaller than unity ($\eta_u < 1$).

For the dielectric grating at $x = 0$, dielectric grating is created by laser illumination. The periodically modulated permittivity profile can be expressed by Fourier expansion [10]:

$$\varepsilon(z) = \varepsilon_{av}[1 + \delta\eta_l \cos(K_l z + \theta)] \quad (5.2)$$

that is,

5.2 Analysis Model

Figure 5.3: Photo-excited silicon and its dielectric expressions.

$$n_2^2 = \varepsilon_{av}[1 + \delta\eta_l \cos(K_l z + \theta)] \quad (5.3)$$

where K_l and η_l are the low grating vector and index of the undulation of the lower surface, $x = 0$, and η_l is also much smaller than unity. ε_{av} is the average permittivity over which the permittivity modulation is considered. In this analysis, ε_{av} is taken as the permittivity of silicon substrate (ε_s).

Figure 5.3 shows the details of the photo-excited semiconductor. When the semiconductor is illuminated with the photon energy greater than the band gap energy between valence band and conduction band of the semiconductor, photons are absorbed; and electron/hole pairs are generated. A thin layer plasma region is then formed near the surface of the waveguide. The presence of free carrier/electron-hole plasma results in the modification of conductive and dielectric properties of the semiconductor material according to Drude-Lorentz formula as shown in Eq. (4.1).

As have been mentioned in Chapter 4, real part and imaginary part of permit-

5.3 Formulation of the Problem

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tivity for different optically induced plasma carrier densities are shown in Fig. 4.3.

Eq. (5.2) can be rewritten as below:

$$\varepsilon(z) = \varepsilon_s \left[1 + \frac{\Delta\varepsilon_{pr} + \Delta\varepsilon_{pi}}{\varepsilon_s} \cos(K_l z + \theta) \right] \quad (5.4)$$

5.3 Formulation of the Problem

To solve a problem using perturbation method, one should first expand the fields on multiple space scales $z = \delta^n z_n$, then one should get the differential functions satisfying the fields for each order δ^n , that is, get the Maxwell equations according to each order of δ^n , also the boundary conditions are not the same because of the multiple space scales, so the equivalent boundary condition should be obtained for each order of δ^n . In this case, up to δ^1 scale has been used. So one can now solve the problem δ^0 and δ^1 problem and finally evaluate coupling characteristics of the structure.

The perturbation is carried up to the first order by introducing space scales in the $\hat{z}$ direction and in time t as:

$$z_0 = z \quad (5.5)$$

$$z_1 = \delta z$$

$$t_0 = t \quad (5.6)$$

$$t_1 = \delta t$$

5.3 Formulation of the Problem

The expression of E_y can be expanded as follows:

$$E_y(x; z, t) = E_y^0(x; z_0, z_1, t_0, t_1) + \delta E_y^1(x; z_0, z_1, t_0, t_1) \quad (5.7)$$

where z_1 characterize the slow amplitude modulations of the unperturbed zeroth order field E_y^0 and δE_y^1 is the perturbed first-order field because of the weak surface and dielectric corrugations.

Considering the derivative expressions in expanded form

$$\begin{aligned} \frac{\partial}{\partial z} &= \frac{\partial}{\partial z_0} + \delta \frac{\partial}{\partial z_1} \\ \frac{\partial}{\partial t} &= \frac{\partial}{\partial t_0} + \delta \frac{\partial}{\partial t_1} \end{aligned} \quad (5.8)$$

and equating the coefficients of the same powers of δ , we can get the differential equations as shown below:

$$\begin{aligned} \left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} - \frac{n_l^2}{c^2} \frac{\partial^2}{\partial t^2} \right) E_y^0 &= 0 \\ \left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} - \frac{n_l^2}{c^2} \frac{\partial^2}{\partial t^2} \right) E_y^1 &= -2 \left(\frac{\partial^2}{\partial z_0 \partial z_1} - \frac{n_l^2}{c^2} \frac{\partial^2}{\partial t_0 \partial t_1} \right) E_y^0 \end{aligned} \quad (5.9)$$

In this case, only the transverse electric (TE) waves are considered, which propagate in the $\hat{z}$ direction and are uniform in the $\hat{y}$ direction. Therefore, only three components E_y , H_x and H_z are included in the formulation.

In each region of Fig. 5.2, the field expression E_y satisfies the following wave equations:

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} - \frac{n_l^2}{c^2} \frac{\partial^2}{\partial t^2} \right) E_y = 0 \quad (l = 1, 2, 3) \quad (5.10)$$

5.3 Formulation of the Problem

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where c is the velocity of light in free space, n_l is the index of the materials in region l . At boundary surface $x = u(z)$, it is obviously that E_y should be continuous. Substituting Eq. (5.2) into Eq. (5.10) and letting H_z be continuous in $x = u(z)$, we obtain another boundary condition that $(\frac{\partial}{\partial x}E_y + \delta\eta_l dK_u \sin(K_u z + \theta)\frac{\partial}{\partial z}E_y)$ should be continuous at the boundary surfaces. For the surface $x = 0$, E_y and H_z should be continuous, that is, E_y and $\frac{\partial}{\partial x}E_y$ are continuous.

To solve the zeroth order equation, assuming the solution E_y^0 to be of the form:

$$E_y^0(x; z_0, z_1, t_0, t_1) = \sum_v A_v(z, t) \phi_v^0 \exp[-j(\omega t_0 - \beta_v z_0)] \quad (5.11)$$

where ϕ_v^0 is the zeroth order model function of the v_{th} ($v = a, b, c$) mode which can be solved by simply applying the boundary conditions and the detailed expression of ϕ_v^0 can be found in [86]. $A_v(z_1, t_1)$ is the complex amplitude which is involved with the slow scales z_1 and t_1 . ω is the wave frequency and β_v is the propagation constant of the v_{th} mode.

