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Photoconductive Switching using Silicon and its Applications in Antennas and Reconfigurable Metallodielectric Electromagnetic Band Gap (EBG) structures

by

Alford Chauraya, BTech, MSc Eng

A DOCTORAL THESIS

Submitted in Partial Fulfilment of the requirements for the award of the degree of Doctor of Philosophy by Loughborough University

July 2006

DEPARTMENT OF ELECTRONIC AND ELECTRICAL ENGINEERING
Loughborough University
United Kingdom

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To my late parents and late brothers
The aims of this research work were to investigate the microwave properties of photoconductive semiconductor switches (PCSS), and how the properties might be used to optically control microwave and millimetre wave devices. Tunable devices (such as antennas, filters and metamaterials) have the ability to increase flexibility performance in multiband systems for example. In this thesis the performance of microwave switches from microstrip discontinuities, with high resistivity silicon dice placed cross the gaps were investigated. Under optical illumination, the electrons in silicon can be excited from the valence band to the conduction band. This photoconductivity in Silicon has been employed to design a small microwave switch that can be operated using optical signal.

The optically activated switch offers a wide range of applications. Potential applications have been demonstrated in integrating the microswitch in microstrip patch antenna, microstrip couple line filter, and Electromagnetic Band Gap (EBG) structures.

Keywords: Photoconductive switching, Silicon, Optically activated Switches, Switchable filters, antennas, EBG phase shifters and reconfigurable EBG structures.
PUBLICATIONS FROM THE RESEARCH WORK


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CHAPTER ONE

1.0 Introduction to optically controlled devices

1.1 Introduction to photoconductive semiconductor switches (PCSS)

This thesis is concerned with the investigation of the optical control of various microwave circuits using photoconductive semiconductor switches (PCSS), and how the photoconducting properties might be used to optically switch or tune microwave and millimetre wave structures. The ability to control the devices' performance depends largely on the transformation of the semiconductor's conductivity when exposed to light of a certain wavelength and flux density. An overview of the optical processes that occur within a photoconductor with specific reference to silicon, (Si) is also covered in this thesis. The photoconductivity effect in Si can be exploited to
switch various microwave circuits on and off, and to control the propagation of Electromagnetic (EM) waves.

In this thesis the design and performance of small and compact microwave Si switches was investigated. The microswitch is considered as the foundation and bedrock component for this work. The optically operated switch proposed in this thesis has the ability to be integrated into most of today’s communication systems, such as reconfigurable smart antennas, tunable metamaterials, and some novel filters. The switch was based upon microstrip gap underneath a high resistivity silicon dice. The main thrust of the work is on investigating potential microwave applications of silicon as a photoconductor in order to be employed a practical optically activated switch that can find various applications in modern communication systems.

Research work on the topic of photoconductive switches was initially conducted and first demonstrated in silicon by Auston as early as 1975 [1]. Approximately two decades later, in 1989, Anderson and Sverre described a modified interdigitated gap photoconductive microwave switches [2]. It has been reported that the interdigitated gap approach reduced the effective gap capacitances and thereby improved the switch performance. The switch was operated by illuminating the gap using a high power GaAlAs/GaAs semiconductor laser emitting light at 805nm, and a maximum “on” state conductance of 5mS and a corresponding transmission of −10dB was measured. The paper by Anderson et al [2], concluded by saying that although the measurements had been carried out of a In : Fe based device, the results were applicable to similar devices based on other high resistivity materials. Similar work was carried by Horri and Tsutsumi [3] where it was demonstrated that the scattering characteristics of a
microstrip line containing an optically controlled interdigital gap had superior characteristics over the commonly used single straight gap.

The investigation on lumped element equivalent circuit of the gap on silicon substrate was also pursued by Gevorgian [4]. The equivalent circuit was derived in order to analyse the losses of the optical power and to characterise the microwave performance of semiconductor microstrip gaps. The equivalent circuit can be a useful tool in the optimisation of the switch performance. Such optimisation allows switch system design trade off between the desired optical drive level and the acceptable power loss in the switch components.

Optically controlled microwave switches are suitable for connecting microwave elements. The investigation of integrating photoconductive microswitches in microwave devices was also pursued by Panagamuwa, where a pair of optically operated microswitches have been employed to create a frequency tunable antenna [5]. Optical illumination from a laser diode via optic cables was focused on two Si switches placed on small gaps in both dipole arms equidistant from the centre feed. With the light switched on, the dipole arms are extended and the antenna resonates at a lower frequency. The results from [5] showed that the antenna’s centre frequency could be shifted by almost 40% when both microswitches are activated simultaneously, and operating each switch individually results in a near 50° shift in beam nulls.
1.2 Potential Applications

The transformation of the steady state behaviour of various microwave devices to active state has significantly broadened the applications of such devices in communication systems. During the past few years, it has become apparent that optically operated switches can be applied to create RF devices and circuits with significant advantages compared to their conventional counterparts. RF switches have a variety of applications in wireless communication industry. The implementation of microwave switches in devices to control its performance has not only produced cutting edge smart products, but has also created employment opportunities as new businesses and products are developed.

The photoconductivity of Silicon can be used to optically control Electromagnetic Band Gap (EBG) based structures. Recently there has been growing interest in studying the use of these materials, as a means of manipulating the electromagnetic behaviour of devices. This is due to the fact that EBG structures have unique frequency selective properties and vast number of applications in microwave, millimetre wave and optical devices. The concept of photonic crystals (or simply periodic structures) was first introduced by Yablonovitch in 1987, and soon attracted considerable interest due to their ability to be engineered for use in many different fields, including communication systems [6]. The number of publications reporting the possible applications has significantly increased over the years. Electromagnetic Band-Gap (EBG) materials, or Photonic Band-Gap (PBG) materials are found to have the unique property that within certain bands of frequency, electromagnetic waves cannot propagate through the structure [7-9]. EBG structures exhibit electromagnetic
responses which are not generally found in nature. The term "Electromagnetic Band Gap" (EBG) is now been accepted as better terminology for these structures than "Photonic Band Gap," (PBG) when applied in the microwave regime. The term EBG is more appropriate since structures can be fabricated for use in any part of the electromagnetic (EM) spectrum. The term PBG has been used to describe microwave phenomena, which occur within EBG and has resulted in some controversy. EBG structures not only exhibit a forbidden range of frequencies, (or a band gap) in which the electromagnetic waves cannot propagate, but also have many technological applications within the fields of optical and microwave communications, including applications in laser and detector technology [10]. There are also several applications that employ EBG structures. EBGs can easily be integrated into microstrip lines based filters [11-13], and can find application as microwave components [14], and as substrates for printed antenna structures [15-20]. Most recently attention has been focused on the elimination of surface-wave formation in planar antennas, and realisations of Artificial Magnetic Conductors (AMCs). A number of numerical techniques, including finite difference methods, finite element methods, spectral domain analysis, and integral equation (IE)/moment methods (MoM), have been used to increase understanding of the properties of surface and leaky waves on such layered, periodic structures.