The zeroth order modes would couple to each other in the presence of the weak surface and dielectric corrugations. It is assumed that the modes a and b are coupled through the upper surface corrugation with the wave number K_u , and that the modes b and c are coupled through the dielectric corrugation with the wave number K_l . Thus the following condition for the first-order Bragg interactions

5.3 Formulation of the Problem

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between the three modes a , b and c should be satisfied:

$$\begin{aligned}\beta_b - \beta_a &= K_u \\ \beta_c - \beta_b &= K_l\end{aligned}\tag{5.12}$$

That is, the mode b is directly coupled to the other two modes a and c .

By solving the first-order field equations, we can obtain the coupled mode equations. The coupled mode equations can determine the dependence of the $A_v(z_1, t_1)$ on the slow scales z_1 and t_1 . Assuming the expressions of the first-order solution for E_y^1 of the following form:

$$E_y^1 = \sum_v \phi_v^1(x) \exp[-j(\omega t_0 - \beta_v z_0)]\tag{5.13}$$

where $\phi_v^1(x)$ denotes the first-order correction to the modal field $A_v \phi_v^0$.

Following the procedure mentioned in [84] and [86], the coupled mode equations are derived as given below:

$$\begin{aligned}\left(\frac{\partial}{\partial t_1} + v_a \frac{\partial}{\partial z_1}\right) A_a &= j e^{j\theta} C_{ab} A_b \\ \left(\frac{\partial}{\partial t_1} + v_b \frac{\partial}{\partial z_1}\right) A_b &= j e^{-j\theta} C_{ab} A_a - j C_{bc} A_c \\ \left(\frac{\partial}{\partial t_1} + v_c \frac{\partial}{\partial z_1}\right) A_c &= -C_{bc} A_b\end{aligned}\tag{5.14}$$

In Eq. (5.14), v_ν is the group velocity of the ν_{th} mode, $C_{\nu\mu}$ is the coupling coefficient between the ν_{th} and μ_{th} modes. The detailed expressions of v_ν and $C_{\nu\mu}$ are derived based on the approach given in [86]. The above equations are

5.3 Formulation of the Problem

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for the asymmetric slab waveguide with doubly corrugations. They give out the general expressions governing the first-order Bragg interactions of three TE modes propagation. For a special case, it is assumed that the three layered structure is symmetric and only take the interaction of the lowest three TE modes into consideration. That is, it has been assumed that mode b is TE_0 mode propagating in the $+\hat{z}$ direction and modes a and c are TE_0 and TE_1 modes propagating in the $-\hat{z}$ direction, respectively. Then Eq. (5.14) can be simplified as the followings [84, 86]:

$$\begin{aligned} \left(\frac{\partial}{\partial t_1} - v_0 \frac{\partial}{\partial z_1}\right) A_0^- &= j e^{j\theta} C_{00} A_0^+ \\ \left(\frac{\partial}{\partial t_1} + v_0 \frac{\partial}{\partial z_1}\right) A_0^+ &= j e^{-j\theta} C_{00} A_0^- - j C_{01} A_1^- \\ \left(\frac{\partial}{\partial t_1} + v_1 \frac{\partial}{\partial z_1}\right) A_1^- &= -C_{01} A_0^+ \end{aligned} \quad (5.15)$$

where

$$\begin{aligned} v_\nu &= \frac{\omega \beta_\nu (2 + dk_{1\nu})}{2\beta_\nu^2 + k_0^2 n_0^2 dk_{1\nu}}, \quad \nu = 0, 1 \\ C_{00} &= \frac{\eta_u \omega}{2} \frac{dk_{10} k_{20}^2}{2\beta_0^2 + k_0^2 n_0^2 dk_{10}} \\ C_{01} &= \eta_l \omega \frac{n_2^2 (k_{10} - k_{11}) k_{20} k_{21} (k_{10} k_{11})^{1/2}}{(n_2^2 - n_1^2)(\beta_0^2 - \beta_1^2)[(2\beta_0^2 + k_0^2 n_0^2 dk_{10})(2\beta_1^2 + k_0^2 n_0^2 dk_{11})]^{1/2}} \end{aligned} \quad (5.16)$$

and the $+$ and $-$ in the superscripts are to denote the forward and backward waves of the TE modes. The wave numbers β_0 and β_1 should satisfy the Bragg condition:

$$\begin{aligned} 2\beta_0 &= K_u \\ \beta_1 + \beta_0 &= K_l \end{aligned} \quad (5.17)$$

5.4 Numerical Results

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and the dispersion equation for the zeroth order is:

$$\tan k_{2\nu}d = \frac{2k_{1\nu}k_{2\nu}}{k_{2\nu}^2 - k_{1\nu}^2} \quad (5.18)$$

If we assume that the modal amplitudes $A_\nu^\pm$ vary in the form $\exp[-j(\Delta\omega t_1 - \Delta\beta z_1)]$, where $\Delta\omega$ indicates the small frequency deviation from the Bragg frequency ω , and following the steps in [86], one can determine the reflection coefficient R_0 of the TE_0 mode, the reflection coefficient R_1 of the TE_1 mode, and the transmission coefficient T of the TE_0 mode. They are given by:

$$\begin{aligned} R_0 &= |A_0^-(z_1 = 0)|^2 = \frac{C_{00}^2}{\nu_0^2} \left| \frac{\Gamma_1}{S} \right|^2 \\ R_1 &= \frac{\nu_1}{\nu_0} |A_1^-(z_1 = 0)|^2 = \frac{C_{01}^2}{\nu_0\nu_1} \left| \frac{\Gamma_0}{S} \right|^2 \\ T &= |A_0^+(z_1 = L)|^2 = \left| \frac{(\Delta\beta_1 - \Delta\beta_2)(\Delta\beta_2 - \Delta\beta_3)(\Delta\beta_3 - \Delta\beta_1)}{S} \right| \end{aligned} \quad (5.19)$$

The detailed expression for Γ_ν , S and $\Delta\beta_\nu$ is found in [86]. The reflection coefficient R_0 and R_1 , and the transmission coefficient T satisfy the following energy conservation relation:

$$R_0 + R_1 + T = 1 \quad (5.20)$$

5.4 Numerical Results

Before the calculating of the doubly perturbed structure, special case of single structure grating on the upper side of the slab waveguide was studied used the Eq. (5.19). To verify the results, they were compared with the results in [86] by yasumoto. The dimensions of the slab waveguide were the same as those in [86].

5.4 Numerical Results

Both results were plotted in Fig. 6.4 and the results matched very well.

Figure 5.4: Comparison of the results of single structural grating on the slab waveguide.

In this chapter, it is considered that a slab waveguide characterized by $dK_0 = 12.0$, $n_2^2/n_1^2 = 11.8$, $K_u/K_0 = 1.0958$, $K_l/K_0 = 1$, where K_0 is an arbitrary standard wave number introduced for the sake of normalization. Thus one can obtain that $\beta_0 = 0.5477$ and $\beta_1 = 0.4470$.

In this case, the index of surface undulation is fixed as 0.05, and the index of dielectric undulation varies from 0 to 0.2. Fig. 5.4(a), Fig. 5.4(b) and Fig. 5.4(c) are the results of the reflection R_0 and R_1 , and transmission coefficient T ,

5.4 Numerical Results

(a)

(b)

5.4 Numerical Results

Figure 5.5: (a) Reflection coefficient R_0 ; (b) Reflection coefficient R_1 and (c) Transmission coefficient T

respectively.

It is seen clearly in the figures that when the dielectric undulation changes, the reflection and transmission coefficients change correspondingly. It can also be observed that this structure exhibits a bandstop characteristic. When the intensity of the laser becomes higher, the index of the dielectric undulation becomes higher. This in turn increases the reflection coefficient of both TE_0 and TE_1 modes and decreases the transmission coefficient in the stop band. The bandwidth of the stopband can also be increased at the same time. Hence it is inferred from the results that it is possible to design a tunable design filter by means of optical

5.5 Conclusion

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excitation. So by controlling optical intensity of the laser sources, one could have tunable feature.

5.5 Conclusion

In the chapter, the characteristics of the three-mode coupling in a dielectric slab waveguide having doubly periodic corrugations have been investigated using perturbation method with multiple scales. The coupled-mode equations governing the first order Bragg interactions of the three propagating TE modes have been given out. The equations are solved for the case where the incident guided wave couples to the coherently reflected two guided waves. For the structure with fixed surface corrugation and variable dielectric corrugation, the results are given out for the reflection coefficients R_0 and R_1 , and the transmission coefficient T . The possibility of fabricating a tunable rejection filter enhanced by optical excitation is mentioned.

Chapter 6

Leaky Wave Analysis on Periodically Photo-Induced Double Grating Structures

6.1 Introduction

As has been discussed in Chapter 5, the optical controlled microwave and millimeter wave devices are gaining more and more interests. This chapter will continue to discuss on a double grating structure with periodic photo induced effects.

A dielectric waveguide with a periodic surface corrugation or permittivity modulation was shown to find applications as leaky wave antennas at millimeter waves and grating couplers used in integrated optics [105]. A large amount of effort was devoted to the corrugated and index modulated periodic structures as reported in [85, 106]. The grating structures used in [85, 106] were permanent configuration and unchangeable, in order to obtain a controllable structure, a semiconductor

6.2 Formulation of the Problem

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material was used in [10] as the substrate of a dielectric waveguide and the grating configuration was created using periodic optical illumination. This illumination could cause permittivity modulation inside the material, which resulted in a controllable grating configuration. In this chapter, to get a higher radiation coefficient with a controllable configuration, a periodical photo induced double grating structures using semiconductor substrate has been analyzed by an asymptotic method of singular perturbation procedure based on multiple space scales.

6.2 Formulation of the Problem

Figure 6.1 shows the schematic of the double grating structure. The substrate used in the structure is semiconductor. In Fig. 6.1, a surface with a weak periodic corrugation is located at $x = d(z)$ which is very near the $x = d$ plane. The region between $x = 0$ and $x = d$ except for the dotted area is occupied by a dielectric film with $(\varepsilon_{av}\varepsilon_0, \mu)$. In region $x > d(z)$ and $x < 0$, it is free space, occupied by air. In the later part of this chapter, an index i is used to indicate the three regions of the structure. $i = 1$ indicates the region $x > d(z)$, $i = 2$ means the region $0 < x < d(z)$ and $i = 3$ means the region $x < 0$. By a Fourier analysis, the periodic function $d(z)$ can be expressed as [10]:

$$d(z) = d(1 + \delta\eta_{u1} \cos(K_1 z + \theta_1) + \delta^2\eta_{u2} \cos(2K_1 z + \theta_2) + \dots) \quad (6.1)$$

The dotted region in Fig. 6.1 is the permittivity modulation region. It is created by the optical illumination, and because of absorption of photon, a thin layer of plasma is formed. The permittivity of the plasma varies with the density of

6.2 Formulation of the Problem

Figure 6.1: Schematic of the double grating structure.

the plasma and the frequency. Since the created area is periodic, the permittivity of the area can be expressed by a Fourier series:

$$\varepsilon(z) = \varepsilon_{av} (1 + \delta \eta_{l1} \cos(K_2 z + \phi_1) + \delta^2 \eta_{l2} \cos(2K_2 z + \phi_2) + \dots) \quad (6.2)$$

In Eq. (6.1) and Eq. (6.2), δ is the smallness parameter. In Eq. (6.1), η_{u1} is the amplitude of the fundamental harmonic of the surface corrugation, K_1 is the grating vector related to the grating period Λ_1 by $K_1 = 2\pi/\Lambda_1$. Electromagnetic waves in a grating region can be represented in terms of space harmonics whose phase constants are:

$$\beta_m = \beta_0 + \frac{2m\pi}{\Lambda_1}, \quad m = 0, \pm 1, \pm 2 \quad (6.3)$$

In Eq. (6.2), β_0 is the phase constant of unperturbed case, and η_{l1} is the amplitude of the fundamental harmonic of the permittivity grating, and K_2 is the grating vector by $K_2 = 2\pi/\Lambda_2$.