Although the level of research into Electromagnetic Band Gap (EBG) structures has grown rapidly over the years, the majority of research work has been focused on developing steady state structures, and this thesis discuses some examples of EBG structures integrated with optical microswitches. The rapidly growing communication market demands novel and smart devices. Vardaxoglou and his team covered
considerably the subject on optical control of microwaves [21-31]. Their work made use of the variable (complex) dielectric behaviour of silicon to switch on or off the passband of a filtering response of a Frequency Selective Surface (FSS). Vardaxoglou et al explored the optical excitation on silicon wafers for generation and switching of FSS for a plane wave incident. Figure 1, adapted from [29] shows how an inductive grid array was generated by fixing an optical mask between a light source and a single silicon wafer.

![Figure 1. An array generation using an optical mask. (Adapted from [29])](image)

The optical control of EBG structures are investigated with a focus that they could be utilised to enhance the performance of antenna's systems, and reconfigurable stop band characteristics. Investigations into the applications of EBG in reconfigurable
antennas, filters, and resonators are discussed in this thesis. Proof-of-concept EBG structures were fabricated and measured. Comparison of the measured results of devices integrated with EBGs and the conventional circuits revealed that the presence of the EBGs improved the performance of device.

Potential applications of optically operated switches discussed in this thesis can be in the topic of metamaterials. The integration of EBG and Si microswitches give rise to EBG characteristics can be switched on or off by optical means. For example, one may alter the conducting element dimensions, L1 and L2 by using Si switches as illustrated in Figure 1.2. Operational frequency can be shifted by extending the dimensions of the elements.

![Integration of optically activated switches in metamaterials](image)

Figure 1.2. Integration of optically activated switches in metamaterials
The ability to control the propagation of electromagnetic waves within the band gap by including periodic defects within the structure make EBGs very attractive for the fabrication of devices in communication systems. Tuning the EBG structure was demonstrated by changing the size of the defect. The EBG structures were realised by placing a transmission line above a periodic array of dipoles printed onto a grounded dielectric substrate. This is as an alternative to the conventional method of EBG manufacture, which involves either drilling holes in the dielectric substrates or etching periodic patterns in the ground plane. The traditional methods of EBG fabrication can present a major drawback applications that do not require the disturbance of the ground plane such in realisations of MMICs [25, 32].

Further examples of applications of optically controlled switches were demonstrated in the following devices:

1. frequency switchable filter, [33]
2. optically operated phase shifter [34] and
3. optically controlled microstrip patch antenna [35].

These optical reconfigurable approaches increase the design flexibility of microwave structures to an extend that was previously impossible and is promising for the future application and development. From the literature in [33], a frequency switchable microstrip filter for microwave applications was realised by modifying a single pole parallel-coupled line filter, and the wave propagation in the filter is controlled by varying the structural parameters. Silicon switches were employed to alter the dimensions of the filter, resulting in a reconfigurable centre frequency device.
An simple and cheap optically controlled EBG phase shifter has been proposed [34]. The main beam of an antenna array can be steered by using low cost variable EBG phase shifter module that may contain a number of microswitches. Cheap and low power optical sources such as LEDs can be used to alter the phase shift range. The EBG phase shifter presented in this thesis provide good electromagnetic compatibility (EMC), and thus, can be considered as an alternative to ferrite or solid state phase shifters [36, 37].

An optically switched microstrip patch antenna prototype has be realised by implementing a small Si switch encapsulated along the transmission [35]. In this configuration frequency switching was successfully demonstrated by localised illumination. The predominant advantages of using microswitch and localised illumination are that the problem of diffusion in silicon and the optical power level are greatly reduced.
1.3 Thesis Outline

The six chapters in this thesis are organised as follows:-

Chapter 1 presents an outline of the work in the thesis, and serves as the main introduction of the research work described in Chapters 2, 3, 4 and 5. A summary of the background work relevant to the subject of optically controlled devices is discussed. Potential applications of optically activated switches in communication systems is also highlighted.

In Chapter 2 issues related to photoconduction in semiconductors, and in particular silicon are discussed in order to provide a clear understanding of the theory related to specific cases in this work.

Chapter 3 covers the designs of optically operated microwave switches. The electromagnetic properties of these silicon-based switches were investigated. Three switch topologies were studied. The performance of the each individual design at various optical signal intensity was evaluated. Measured and predicted results were compared in order to evaluate validity of the modelling techniques used. The experimentally observed results closely agree with the theoretical predictions.

In Chapter 4, three examples of the applications for optically operated microswitches are demonstrated. These examples are (1) a switchable microstrip patch antenna and (2) an optically controlled phase shifter, and (3) a reconfigurable microstrip patch
antenna. Both measured and predicted results are presented, and comparisons are given.

Chapter 5 describes techniques of optically controlling EBG structures using micro switches. Two techniques of altering the bandgap are presented. The first technique to be adopted was to enlarge or extend the length of the elements within the EBG structure, since altering the length of the elements shifts the resonant frequency and position of the stopband. The second method employed periodic defects at strategic locations. Defects or perturbations were introduced by means of reducing the length of some of the array elements and therefore altering their resonant frequency. The size of the defects were optically altered in order to shift the stopband.

Finally Chapter 6 concludes the work presented in this thesis and also presents discussion on the possible future work. Considerations are also made to improvements that could be implemented.
1.4 References


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Introduction


2 Introduction to the theory of photoexcitation of semiconductors

2.1 Introduction

The objective of this chapter is to present an overview of the optical mechanism that takes place within a photoconductor, particularly silicon. The photoconductive property of silicon is that free carriers are produced when exposed to an optical signal, and when there is sufficient concentration, it behaves like a conductor \([1]\). These free carriers are made up of electron – hole pairs, and can be referred to as plasma in this thesis. The theory outlined in this chapter is by no means a comprehensive coverage on this subject but serves to highlight some of the key issues regarding photoconductivity in semiconductors that are important for the research here. This unique property of semiconductors has been exploited to effect a transformation of silicon substrate from dielectric state to a pseudo-metallic state because of the induced free carriers. Naturally, silicon exists as an insulator, and when illuminated with light
of appropriate wavelength and flux density it behaves as a conductor. The presence of light simultaneously modifies the silicon's relative permittivity and conductivity, which causes a change of the microwave component response. This unique property of silicon has been utilised to design an optically controlled microswitch. The switch could be implemented in devices where low electrical interference, wide bandwidth operation and size are important features in efficient operation. Examples of the applications of the photoconductivity of silicon were demonstrated in silicon-based photoconductive switches, discussed in the next chapter. The optically controlled switch has the capability to be integrated into reconfigurable microstrip patch antennas and switchable resonator, described in chapter four, including optically tunable electromagnetic band gap (EBG) structures outlined in the fifth chapter. The fundamental techniques on the generation of plasma from high resistivity silicon substrate can be extended to modify the response of several microwave passive circuits such as couplers, and filters. The use of optical switching has some advantages over other methods, including low electrical interference, easy integration, fast switching and robustness. These attributes are discussed in Chapters 3-5.

A mathematical model from Vardaxoglou's extensive work on this subject was used to characterise the optically illuminated silicon in terms of its surface impedance and the plasma density [2]. The surface impedance of the plasma, which is derived from the permittivity, has three distinct phases where the silicon substrate takes on different properties depending on both the free electron - hole pairs and the light intensity. The expression for the permittivity contains distinct terms for the effect of the induced plasma in terms of both electrons and holes [2]. In this chapter, an outline of the optical terms and definitions of the complex dielectric properties of silicon substrate
are given, as well as the derivation of the surface impedance of the plasma by applying the standard transmission line equations.