6.2 Formulation of the Problem

Figure 6.2: Wave propagation in the structure.

Figure 6.2 depicts out the propagation of the waves in the structure, with the incident waves a_i , the guided waves A_g and radiated waves b_i . The transverse magnetic (TM) mode is considered here with field components of H_y, E_x, E_z , propagating in the $\hat{z}$ direction and having no variation in $\hat{y}$ direction with $\exp(j\omega t)$ time dependence. The perturbation is carried up to the second order by introducing space scales in the $\hat{z}$ direction as:

$$\begin{aligned} z_0 &= z, \\ z_1 &= \delta z, \\ z_2 &= \delta^2 z \end{aligned} \tag{6.4}$$

Thus the expression of H_y is given by:

$$H_y(x, z) = H_{y0}(x, z_0, z_2) + \delta H_{y1}(x, z_0, z_2) + \delta^2 H_{y2}(x, z_0, z_2) \tag{6.5}$$

6.2 Formulation of the Problem

Furthermore:

$$\frac{\partial}{\partial z} = \frac{\partial}{\partial z_0} + \delta^2 \frac{\partial}{\partial z_2} \quad (6.6)$$

Incorporating inhomogeneous variation of permittivity in the propagation direction, the Helmholtz Equation for H_y is written as:

$$\left[\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + \omega \mu \varepsilon_0 \varepsilon(z) \right] H_y - \frac{1}{\varepsilon(z)} \frac{\partial \varepsilon(z)}{\partial z} \frac{\partial H_y}{\partial z} = 0 \quad (6.7)$$

Substituting Eq. (6.2) into Eq. (6.7), we get

$$\begin{aligned} & \left\{ \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + \omega^2 \mu \varepsilon_0 \varepsilon_{av} [1 + \delta \eta_{l1} \cos(K_2 z + \phi_1) + \delta^2 \eta_{l2} \cos(2K_2 z + \phi_2)] \right\} H_y \\ & + \frac{K_2 \delta \eta_{l1} \sin(K_2 z + \phi_1) + 2K_2 \delta^2 \eta_{l2} \sin(2K_2 z + \phi_2)}{1 + \delta \eta_{l1} \cos(K_2 z + \phi_1) + \delta^2 \eta_{l2} \cos(2K_2 z + \phi_2)} \frac{\partial H_y}{\partial z} = 0 \end{aligned} \quad (6.8)$$

Substituting Eq. (6.4) and Eq. (6.5) into Eq. (6.8) and equating the coefficients of equal powers of δ , we get the differential equations for each order,

6.2 Formulation of the Problem

$$\begin{aligned}
 o(\delta^0) : \quad & \left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + \omega^2 \mu \varepsilon_0 \varepsilon_{av} \right) H_{y0} = 0 \\
 o(\delta^1) : \quad & \left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + \omega^2 \mu \varepsilon_0 \varepsilon_{av} \right) H_{y1} = - \eta_{l1} K_2 \sin(K_2 z_0 + \phi_1) \frac{\partial H_{y0}}{\partial z_0} \\
 & \quad - \omega^2 \mu \varepsilon_0 \varepsilon_{av} \eta_{l1} \cos(K_2 z_0 + \phi_1) H_{y0} \\
 o(\delta^2) : \quad & \left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial z^2} + \omega^2 \mu \varepsilon_0 \varepsilon_{av} \right) H_{y2} = - 2 \frac{\partial^2 H_{y0}}{\partial x \partial z_0} - \omega^2 \mu \varepsilon_0 \varepsilon_{av} \eta_{l1} \cos(K_2 z_0 + \phi_1) H_{y1} \\
 & \quad - \omega^2 \mu \varepsilon_0 \varepsilon_{av} \eta_{l2} \cos(2K_2 z_0 + \phi_2) H_{y0} \\
 & \quad - \eta_{l1} K_2 \sin(K_2 z_0 + \phi_1) \frac{\partial H_{y1}}{\partial z_0} \\
 & \quad - 2 \eta_{l2} K_2 \sin(2K_2 z_0 + \phi_2) \frac{\partial H_{y0}}{\partial z_0} \\
 & \quad + \frac{K_2 (\eta_{l1})^2}{2} \sin 2(K_2 z_0 + \phi_1) \frac{\partial H_{y0}}{\partial z_0}
 \end{aligned} \tag{6.9}$$

The boundary conditions in this problem are that H_y and E_x must be continuous at $x = 0$ and $x = d$ for each order δ_n .

The boundary conditions of each order of at $x = 0$ are expressed as:

$$o(\delta^0) : H_{y20} = H_{y30} \tag{6.10}$$

$$\frac{\partial H_{y20}}{\partial x} = \varepsilon_{av} \frac{\partial H_{y30}}{\partial x}$$

6.2 Formulation of the Problem

$$o(\delta^1) : H_{y21} = H_{y31} \quad (6.11)$$

$$\frac{\partial H_{y21}}{\partial x} - \eta_{l1} \cos(K_2 z_0 + \phi_1) = \varepsilon_{av} \frac{\partial H_{y31}}{\partial x}$$

$$o(\delta^2) : H_{y22} = H_{y32}$$

$$\frac{\partial H_{y22}}{\partial x} - \eta_{l2} \cos(2K_2 z_0 + \phi_2) \frac{\partial H_{y20}}{\partial x} - \eta_{l1} \cos(K_2 z_0 + \phi_1) \frac{\partial H_{y21}}{\partial x} = \varepsilon_{av} \frac{\partial H_{y32}}{\partial x} \quad (6.12)$$

The boundary conditions of each order of δ at $x = d$ are expressed as:

$$o(\delta^0) : H_{y10} = H_{y20} \quad (6.13)$$

$$\frac{\partial H_{y20}}{\partial x} = \varepsilon_{av} \frac{\partial H_{y10}}{\partial x}$$

$$o(\delta^1) : H_{y21} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial H_{y20}}{\partial x} = H_{y11} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial H_{y10}}{\partial x}$$

$$\begin{aligned} & \frac{\partial H_{y21}}{\partial x} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y20}}{\partial x^2} - \eta_{l1} \cos(K_2 z_0 + \phi_1) \frac{\partial H_{y20}}{\partial x} \\ & = \varepsilon_{av} \left(\frac{\partial H_{y11}}{\partial x} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y10}}{\partial x^2} \right) \end{aligned} \quad (6.14)$$

6.2 Formulation of the Problem

$$\begin{aligned}
o(\delta^2) : & H_{y22} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial H_{y21}}{\partial x} + \frac{1}{2} d^2 (\eta_{u1})^2 \cos^2(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y20}}{\partial x^2} \\
& + d\eta_{u2} \cos(2K_1 z_0 + \theta_2) \frac{\partial H_{y20}}{\partial x} \\
= & H_{y12} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial H_{y11}}{\partial x} + \frac{1}{2} d^2 (\eta_{u1})^2 \cos^2(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y10}}{\partial x^2} \\
& + d\eta_{u2} \cos(2K_1 z_0 + \theta_2) \frac{\partial H_{y10}}{\partial x} \\
\\
& \frac{\partial H_{y22}}{\partial x} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y21}}{\partial x^2} + \frac{1}{2} d^2 (\eta_{u1})^2 \cos^2(K_1 z_0 + \theta_1) \frac{\partial^3 H_{y20}}{\partial x^3} \\
& + d\eta_{u2} \cos(2K_1 z_0 + \theta_2) \frac{\partial^2 H_{y20}}{\partial x^2} - \eta_{l2} \cos(2K_2 z_0 + \phi_2) \frac{\partial H_{y20}}{\partial x} \\
& - \eta_{l1} \cos(K_1 z_0 + \theta_1) \left[\frac{\partial H_{y21}}{\partial x} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y20}}{\partial x^2} \right] \\
= & \varepsilon_{av} \left(\frac{\partial H_{y12}}{\partial x} + d\eta_{u1} \cos(K_1 z_0 + \theta_1) \frac{\partial^2 H_{y11}}{\partial x^2} + \frac{1}{2} d^2 (\eta_{u1})^2 \cos^2(K_1 z_0 + \theta_1) \frac{\partial^3 H_{y10}}{\partial x^3} \right. \\
& \left. + d\eta_{u2} \cos(2K_1 z_0 + \theta_2) \frac{\partial^2 H_{y10}}{\partial x^2} \right)
\end{aligned} \tag{6.15}$$

Once solving the zeroth-order, first-order and second order problems, one can get the dispersion characteristics, the relations between the radiated wave, the guided wave and the incident wave, and the amplitude transport equation. Using these results, radiation coefficient can be calculated.

6.2 Formulation of the Problem

6.2.1 Zeroth Order Problem

The zeroth order fields are referred to as the unperturbed guided wave. The zeroth order fields in the slab are given as:

$$H_{y10} = N_g A_g(z_2) \left(\frac{\alpha_{1i}}{k_i} \sin k_i d + \cos k_i d \right) e^{-\alpha_{1i}(x-d)} e^{-j\beta_i z_0}, \quad (x > d) \quad (6.16)$$

$$H_{y20} = N_g A_g(z_2) \left(\frac{\alpha_{1i}}{k_i} \sin k_i x + \cos k_i x \right) e^{-j\beta_i z_0}, \quad (0 < x < d) \quad (6.17)$$

$$H_{y30} = N_g A_g(z_2) e^{-\alpha_{1i} x} e^{-j\beta_i z_0}, \quad (x < 0) \quad (6.18)$$

And,

$$\begin{aligned} k_i &= (\omega^2 \mu_0 \varepsilon_0 \varepsilon_{av} - \beta_i^2)^{1/2} \\ \alpha_i &= (\beta_i^2 - \omega^2 \mu_0 \varepsilon_0)^{1/2} \end{aligned} \quad (6.19)$$

Where β_i is the zeroth order propagation constant in the $\hat{z}$ direction. The zeroth order dispersion is given by

$$2\alpha_{1i} k_i \cos k_i d = (k_i^2 - \varepsilon_{av} \alpha_{1i}^2) \sin k_i d \quad (6.20)$$

6.2.2 First Order Problem

The perturbation first order fields are caused by the originally excited wave due to corrugations. The propagation constants of the first order fields are $\beta_i - K_1$, $\beta_i + K_1$, $\beta_i - K_2$ and $\beta_i + K_2$ and are restricted to the first order space harmonics.

6.2 Formulation of the Problem

The first order solution can be expressed as a combination of four scattered Floquet modes for each region.

$$\begin{aligned}
 H_{y11} = & N_r e^{j\theta_1} (\alpha_{i1} e^{jk_{-1c}(x-d)} + b_{i1} e^{-jk_{-1c}(x-d)}) e^{-j(\beta_i - K_1)z_0} + F_{1c} e^{-\alpha_{1c}(x-d)} e^{-j(\beta_i + K_1)z_0} \\
 & N_r e^{j\phi_1} (\alpha_{i2} e^{jk_{-2c}(x-d)} + b_{i2} e^{-jk_{-2c}(x-d)}) e^{-j(\beta_i - K_2)z_0} + F_{2c} e^{-\alpha_{2c}(x-d)} e^{-j(\beta_i + K_2)z_0} \\
 & (x > d)
 \end{aligned} \tag{6.21}$$