2.2 Photoelectric effect of semiconductors

This is the phenomenon in which charge particles are released from a semiconductor when it absorbs radiant energy [3, 4]. The photoelectric effect is commonly thought of as the promotion of electrons from the valence to conduction bands of a semiconductor's atoms when illuminated by light of an appropriate wavelength (typically in the range 800-1000 nm) [3]. In the broad sense, however, the phenomenon can take place when the semiconductor is exposed to radiant energy, and when the particles released are excess charge carriers. The photoelectric effect in semiconductors can be understood only in terms of quantum mechanics. It is possible to think of solids as being composed of tiny crystals, each of which has an orderly array of atoms in three dimensions called a lattice structure. Every electron in the lattice, whether bound to an atom or unbound, has a definite energy, called its energy state. Energy states are influenced by the lattice structure of the crystal. Electrons can change to a higher energy state by absorbing heat or electromagnetic energy. According to the laws of quantum mechanics, only specific discrete energy states may exist, and these are grouped together in energy bands, with the regions between them called forbidden bands. The band of highest energy that is occupied when the crystal is at a temperature of absolute zero is the valence band. If this band contains states that are unoccupied, it is called a conduction band and the crystal is a metal. If all the
states are occupied, the crystal is either an insulator or a semiconductor, depending on whether or not electrons from the valence band can be transferred to the next unoccupied band, which then becomes a conduction band. In a semiconductor an input of energy, in the form of light, can raise a valence band electron into the conduction band. Electrons can also be supplied to a semiconductor by atoms having energy levels in the forbidden gap, called impurity atoms. In an insulator the forbidden gap is too large to be bridged easily.

Charge carriers exist for a certain period of time before the electron losses the absorbed energy and return to the valence band. Recombination lifetime is the term given to the average time the electron takes to loose the energy, and this process can either be described as bulk recombination or surface recombination. Bulk recombination can be expressed as three recombination mechanisms, namely, Shockley Read Hall (SRH), Auger, and radioactive, and for a detailed review the reader is referred to the work published by Schrodor [5]. The surface of a untreated silicon wafer consists of a number of dangling bonds and some impurities. These suspended bonds act as strong recombination locations that reduce the carrier lifetime at the surface. This recombination event at the surface is known as surface recombination.

The silicon used in this work was passivated with a silicon dioxide layer to increase the carrier lifetime, and to achieve a high level of photo-induced conductivity. The method of passivation of the silicon wafers used in this research is commercially confidential.
2.2.1 Energy bands of semiconductors

Figure 2.1 depicts a simplified energy band diagram, used to describe semiconductors. The positions of the valence band, conduction band (as indicated by the valence band edge, $E_V$, and the conduction band edge, $E_C$), the vacuum level, $E_{\text{vacuum}}$, and the electron affinity, $\chi$, are indicated in this figure in relationship to each other [2]. The distance between the conduction band edge, $E_C$, and the energy of a free electron outside the crystal (called the vacuum level labelled $E_{\text{vacuum}}$) is quantified by the electron affinity, $\chi$ multiplied with the electronic charge $q$. The diagram identifies the almost-empty conduction band by a horizontal line. This line signifies the bottom edge of the conduction band and is labelled $E_C$. Similarly, a horizontal line labelled $E_V$ indicates the top of the valence band. The energy band gap is located between the two lines, which are separated by the bandgap energy $E_g$. 

![Figure 2.1. Simplified energy band diagram of a semiconductor.](image)
2.2.2 Photoconductivity

Many materials exhibit a marked change in electrical conductivity under optical illumination. Usually the conductivity of many materials increase when light energy is applied to it. This increase in conductivity produced by the absorption of photons is known as photoconductivity [5-7]. In highly conductive substances, such as metals, the change in conductivity is insignificant, while in semiconductors it is significant and has been used for a wide range of applications [8-10]. In a semiconductor there are a certain number of "free" electrons occupying energy levels lying within the conduction band and a certain number of positively charged "holes" left in the valence band by free electrons that migrated into the conduction band as a result of the natural vibrations of atoms. During illumination of the substance, the number of free charge carriers generated (electrons and holes) increases, producing an increased conductivity, this effect will continue until the carriers generated by photon absorption become immobile. However, a high level of conductivity is only sustainable under a constant illumination. One of the most important factors in determining photosensitivity (i.e., the increase in conductivity per photon absorbed) is the free-carrier lifetime, which must be finite because recombination or neutralization takes place between negatively charged electrons and positively charged holes, and there are thus fewer electrons available for conduction. This lifetime can vary from $10^{-2}$ second to $10^{-12}$ second, not only from substance to substance but also in a single substance having varying degrees of imperfections or impurities [11].

Another important factor determining photosensitivity is the light absorption characteristic of the photoconductors. When the photon energy is greater than the
forbidden energy gap of the solid, each photon creates a free charge pair, i.e. a free electron and a free hole. If the photon energy is much larger than the band gap (as, for instance, the photon energy of X rays), the absorption of a single photon can give rise to many free charge pairs.

2.2.3 Illumination intensity

As remarked earlier, illuminating a semiconductor with light of constant intensity generates charge carries, (electrons and holes), which contribute to the steady photoconductivity. Plasma is the term used to describe the electron – hole pairs, which have been excited in this way [12-13]. The relationship between the illumination intensity and optically generated plasma (assuming the number of electron and holes are equal) is given by the following equation:

\[ I = \frac{h \omega LN_S}{(1 - R)\eta \tau} \]  

(2.1)

where 

- \( I \) = illumination intensity, (W/cm\(^2\))
- \( S_p \) = thickness of the plasma layer, (cm)
- \( N \) = charge carrier density, (cm\(^{-3}\))
- \( h \) = Planck’s constant/2\( \pi \), (Js)
- \( R \) = Reflectivity of the light source of the substrate (dimensionless)
- \( \omega \) = angular frequency of the light source, (rad/s)
- \( \eta \) = quantum efficiency, defined as the ratio of the number of photo excited electrons to the number of incident photons (at a given wavelength of light source)
\( \tau = \) carrier life time is the length of time that the carrier is available to contribute to electric conduction, (s)

The reflectivity, \( R \) and quantum efficiency, \( \eta \) depend on the wavelength of light, whilst the carrier lifetime depends on the nature of carrier.

The semiconductor selected in this work is silicon. Silicon was selected because a considerable research has been conducted into it’s properties. Silicon is also one of the widely used material throughout the electronics and photonics industries. Uniform illumination intensity was assumed in examples discussed in this work. Subsequent to this assumption the concentration of the plasma generated was regarded to be uniform throughout the silicon dice.

2.2.4 Complex dielectric constant, \( \varepsilon_r \), of plasma

This section details the derivation of the complex dielectric constant, \( \varepsilon_r \) of plasma from first principles. The electric force exerted on the free charge carriers, namely electrons and holes in silicon is given by the equations:

Force exerted on electron:

\[
eE \cdots = jm_e \omega \nu_e V_e + m_e \nu_e V_e
\]

\[\Rightarrow V_e = \frac{eE}{m_e \nu_e + jm_e \omega} \quad (2.2)\]

Force exerted on holes:
$eE \cdot \cdot \cdot = jm_h \omega V_h + m_v v_h v_h$

$\Rightarrow V_h = \frac{eE}{m_h v_h + j m_h \omega}$  \hspace{1cm} (2.3)

where $e =$ electronic charge of each carrier, (C)

$E =$ incident electric field, (V/m)

$\omega =$ angular frequency, (rad/s)

$V_e, V_h =$ collision frequency of electron and hole, (s$^{-1}$), which is the reciprocal of mean time between collision for the carriers

$m_e, m_h =$ effective mass electrons and holes respectively, (kg)

$V_e, V_h =$ drift velocity of electrons and holes respectively, (m/s)

The total conduction current density, $J_c$ is given by sum of the electrons and holes current densities:

$J_c = N_e e V_e + N_h e V_h$  \hspace{1cm} (2.4)

After substituting (2.2) and (2.3) into (2.4), and assuming that $N_e \approx N_h$ then

$J_c = N e^2 E \left\{ \frac{(v_e - j \omega)}{m_e (v_e^2 + \omega^2)} + \frac{(v_h - j \omega)}{m_v (v_h^2 + \omega^2)} \right\}$  \hspace{1cm} (2.5)

The displacement current density, $J_d$ is:

$J_d = j \omega \varepsilon_0 \varepsilon_a E$  \hspace{1cm} (2.6)

where $\varepsilon_0 =$ permittivity of free space (8.854x10$^{-12}$ F/m)

$\varepsilon_a =$ relative dielectric constant of silicon (11.8)

Total current density, $J_{tot}$ is:

$J_{tot} \cdot \cdot \cdot = J_c + J_d$

$\Rightarrow J_{tot} = j \omega \varepsilon_0 \varepsilon_a E$  \hspace{1cm} (2.7)
By substituting (2.4), and (2.5) into (2.6) and comparing terms, the complex relative permittivity of the plasma layer is expressed by:

$$\varepsilon_r = \varepsilon_0 - \sum \left\{ \frac{\omega^2 p_i}{\nu_i^2 + \omega^2} \left( 1 - j\frac{\nu_i}{\omega} \right) \right\}$$  \hspace{1cm} (2.8)

where $\omega^2 p_i = \frac{N e^2}{m_i \varepsilon_0}$; $\omega p_i$ = plasma frequency, (rad/s)

For the silicon substrate the following values have been used in the model $\varepsilon_0 = 11.8$, $v_e = 4.53 \times 10^{12}$ s$^{-1}$, $v_p = 7.17 \times 10^{12}$ s$^{-1}$, $m_e = 0.259 \: m_o$, $m_p = 0.38 \: m_o$, where $m_o$ is the free electronic mass.

### 2.2.5 Complex refractive index, $n_r$ of plasma

Accurate measurements of the complex refractive index can be obtained by using a technique called spectroscopic ellipsometry [14]. Spectroscopic ellipsometry is a powerful optical measurement technique which enables the thickness and complex refractive index ($n_r$ plus the extinction coefficient $k$) of a wide variety of materials in multiple layer film stacks to be calculated. Due to the contribution of plasma, the imaginary part of $\varepsilon_r$ increases significantly and becomes comparable to the real part. This change marks a turning point in the electromagnetic response of the material.

The corresponding refractive index, $n_r$ is also complex.

$$n_r = \sqrt{\varepsilon_r}$$

$$\Rightarrow n_r = n - jk$$  \hspace{1cm} (2.9)

where $n$ = refractive index, is the ratio of phase velocity of light in vacuum to that in the medium.
\[ k = \text{extinction index, is related to exponential decay in the amplitude of light wave as it passes through the material.} \]

Figure 2.2. Optical constants of silicon as a function of plasma concentration.

Figure 2.2 is a representation of the change of complex refractive index with plasma density according to equations (2.7) and (2.8). The silicon had the following parameters:

- The relative dielectric constant of Si in its "off" state, \( \varepsilon_0 = 11.8 \),
- collision frequency of electron, \( v_e = 4.53 \times 10^{12} \text{s}^{-1} \),
- collision frequency of hole, \( v_h = 7.71 \times 10^{12} \text{s}^{-1} \),
• effective mass of electron, \( m_e = 0.259m_0 \),

• effective mass of hole, \( m_h = 0.38m_0 \).

The free electronic mass, \( m_o \) and electronic charge, \( e \) are \( 9.109 \times 10^{-31} \) kg and \( 1.602 \times 10^{-19} \) C respectively. The graph indicates that at low plasma concentration, (up to \( N \approx 1 \times 10^{14} \) cm\(^{-3}\)) the real part of the refractive index, (shown by the solid curve) initially remains almost constant for a given plasma concentration. However, as can also be seen from Figure 2.2 real part then increases in such a way that (when \( N \approx 1 \times 10^{16} \) cm\(^{-3}\)) it coalesces with the imaginary part reaching a value of about 10. Subsequently, both constants increase with same rate and have similar values for any given value of plasma density. This is region where the conduction process takes place, and for regions exhibiting refractive indexes above 100, the plasma density has reached its near metallic state. It is within the scope of this research study to utilise this unique property of silicon substrate to modify the performance of a microwave device incorporating silicon switches.

### 2.2.6 Complex propagation constant, \( \gamma \) of plasma

The complex propagation constant can be written as

\[
\gamma = jk_0 n,
\]

(2.10)

where \( k_0 = \) free space wave number

\[
= \frac{2\pi}{\lambda}
\]
2.2.7 Complex surface impedance, $Z_{sp}$ of plasma

The complex surface impedance of the plasma layer at different plasma concentrations and different layer thickness were computed by applying the standard transmission line equations for lossy media. [15-17]. The surface impedance was calculated assuming that there is no diffusion of charge carriers into the substrate.

To obtain accurately the scattering parameters, the plasma thickness ($t$), as in the case of Frequency Selective Surfaces (FSS), must be small, (i.e. $t \ll \lambda$) [16, 17]. Films of infinitesimal thickness and specified sheet conductance can then be used to approximate the materials in the FSS. The sheet conductance will be derived by first considering Figure 2.3, which shows a continuous, film that is not FSS. The film is characterised by a complex index reflection ($\hat{N} = N - jK$), from which one can obtain the complex conductivity ($\hat{\sigma} = \sigma_1 + j\sigma_2 = j\omega\varepsilon_0\hat{N}^2$); the intrinsic admittance ($\hat{Y} = (1/377)\hat{N}$), and the propagation constant ($\hat{\gamma} = j(2\pi/\lambda)\hat{N}$).