$$\begin{aligned}
 H_{y21} = & (F_{-1f} \frac{\cos k_{-1f}x}{\cos k_{-1f}d} + G_{-1f} \frac{\sin k_{-1f}x}{\sin k_{-1f}d}) e^{-j(\beta_i - K_1)z_0} \\
 & + (F_{1f} \frac{\cos k_{1f}x}{\cos k_{1f}d} + G_{1f} \frac{\sin k_{1f}x}{\sin k_{1f}d}) e^{-j(\beta_i + K_1)z_0} \\
 & (F_{-2f} \frac{\cos k_{-2f}x}{\cos k_{-2f}d} + G_{-2f} \frac{\sin k_{-2f}x}{\sin k_{-2f}d} + C_1 \frac{\frac{\alpha_{1i}}{k_i} \sin k_i x + \cos k_i x}{2\beta_i K_2 - K_2^2}) e^{-j(\beta_i - K_2)z_0} \\
 & (F_{2f} \frac{\cos k_{2f}x}{\cos k_{2f}d} + G_{2f} \frac{\sin k_{2f}x}{\sin k_{2f}d} + C_2 \frac{\frac{\alpha_{1i}}{k_i} \sin k_i x + \cos k_i x}{-2\beta_i K_2 - K_2^2}) e^{-j(\beta_i + K_2)z_0} \\
 & (0 < x < d)
 \end{aligned} \tag{6.22}$$

$$\begin{aligned}
 H_{y31} = & N_r e^{j\theta_1} (A_i e^{jk_{-1c}x} + B_i e^{-jk_{-1c}x}) e^{-j(\beta_i - K_1)z_0} + F_{1a} e^{\alpha_{1c}x} e^{-j(\beta_i + K_1)z_0} \\
 & N_r e^{j\phi_1} (A_i e^{jk_{-2c}x} + B_i e^{-jk_{-2c}x}) e^{-j(\beta_i - K_2)z_0} + F_{2a} e^{\alpha_{2c}x} e^{-j(\beta_i + K_2)z_0} \\
 & (x < 0)
 \end{aligned} \tag{6.23}$$

6.2 Formulation of the Problem

In the above equations,

$$\begin{aligned} C_1 &= \frac{-\omega\mu\varepsilon_0\varepsilon_{av}\eta_{l1} + \eta_{l1}K_2\beta_i}{2} N_g A_g e^{j\phi_1} \\ C_2 &= \frac{-\omega\mu\varepsilon_0\varepsilon_{av}\eta_{l1} - \eta_{l1}K_2\beta_i}{2} N_g A_g e^{-j\phi_1} \end{aligned} \quad (6.24)$$

The propagation constants are given by:

$$\begin{aligned} k_{-1c} &= (\omega^2\mu_0\varepsilon_0 - (\beta_i - K_1)^2)^{1/2} \\ \alpha_{1c} &= ((\beta_i + K_1)^2 - \omega^2\mu_0\varepsilon_0)^{1/2} \\ k_{-1f} &= (\omega^2\mu_0\varepsilon_0\varepsilon_{av} - (\beta_i - K_1)^2)^{1/2} \\ k_{1f} &= (\omega^2\mu_0\varepsilon_0\varepsilon_{av} - (\beta_i + K_1)^2)^{1/2} \end{aligned} \quad (6.25)$$

and

$$\begin{aligned} k_{-2c} &= (\omega^2\mu_0\varepsilon_0 - (\beta_i - K_2)^2)^{1/2} \\ \alpha_{2c} &= ((\beta_i + K_2)^2 - \omega^2\mu_0\varepsilon_0)^{1/2} \\ k_{-2f} &= (\omega^2\mu_0\varepsilon_0\varepsilon_{av} - (\beta_i - K_2)^2)^{1/2} \\ k_{2f} &= (\omega^2\mu_0\varepsilon_0\varepsilon_{av} - (\beta_i + K_2)^2)^{1/2} \end{aligned} \quad (6.26)$$

The presence of periodic corrugation renders at least one of the scattered Floquet modes to be a fast wave in order to obtain leaky wave phenomena. This can be accomplished by appropriately choosing K_1 and K_2 such that only $\beta_i - K_1$ and $\beta_i - K_2$ lie in the fast wave region, and all other Floquet modes are slow waves.

From the boundary conditions of the first order fields, a relation between the amplitudes of the principal guided wave A_g , the incident wave a_{i1} and a_{i2} , and the

6.2 Formulation of the Problem

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radiated wave b_{i1} and b_{i2} are given by:

$$b_{i1} = C_{rg1}A_g + C_{rr1}a_{i1} + C_{rr2}a_{i2} \quad (6.27)$$

$$b_{i2} = C_{rg2}A_g + C_{rr3}a_{i1} + C_{rr4}a_{i2} \quad (6.28)$$

where C_{rg1} and C_{rg2} are the coupling coefficients for b_{i1} and b_{i2} , C_{rr1} and C_{rr2} are reflection coefficients at the top and bottom surfaces.

6.2.3 Second Order Problem

The analysis of the second order fields determines the interaction between the principle guided wave and the first order incident and reflected waves. The particular solution for the second order problem is assumed to be forms of:

$$H_y = \phi_i(x)e^{-j\beta_i z_0}, \quad i = 1, 2, 3 \quad (6.29)$$

Substituting Eq. (6.29) into Eq. (6.15), the solution for $\phi_i(x)$ can be obtained, and from the boundary condition of the second order field at $x = 0$ and $x = h$, an amplitude transport equation is obtained as:

$$\frac{\partial A_g}{\partial z_2} = C_{gg}A_g + C_{gr1}a_{i1} + C_{gr2}a_{i2} \quad (6.30)$$

Eq. (6.27), Eq. (6.28) and Eq. (6.30) constitute a pair of canonical equations relating guided wave, incident waves, and radiated waves. C_{gg} is called the extinction coefficient whose real part is the leakage coefficient, C_{ggr} . The imaginary

6.3 Numerical Results

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part of the extinction coefficient, C_{ggi} , determines the exact radiation angle for optimum radiation efficiency.