At the input and output surfaces, the complex electric and magnetic fields are denoted $E_{IN}, H_{IN}$ and $E_R, H_R$ respectively. $E_R$ and $H_R$ are formed from the sum of the incident and reflected fields. The admittance of free space region to the right of the film is $Y_R = H_R/E_R = 1/377$ $\Omega^{-1}$. The film admittance incorporating the effect of the right hand half space is $\hat{Y}_{IN} = H_{IN}/E_{IN}$. The desired sheet conductance is $\hat{\sigma}_s = H_F/E_{IN}$ $\Omega$($\text{square}$)$^{-1}$, where $H_F$ is the contribution to the magnetic field at the input due to the conduction and displacement current in the film alone:

$$\frac{\hat{\sigma}_s}{E_{IN}} = \frac{H_F}{E_{IN}} = \frac{H_{IN} - H_R}{E_{IN}} = \hat{Y} - \frac{E_R}{E_{IN}}Y_R$$

(2.11)
The quantity can be obtained by solving the Maxwell-Ampere integral equation for \( H_F \) or simply by applying standard transmission line equations:

\[
Y_{in} = \left[ \frac{\hat{Y}}{\hat{N}} \right] \frac{\cosh \gamma t + \hat{N} \sinh \lambda t}{\cosh \lambda t + \sinh \lambda t}
\]  

(2.12)

\[
\frac{E_R}{E_{IN}} = \frac{\hat{N}}{\hat{N} \cosh \lambda t + \sinh \lambda t}
\]

(2.13)

\[
\hat{\sigma}_s = \left[ \frac{\hat{Y}}{\hat{N}} \right] \frac{\cosh \gamma t + \hat{N} \sinh \lambda t - 1}{\cosh \lambda t + \sinh \lambda t}
\]

(2.14)

Because of the open areas in the thin FSS described above, the electric field is impressed almost equally on both sides of the film. Applying the superposition theorem principle as indicated in Figure 2.3, the current in the film is doubled and the sheet conductance, is given by:

\[
\hat{\sigma}_s = \frac{2H_F}{E_{IN} + E_r} = \frac{2\hat{\sigma}_s}{1 + (E_r/E_{IN})}
\]

(2.15)

\[
\Rightarrow \hat{\sigma}_s = \left[ \frac{\hat{Y}}{\hat{N}} \right] \frac{\cosh \gamma t + \hat{N} \sinh \gamma t - 1}{\cosh \gamma t + \sinh \gamma t + \hat{N}}
\]

If the film thickness is much smaller or larger than the skin depth, equation 2.15 reduces to:

\[
\hat{\sigma}_{ss} \approx \hat{\sigma} \quad (t \ll \delta)
\]

and

\[
\hat{\sigma}_{ss} \approx 2\hat{Y} \quad (t \gg \delta)
\]
2.3 Electromagnetic Simulations

2.3.1 Modeling techniques

Two simulation softwares employing, a 2D Method of Moments (MoM), and a 3D Transmission Line Matrix (TLM) method were employed to perform simulations. The 2D simulation software is a commercially available microwave computer-aided-design (CAD) package called Microwave Office [18]. This package employs EMSight as it’s electromagnetic simulator. EMSight is a Windows based electromagnetic and linear circuit simulator for analysing the behaviour of circuits.
and structures at high frequencies. EMSight is capable of analysing arbitrary circuits incorporating interconnecting vias on an unlimited number of dielectric layers with unlimited number of materials, geometries, and ports. However this package has a drawback of long run times.

The 3D simulation software used is called MicroStripes [19]. MicroStripes employs the Transmission Line Matrix (TLM) method to solve Maxwell’s equations in the time domain. This package is ideal for solving extremely complex and electrically large designs. The TLM solver tolerates large aspect ratios for grid cells and rapid changes in grid density. This enables localized gridding and keeps mesh requirements to an absolute minimum.

2.3.2 Characterisation of plasma

Within the simulation environment the electrical or physical properties of conductors and of the plasma both in its static and photo-illuminated state are specified. The electrical parameters are defined in terms of the material’s low frequency conductivity, high frequency loss coefficient and excess surface reactance. The physical parameters consist of the conductor thickness and material conductivity.

The low frequency parameter may be used to specify the DC resistance of the planar conductor (ohms/square). The DC resistance relates to the resistance of the conductor assuming a uniform current distribution throughout the cross-section of the conductor. The DC resistance is typically set to the following equation where, \( \sigma \) is the conductivity of the conductor and \( t \) is its thickness.
The high frequency loss coefficient specifies the loss associated with the conductor at frequencies where the thickness of the conductor is significantly larger than the skin depth. Since the loss associated with the skin depth effects are proportional to the square root of frequency, the skin depth loss coefficient is multiplied by the square root of frequency to provide a value in units of ohms/square value that is used for loss computations.

The high frequency loss coefficient is multiplied by a factor of \((1+j)\) to ensure that the real and imaginary components of the surface impedance are equal as associated with the skin depth phenomena. The high frequency loss coefficient is given by:

\[
R_{\text{HF}} = \sqrt{\frac{\pi \mu}{\sigma}}
\]  

At low frequencies, the software in the computation of conductor loss uses the DC resistance, while at high frequencies the high frequency loss coefficient is employed to compute conductor loss. In the intermediate or transition region (frequencies where the skin depth is close to the thickness of the conductor) both factors are used.

A simpler alternative method for specifying the electrical parameters of a material is to enter them as physical parameters and use EMSight to convert them to the correct electrical parameters [18]. To specify the physical parameters, the thickness of the conductor and the conductivity of the conductor must be specified.
The loss value computed using the physical parameters assumes an ideal material. Impurities and surface roughness tend to make the actual measured loss value somewhat higher than the value obtained through simulation. One technique for improving the agreement is to decrease the conductivity (typically 10% - 20%) of the conductor so its simulated loss agrees more closely with the measured value. Measured loss data obtained from a transmission line can be matched to the simulation data value of the same line by iteratively adjusting the "effective" conductivity of the conductor. This "effective" conductivity can then be employed in place of the ideal conductivity.

2.3.3 Modelling surface impedance of plasma

The surface impedance is estimated using the known value for the relative permittivity of the semiconductor, which contains electron-hole pairs, and the relationship between the impedance variation and plasma concentrations and thickness has been extensively studied by Vardaxoglou and Lockyer [1, 2]. In this work, the model is improved to accommodate the properties of silicon under optical illumination perpendicular to the direction of travel of the microwave signal. This relationship is plotted in Figure 2.5. The graph depicts three different trends. Firstly, at low plasma concentration there is high value of negative reactance, and increases in proportion to the plasma thickness, whereas the resistance remains small and almost constant. The second trend occurs for plasma density values between $10^{13}$ cm$^{-3}$ and $10^{16}$ cm$^{-3}$, the reactance increases gradually towards zero and the resistance exhibits its maximum value at a concentration of about $10^{15}$ cm$^{-3}$. Within this region the plasma behaves like a lossy medium. Finally, both the resistance and reactance decay
exponentially to zero as the concentration increases. Where the impedance drops to zero the plasma enters its fully conducting state.

In order to simulate any structure integrated with silicon switches, the derived values for surface impedance or equivalent conductivity ($\sigma_e$) plotted in Figure 2.5 are used to define the properties of the silicon dice. The graphs in this figure were plotted using MathCAD. The simulated scattering parameter results are compared with measured results obtained by varying the optical power incident on the small piece of silicon. By matching simulated and measured S-Parameter results, each optical power used is assigned equivalent conductivity or impedance values that can be used in a simulation environment.

Figure 2.5. The impedance pattern for different plasma concentrations and thickness
2.4 Conclusion

The basic theory covered in this thesis is based on the optical control of microwave circuits using high resistivity silicon. The theory serves to highlight some of the key issues regarding photoconductivity. Photoconducting processes that occur within silicon under optical illumination has been analysed, and illustrated with graphs. The objective of this research work was on implementing the photoconducting silicon in various microwave circuits. Silicon, a photoconductor, can have its conductivity varied simply by exposing it to light of appropriate wavelength and flux density. It has been pointed out that in order to achieve high plasma concentration, long carrier lifetimes are required. The two main recombination events discussed are the bulk and surface recombination, and these mechanisms reduce the carrier lifetime of silicon. Subsequent to this, it is of paramount importance to uniformly apply silicon oxide layer on the silicon wafers to minimise recombination mechanisms.

The characterisation of the silicon and the photo-induced plasma has been described. It has also been shown that values for surface impedance and equivalent conductivity ($\sigma_e$) are used to define the properties of silicon in the simulation environment. The surface impedance and equivalent conductivity parameters are used in the simulation package to simulate the optical light. The relationship between the surface impedance and concentration of the plasma exhibits three distinct phases, namely, the semiconductor state, lossy and conducting state, and each region gives different characteristics.
2.5 References


3 OPTICALLY CONTROLLED SWITCHES

3.1 Introduction

Microwave switches are widely used components in a variety of communication devices, for example, phase shifters, and antennas. The increasing number of publications covering the topic of optically controlled microwave switches continuously confirms the interest in this area [1-34]. These types of switches have been proposed and successfully implemented for various applications, including high-speed microwave switching and controlling [1-4], high-speed sampling [5-6] and the generation of high-frequency waveforms and ultrafast kilovolt pulses [7-10]. In these publications investigations on how the performance of microwave structures can be altered by photoinduced carriers created within semiconductors were studied.
Without considering the different substrate materials such as Si, GaAs, or InP, microstrip based switches can be classified by the method of illumination. The first group includes switches with a single microstrip gap discontinuity structure, in which both the light on and light off processes are achieved at the same location in the semiconductor substrate [1, 4-5, 7-9]. In this case, two optical pulses of greatly differing wavelengths are required to produce regions of highly conductive surface plasma and bulk plasma, respectively. Using materials such as Cr:GaAs with fast-recombination properties are used, the device can turn off automatically due to short lifetime of the excess carriers [10].

The second group includes microwave switches investigated by Platte, that consists of a combined gap shunt microstrip structure, [2-3, 6], in which the light on and light off processes occur at different locations in the supporting semiconductor substrate.

So far majority of microstrip based photoconductive switches have been realised by placing a narrow gap or slot incorporated within a semiconductor substrate and laser pulses are employed to switch on and off the switch. This technique has a major drawback of generating problems associated with diffusion of the plasma within the semiconductor substrate. Moreover, great care has to be taken in fabricating and mounting these switches because the semiconductor materials used are brittle. In contrast to the two groups mentioned above, this thesis presents an alternative method of realising a simple photoconductive micro-switch diced from silicon wafer of high resistivity ($\rho$), (typically $\rho > 6000 \ \Omega \ cm$). The light on and light off processes occur across the silicon dice. When illuminated by light of an appropriate wavelength, silicon has a unique property of changing from an insulator state to a pseudo metallic state by forming electron-hole pairs. This proposed design can be easily fabricated.
and mounted on any arbitrary substrate material using a relatively small silicon dice incorporated within a microstrip gap discontinuity acting as the active region. The capacitance across the microstrip gap can be controlled by illuminating the silicon dice. This approach can be considered novel for a number of reasons such as: (1) the area of excitation is smaller than the conventional photoconductive microwave switches, which offers the advantage of employing low power optical illumination from sources such as LEDs, (2) the switching processes takes place at the same location employing a light pulse of a particular wavelength to activate the switch, and (3) size matters, smaller is better for easy integration.

The behaviour and performance of the micro-switch have been studied using measurements and Electro-Magnetic (EM) simulations. An equivalent circuit represented by resistors, inductors and capacitors has been found to accurately model the micro-switch. The resistor and capacitor across the gap represent the photoconductance. The circuit has enabled the characterisation of the silicon dice incorporated in a transmission line under varying optical powers. The objective of designing the equivalent circuit is to relate the circuit components to the optical power and plasma concentration. In addition to EM linear simulations, this circuit can be a very useful tool in determining and ensuring optimum switch performance.

This chapter presents an in depth study of a microwave micro-switch activated by optical illumination. Light signals do not interfere with RF radiation and therefore immunity is preserved for high performance systems. Three major goals for an acceptable switch for commercial applications are as follows:
1. low magnitude of the electric fields across the gap when the switch is off (static case) that create high electrical isolation,

2. losses within active elements below a specific level when the switch is on (active case), and

3. switching time which meets the demands of microwave and millimeter frequency applications.

Grant et al stated that most applications in telecommunications and radar require switching times of microseconds to milliseconds [11]. In recent literature [12], a microwave switch based on micro electromechanical systems technology exhibited turn-on times on the order of 200 $\mu$s, and turn-off times below 20 $\mu$s. This switch was cycled over 200 000 times before failing. In most cases, failure was caused by metal stiction. However, the switching time of an optical microwave switch depends on the optimum carrier lifetime of the semiconductor. Studies by Platte revealed that the optimum carrier lifetime for silicon substrate was in the range $10^{-6}$s to $10^{-5}$s [13]. There are no mechanical parts to wear out in an optically operated switch. While it is generally accepted that mechanical switches offer the benefits of low insertion loss and high isolation, semiconductor based switches on the other hand offer attractive features such as small size, weight and low fabrication cost [11].

3.2 Fundamental advantages of switches using optical control

Microwave switches constitute as building block components and, as such, they are employed in a diverse range of applications, including cell phones, military communication hardware, radar systems and automated test equipment. The most
common types of microwave switch currently employed within the microwave industry are Micro ElectroMechanical Switches (MEMS), and semiconductor switches based on FET and p-i-n diode [14, 15].

### 3.2.1.1 Semiconductor switches

#### MESFET

In the conducting state, the MESFET is operated with zero bias and the channel is approximately a linear conductor. In the insulating state, the gate to channel Schottky diode is reverse biased depleting the channel. The off state capacitance is determined by the capacitance values of the MESFET. The performance of MESFETs are limited by the conductivity of the AlGaAs channel. The channel is heavily doped, near the solubility limit, but it must be made thin enough that the Schottky diode can pinch it off before breakdown. There is a further limitation on the performance because of the planar structure of the electrodes which add fringing fields and creating relatively large capacitance. Some manufacturers, such as Blackwell’s have achieved improved performance by selectively thinning the substrate in order to reduce the fringing fields.

#### P-I-N

A P-I-N diode can be regarded simply as a parallel plate device in the off state, and its performance depends on the volume of the conducting region. The on-state resistance of this device depends on the optical power, the carrier density and lifetime across the conducting region.

#### Photoconductive

Photoconductive switches have been used for very high-speed optically controlled switching of microwave signals and there has been an interest in making efficient optically controlled switches. A photoconductive switch has the geometry of an
discontinuity. Under illumination photo-carriers generated in the semiconductor regions provide an on-state resistance which closes the switch. The structure of this switch is very similar to that of a MESFET in that the electrodes are planar. However, the isolation is somewhat better because the gap is usually made longer than the short channel length of a MESFET.

### 3.2.1.3 MEMS Switches

There are both resistive MEMS switches that make metal to metal contact and variable capacitor MEMS switches that make metal to insulator contact. However, this thesis will consider the metal to metal contact type. The microwave structure is based on a gap in a coplanar waveguide. A top contact suspended from an insulating beam can be lowered to make electrical contact across the gap thus closing the switch. Two large capacitors act as the actuator to bend the bridge and lower the contact. Subsequently, there exists some degree of freedom available to the switch designer that does not exist in a semiconductor switch.