6.3 Numerical Results

The zeroth order fields are referred as the unperturbed guided wave. As mentioned in Eq. (6.19), β_i is the zeroth order propagation constant in the $\hat{z}$ direction, α_{1i} and k_i are the propagation constant in the $\hat{x}$ direction in region 1 and region 3 respectively, hence the dispersion equation is obtained as stated in Eq. 6.20.

The dispersion characteristics are shown in Fig. 6.3. In the calculations, the thickness of the film is 1 mm and Λ_1 is 6 mm and Λ_2 is 5 mm. It is clearly seen from the diagram, from 24.3 GHz to 29 GHz, $\beta_i - K_2$ is in the fast wave region, but $\beta_i - K_1$ is still in the slow wave region; when the frequency is above 29 GHz, $\beta_i - K_2$ and $\beta_i - K_1$ are both in the fast wave region.

Since the experiments could not be carried out because of lacking multiple laser source and high resistivity semiconductor samples, the validation of the equations in Section 6.2 was tested by considering the specific cases with single surface grating structure and comparing the results with those in [107]. The proposed structure in [107] was single structural corrugations in optical frequency domain, however this analysis is applicable in that regime too. Other parameters used in Fig. 6.4 are the same as used in [107]. The comparison of the leakage coefficients are shown in Fig. 6.4. The α in the figure is corresponding to the C_{ggr} in Eq. (6.30). It is found

6.3 Numerical Results

Figure 6.3: Dispersion diagram for TM mode in the structure.

that the results are matched with single grating case which proved the validity of the results generated in this work.

As mentioned in Eq. (6.30), C_{ggr} is the leakage coefficient. In Fig. 6.5, C_{ggr} varies with the density of the plasma, indicating by η_{l1} . In the calculation, η_{u1} is fixed as 0.05. We can see from the figure, as the plasma density increased, the leakage coefficient also increased.

6.3 Numerical Results

Figure 6.4: Comparison of the single structural grating results.

In a finite length L of the grating structure, the radiation efficiency Q_0 is defined as the ratio of the total power radiated from the modulated region to the guided wave power incident at $z_2 = 0$ and is expressed as:

$$Q_0 = \frac{\int_0^L (|b_{i1}|^2 + |b_{i2}|^2) dz_2}{|A_g|_{z_1=0}^2} = 1 - e^{2C_{ggr}L} \quad (6.31)$$

The radiation efficiency of different plasma density is shown in Fig. 6.6. In the calculation, the whole length is 50 mm and η_{u1} is 0.05. Fig. 6.5 indicates the radiation efficiency increases with the increasing of the plasma density.

Figure 6.7 shows how the radiation coefficient varies with the ratio of the period

6.3 Numerical Results

Figure 6.5: Variation of C_{ggr} with plasma density η_{l1} . ($\eta_{u1} = 0.05$)

of the upper and lower corrugations in this doubly grating structure. K_1 and K_2 are grating vectors related to the grating periods. K_1 is $2\pi/\Lambda_1$ and K_2 is $2\pi/\Lambda_2$. The working frequency is 40 GHz. It can be seen from the figure that when the period of the two grating is about 0.25, the radiation coefficient achieve the maximum.

The exactly radiation angle was determined by the imaginary part of the extinction coefficient C_{gg} . The expression is shown in Eq. (6.32).

$$\theta_r = \frac{k_{1c}}{\beta_i - K_1 + C_{ggi}} \quad (6.32)$$

The radiation angle of the upper side of the waveguide slab is shown in Fig.

6.4 Conclusions

Figure 6.6: Variation of radiation efficiency with plasma density η_{l1} . ($\eta_{u1} = 0.05$)

6.8. The figure shows that the angle varied from 45 to 60 degree when the frequency scanned from 30GHz to 40GHz.

6.4 Conclusions

The leaky wave characteristics on a photo induced double grating silicon slab are analyzed rigorously by singular perturbation method based on multiple space scales. The leakage coefficient and the radiation efficiency are given out numerically and these radiation characteristics are investigated as a function of optically induced plasma density. The optical excitation enhances considerably the radia-

6.4 Conclusions

Figure 6.7: Variation of radiation efficiency with grating vector K_2/K_1 . ($f = 40GHz$)

tion efficiency and also gives flexibility in controlling the radiation behavior. The radiation efficiency of 90% in the analyzed results indicate that it is possible to design leaky wave structures based on doubly corrugated semiconductor waveguide.

6.4 Conclusions

Figure 6.8: Radiation angle of the upper side of the waveguide slab.

Chapter 7

Conclusion and Recommendations

7.1 Conclusion

A thorough study of periodic structures and optically controlled structures have been conducted in this work. Firstly in Chapter 3, photonic bandgap structures have been discussed and the split ring resonator pattern has been introduced. As has been investigated in this thesis, when this special CSRR pattern etched periodically on the ground plane of a microstrip line, the band stop characteristics appeared. The design has been simulated in ADS 2003a and an impressive up to 60 dB insertion loss is achieved. This property is very useful for constructing high insertion loss bandstop (band reject) filters, especially in the case of filtering strong jamming interference to remove large amplitude noises from the signals. After that, the propagation characteristics have been computed and discussed. A good understanding on the propagation constant and attenuation of the whole

7.1 Conclusion

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structure are presented. Parameter studies give out that the resonant frequency is independent of the period of the design. In order to appreciate the advantage of the compact size of the SRR pattern, a 3-cell microstrip line filter with CSRR slots has been designed and tested. Excellent agreement has been observed for the cases considered.

In the later part of Chapter 3, the split ring resonator has been extended into defect ground plane structure to widen the application of the unique SRR pattern. Combined SRR with dumbbell like DGS pattern alternatively introduces an additional pole comparing to the conventional DGS only pattern. So microstrip lines with DGS-SRR and DGS only pattern etched on the ground plane has been designed and simulated, respectively. The simulated results show spurious response in the stopband of DGS-SRR design and is efficiently suppressed comparing with DGS only design and the bandwidth of DGS-SRR design is significantly improved. Further study on the period shows that the size of DGS-SRR design can be reduced by 25% without compromising on the performance. As the rejection band of the DGS-SRR structure is from 3GHz to 10GHz, this useful characteristic can be exploited in suppressing the interference from UWB bands in the coexistence scenario of mobile, WLAN and UWB systems.