### 3.2.1.3 Comparison

The comparison of various switches is summarised in Figure 3 adopted from [11]. Grant et al revealed that the principal advantages of MEMs are their low insertion loss (typically around 0.1 dB), high off-state isolation provided over a wide frequency range, and high power handling capabilities [11]. The low loss associated with the electromechanical switch results because the switch arms are manufactured from a highly conductive material. A high degree of electrical isolation is achieved due to the absence of any electrical connection between the open contacts. However, these devices are bulky. On the other hand FET and p-i-n diodes provide much faster
switching speeds and are smaller in both size and weight. Freeman et al. studied the transient response of a PV/FET switch, and the fall time is less than 10 ns [16]. The major drawbacks associated with FETs and p-i-n diodes are their high level of insertion loss, high DC power consumption, low off-state isolation and low power handling capabilities.

![Graph showing the comparison of the FOM for a number of RF switches. The labelled curves are for 1 - measured S21 NRC opto, 2 - Opto 40 Ω, 80 ff, 3 - PIN 1 Ω, 110 ff, 4 - Opto 100 Ω, 30 ff, 5 - FET 5 Ω, 100 ff, 6 - Rockwell MEMS switch, and 7 - 60 μm coplanar waveguide gap on quartz measured at NRC. Gray boxes are the switching ratio for mechanical coaxial switches.](image)

Microwave switches employing photoconducting materials offer superior electrical performance and attractive properties not afforded by MEMs or conventional semiconductor switches based on p-i-n diodes or FETs. Figure 3.1 illustrates the basic geometry of a photoconducting switch. The switch consists of a microstrip gap discontinuity bridged by a silicon dice. Under optical illumination charge carriers are generated within the semiconductor. When illuminated the silicon forms a conducting path between the metal contacts. Extinguishing the light source has the effect of...
opening the switch. Auston demonstrated that illumination by a pulsed laser could produce pseudo-metallic photoconduction in silicon [17]. The following subsections briefly describe the properties of photoconducting switches, which make them suitable for many consumer/industrial applications:

### 3.2.1 Cost effective and scalable technology

Photoconducting switches are relatively simple devices to fabricate and consequently the associated production costs would generally be low. Not only is this technology cost-effective, but its scalability addresses the problem of wasted frequency bandwidth within the area of electromagnetic spectrum used for communications. There has been an ever increasing level of demand for this type of switch principally driven by the demand for low-cost novel microwave structures for use within portable wireless microwave frequency communications systems [18, 19].

![Diagram of a photoconductive switch](image)

**Figure 3.1** Basic structure of a photoconductive switch. (Design parameters of the switch are given and discussed in section 3.3 of this Chapter.)
3.2.2 Reliable and repeatable operation

Optically activated switches for use in microwave systems are more reliable than conventional microwave switches because they are solid state devices. A problem often associated with MEMs is the possibility of indeterminable switch states, which is unacceptable in safety critical systems, for example. MEMs incorporate a range of mechanical moving parts such as wheels and springs. Photo-illuminated switches have a virtually unlimited operating life and would require very little routine maintenance. When a photoconducting switch is uniformly illuminated with light, the density of charge carriers is approximately uniform throughout the switch and thus the current is also distributed uniformly throughout the conducting regions of the switch. In comparison with conventional electronic switching devices, the photoconducting switching technique for the control of microwave systems is both more reliable and repeatable [20].

3.2.3 Size and ease of systems integration

The photoconducting switch is a good candidate for incorporation into miniaturised and well-developed large-scale integrated circuits. As has been the case for years, miniaturisation continues to be a key trend, though suppliers acknowledge that some products can not shrink much more because of the user interface requirements for example, the size key pad of a mobile phone can be limited to the average size of users fingers. In addition microstrip is one of the most commonly used microwave planar transmission line, and it can be integrated with other passive and active microwave devices. Optical control improves the ease integration and reliability, whilst also reducing manufacturing costs [21, 22].
3.2.4 Speed and efficiency

Photoconducting switches have been used at very high speeds switching microwave signals [23, 24]. These switches are required in telecommunications applications to achieve high data transmission rate. Many researchers in the communication field including Rebeiz, believe that the practical limit of switching time will be around $1 \, \mu s$ for high reliability operation [25].

3.2.5 Off state isolation and insertion loss

The level of electrical isolation provided by photoconducting switches is greater than that of other semiconductor switches because the gap is usually made longer by employing various techniques which will be considered in detail in subsequent sections [26]. Low insertion loss (typically $< 0.5$ dB) is achievable, and the loss depends on the optical power, carrier mobility and lifetime. The relationship between these preceding factors and plasma density has been discussed in Chapter 2.

3.3 Design parameters of semiconductor switches

![Diagram of a photoconductance switch geometry with silicon dice placed between metal contacts.](Figure 3. 2(a). A typical photoconductance switch geometry with silicon dice placed between metal contacts)
Figure 3.2(b). (Not drawn to scale) The configuration and dimensions of a switch with silicon dice above the 0.25 mm microstrip gap discontinuity.

Figure 3.2 shows the general geometry of a photoconducting switch discussed in this work. The switch parameters were chosen such that the characteristic impedance of the transmission line is 50 Ω. The microstrip line width $W$, for given characteristic impedance $Z_0$ on a substrate of known thickness $t$, can be found from [27].

3.4 Illumination and optical sources

Silicon dice having a thickness of 300 μm, and a dielectric constant, $\varepsilon_r = 11.8$ incorporated within the gap was illuminated with a range of optical powers between 0 and 200 mW. Intrinsic silicon has a conductivity of $0.439 \times 10^{-3}$ S/m [31]. Silicon used in this work is n-type doped with Phosphorus to increase its static conductivity to $16.7 \times 10^{-3}$ S/m. In the absence of optical excitation, the high resistivity ($\rho$), silicon
(typically $\rho > 6000 \ \Omega \ cm$) behaves as a low conductivity dielectric slab, ($\sigma_s = 0.0167 \ S/m$). Uniform illumination of the silicon changes its conductivity as high as 250 S/m and leads to the generation of a photoinduced plasma within the silicon.

Two types of optical sources, namely laser diode and LEDs were used in the measurements. A laser diode operating at 980nm delivers light through a 1.0 mm diameter glass fibre optic cable onto the silicon switch. The glass fibre reduces EM interference and provides an electrical and thermal isolation for the switch control circuitry. A large laser spot size covered the silicon dice when the fibre was focused at an angle of approximately 45° from the plane of the dielectric substrate. In contrast to the laser light delivery system, the LED is placed directly above the silicon dice. The type of the LEDs employed for this purpose was a semiconductor based (Gallium Aluminium Arsenide - GaAlAs) infrared emitting diode in a TO-39 package (emitting at wavelength $\lambda = 880 \ nm$). However the optical power delivered from the LED was limited to about 30 mW in the Continuous Wave (CW) mode. These LEDs give more optical power when pulsed compared to continuous operation. A simple design for a LED pulsing circuit was built in order to drive the LEDs. The ability to employ low power optical sources such as LEDs whilst maintaining a good standard of device performance was the primary goal. The low power consumption would be an advantage over the conventional products. A pulsing circuit was made, and this circuit is capable of driving the LEDs with a current of about 1A in pulsed operation, compared to 0.5A for continuous operation. With an aid of this pulsing circuit the speed at which the microwave switch operates was determined. The pulsed measurements were carried out using a Radio Frequency – Direct Current (RF-DC) detector calibrated with respect to the laser diode performance. The detector produced
a DC voltage which is proportional to the RF power at its input terminals. The RF source was the HP 83650L Sweeper, and the DC voltage data was measured using an oscilloscope. A block diagram of the equipment used is shown in Figure 3.3. When pulse modulation from the oscillator is in effect, the RF power from the Sweeper is turned on and off at a rate determined by the frequency of the pulsating input. A snapshot of a pulsed response at 2.75 GHz is shown in Figure 3.4, and a measure of the switching speed is clearly demonstrated from both the rising edge and the falling edge of the pulse. Evaluation of the rise and fall time in Figure 4 with respect to the pulsed light indicates that the rise time is 10.5 μs and the fall time is 140 μs. The higher fall time is due to the high carrier lifetime of silicon. Reducing the carrier lifetime will result in faster switching times, but will require higher optical powers for switching. The tunable devices to be considered in this project will be based on this type of switch.