After that, optical controlled semiconductor has been introduced to the periodic structures in Chapter 4. The permittivity properties of silicon when it is illuminated by different intensities of laser has been investigated and computed. The real and imaginary parts of the permittivity vary according to the plasma densities and also the frequency dependence is considered. With the understanding

7.1 Conclusion

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of this special characteristic, the microstrip line structures loaded with different PBG patterns on the ground plane with silicon substrate have been designed. The proposed configuration can be used without optical excitation as it displays filtering characteristic because of EBG resonators, but for the tunable behavior, optical excitation is needed. Different kinds of PBG patterns have been designed and simulated both with and without illumination of the laser. Those results have been reported in Chapter 4. Among which, the split ring resonator pattern can achieve 80 *dB* insertion loss at 18 *GHz* without illumination, while around only 1*dB* insertion loss when the illuminating laser plasma density is up to $10^{22}/m^3$. The possibility of fabricating an on-off optical controlled filter has been proposed by applying this design. At 18 *GHz*, when the filter is not illuminated, the insertion loss is high, the filter is at on status; and when the filter is illuminated, the insertion loss is only 1 *dB*, the filter is at off status. This kind of on-off filter is switched at pico-second response and has a high isolation in microwave circuits since the controlling signal is optical. To perform the experiments, we need high resistivity silicon ($5000\Omega \cdot cm$) and multiple LEDs/lasers for periodical illuminations, which are currently not available at our laboratory. However, the results investigated show a clear promise and potential of such a high on-off switch. To verify the simulated and numerical results, an alternative simulation CST has been introduced and thereby the results are compared to show the validity.

To further investigate into the optical controlled periodic structure, a singular perturbation method is used to investigate the doubly grating structures. In Chapter 5, the characteristics of three-mode coupling in a dielectric slab waveguide having doubly periodic corrugations have been investigated. This slab consists of

7.1 Conclusion

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silicon, has structural grating on the upper side, and on the bottom, the permittivity periodically controlled with laser illumination. Three propagating TE modes are considered in the analysis. By solving the first order problem in perturbation method, the coupled-mode equations governing the first order Bragg interactions of the three modes have been given out and the reflection coefficients R_0 and R_1 , and the transmission coefficient T of these modes are computed. A tunable rejection filter enhanced by optical excitation can be achieved by this design. Furthermore, the leaky wave characteristics of this optical controlled slab has been analyzed rigorously by singular perturbation method based on multiple space scales. The solution up to second order problem of perturbation method has been given out. The leakage coefficient and the radiation efficiency of the slab due to the double corrugations are presented numerically and these radiation characteristics are investigated as a function of plasma density. The radiation efficiency is enhanced by the increasing optical induced plasma density. It is suitable for constructing leaky wave antennas since the radiation efficiency could be up to 90% with proper optical control. To verify the results of these two chapters, comparison are made with the results in specific cases with reported literatures.

In conclusion, novel SRR structures and optical controlled periodic structures have been extensively studied in this thesis. With the property of compact size and light weight compared to conventional PBG structures, SRR loaded microstrip lines and optical controlled doubly grating dielectric waveguide slab are suitable candidates for various applications, especially filters and optical enhanced antennas in modern wireless communications systems.

7.2 Recommendations for Further Research

7.2 Recommendations for Further Research

In order to improve the performance and convenience of the optical controlled devices, a combination of laser beam emitter and the devices themselves can be employed to overcome the power loss from the laser emitter to the silicon substrate. The integration can efficiently reduce the size of the whole system and a large number of the period can be applied to demonstrate the advantage of compact size and high insertion loss.

There is a lot of research on the multiple functionality in one device. Some new abbreviation have been created such as filtenna, which means a filtering antenna, that is, the device has the functional of bandpass filter and horn antenna [108]. The periodic structure also can be applied and designed to accomplish these kinds of multiple functional devices. The future research on the amplifier and antenna, or amplifier and filter combined devices is of great interest. It is another area that progress can be made with the appropriate periodic structure in microwave/millimeter wave domain where multi-functional devices can be made of high potential.

Author's Publications

Journal Papers

1. Y. Xu and A. Alphones, "Propagation characteristics of complimentary split ring resonator (CSRR) based EBG structure," *Microwave and Optical Tech. Lett.*, vol. 47, no. 5, pp. 409–412, Dec. 2005.

Submitted and under Review

2. Y. Xu and A. Alphones, "Novel DGS-SRR based microstrip low pass filter," *International Journal of Ultra Wideband Communications and Systems*. (under review)
3. Y. Xu and A. Alphones, "Analysis on photo-induced double grating periodic structures by perturbation method," *Microwave and Optical Tech. Lett.* (under review)

Conference Papers

1. L.K. Arnaud, Y. Xu, D. Bajon, J.C. Mollier and A. Alphones, "Optically controlled CPS line/microstrip split ring PBG switches: ON/OFF ratio enhancement," in *Proc. Asia-Pacific Microwave Conference*, New Delhi, Dec. 2004.
2. Y. Xu and A. Alphones, "Design of microstrip line based split ring PBG

7.2 Recommendations for Further Research

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structures,” in *Proc. IEEE International Workshop on Small Antennas and Novel Metamaterials*, , pp.426–430 , Singapore, March. 2005.

3. Ying Xu, A. Alphones, Z. Shen , “Analysis on periodic photo-induced double grating structures by multiple space scales,” in *Proc. Asia-Pacific Microwave Conference*, vol. 2, Suzhou, China, Dec. 2005.
4. Y. Xu and A. Alphones, “Three-mode coupling in an optically excited doubly periodic structures”, in *Asia-Pacific Microwave Conference*, HongKong, Dec. 2008.

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