The intensity of the illumination from both the fibre end and the LED glass dome was measured using a calibrated optical power meter. Microwave measurements were performed, using an HP5337D network analyser, at various optical powers, namely 0, 0.5, 10, 30, and 200 mW to evaluate the performance of the switch.

Figure 3.3. A block diagram of the measurement set-up incorporating pulsed LED.
1. Fabrication techniques of semiconductor switches

A cross sectional view and the configuration of the structure of the photoconducting switch realised from a microstrip gap discontinuity are illustrated in Figure 3.2. The simple construction of this switch makes it easy to fabricate. A small 300 μm thick Si dice, Ls X Ws is bonded in the manner that it overlaps the microstrip gap by a distance, dx as illustrated in the figure. This is a novel technique for realising ‘localised’ photoconducting switches, contrary to the conventional arrangement where the host substrate is a semiconductor that result in a large absorption region [28, 29].

The use of a small Si dice and the illumination of only small and localised area of the surface of silicon to shorten diffusion areas and reduce power dissipation. Three possible ways in which this dice can be placed across the gap were investigated. Experiments were conducted with silicon located in each of the following positions:
(a) **Silicon above the gap (SAG).** The two ends of the silicon substrate were in contact with areas of microstrip transmission line on either side of the gap, over a distance of $G$ mm. A lump of silver epoxy was used to hold the silicon in place and to improve the ohmic contact at the boundary, as shown in Figure 3.5. Silicon oxide used for passivating the switch unfortunately degrades the ohmic contact between the silicon and the copper tracks, thus reducing DC conductivity. However, at frequencies above 1 GHz the effect of this is not observed.

(b) **Gap above silicon (GAS)**, within a cavity in the dielectric substrate to a depth of 300 μm and having an area equal to $W_S \times L_S$ square units.

(c) **Gap above silicon and with the dielectric substrate undisturbed (GBS).** The two ends of the copper microstrip lines were placed above the silicon dice and separated by a gap of $G$ mm.
Female SMA connectors with center tab contacts were connected to each end of the copper microstrip line with a solder joint. The width of the strip was tapered so that it was no wider at the edge of the substrate than the width of the pin from the SMA connector, in order to reduce the capacitance of the centre conductor and the strip line. Generally, excess capacitance due to the fringing fields exists within this transition region [30, 31]. The material used for the substrate was RT-Duroid 5880, 1.125 mm thick, and having ½ ounce copper cladding on both sides. A comparison of the performance of switches with the gaps positioned on either above or below the silicon dice are shown in Figure 3.6. As expected, the silicon properties were altered by optical illumination. These results confirm that a simple microwave Si switch can be realised employing optical injection technique. Optical illumination was directed above a small Silicon dice place above a microstrip discontinuity. Under optical illumination photoconduction takes place in silicon. Consequently, the silicon forms a conducting path between the copper strips. Extinguishing the light source has the effect of opening the switch. Auston demonstrated that illumination by a pulsed laser

![Figure 3.6. Transmission responses of Si switches incorporating gaps in different configurations](image-url)

Frequency (GHz)

-20  -18  -16  -14  -12  -10  -8  -6  -4
Transmission Coefficient (dB)

Light On (200mW)

Light Off

SAG

GAS

GBS

1.00  1.25  1.50  1.75  2.00  2.25  2.50  2.75  3.00
Frequency (GHz)
could produce pseudo-metallic photoconduction in silicon [8]. It was found that the switch constructed employing the SAG configuration gave the largest On-to-Off ratios, and all the switches discussed in subsequent sections will be of this nature. Although noticeable switching exists, the insertion loss is high, and the isolation is low, thus limiting the application of such a switch. To improve the parameters of the switch, the topology of the gap can be optimised.

3.6 Effective microwave switches

Novel techniques to enhance both the optical coupling efficiency and the dark state isolation of the planar photoconductance switches were devised. Three topologies were investigated for microstrip gap discontinuity with:

1. notches, S1
2. interdigitated and interleaved pattern, S2 and
3. chamfered and tapered pattern, S3 as portrayed in pictures in Figure 3.7.

In all the above cases, a highly conductive silver loaded epoxy was used to hold the silicon dice in place. This epoxy also helped to establish a good ohmic contact between the semiconductor and the transmission line. The resistivity from the manufacture's data of this epoxy is less than $1 \times 10^{-3}$ ohm – cm [33].

3.7 Performance of optical switches

Three different types of switches were manufactured using etching techniques. The transmission lines were fabricated on a 1.125 mm high RT Duroid substrate having a dielectric constant of about 2.2. The objective in each case was to optimised the optical switch performance.
Figure 3.7. Photographs of three switch topologies before and after construction. (a) notches, S1 (b) interdigitated and interleaved pattern, S2 and (c) chamfered and tapered pattern.
3.7.1 S1 topology

The geometry of this design was such that an area of 2 mm$^2$ was notched on both sides of the 0.25 mm gap. The purpose of incorporating these notches was to enhance the electrical isolation of the switch. These notches reduced the capacitance across the gap. A 2 mm x 1.5 mm silicon dice was placed across the gap, and firmly fixed to the line using silver loaded epoxy. This small portion of the switch acted as the active region. One may recall that diffusion processes reduce the concentration of the plasma generated within the silicon. The measured scattering parameters exhibited by the switch for the excitation frequencies in the range 1 to 3 GHz are illustrated in Figure 3.8. In the off state for the 1 GHz excitation, the values of insertion loss and return loss are 22.4 dB and 0.37 dB respectively. The electrical isolation at this frequency is about 20 dB, and this was mainly enhanced by the notches across the microstrip gap. The insertion loss and return loss were switched to 2.6 dB and 9.9 dB respectively when the device is illuminated with optical power of 70 mW. It was found out that these losses were reduced by less than 4% when optical power was increased to 200 mW. This suggests that for any given optical power higher than 70mW, the photons incident upon the silicon promote few additional electrons from the valence to the conduction band. It can be noted that the majority of the carriers are generated with just 30mW of optical power.

3.7.2 S2 topology

Unlike a simple straight gap in Case S1, the topology for switch S2 consists of an “interleaved” gap. The size of silicon substrate placed across the gap is 2 mm x 3.5 mm. From Figure 3.9 the insertion loss parameters measured with zero and 200 mW of optical illumination changed from 19.65 dB to 1.52 dB at 1 GHz. This switch gave
a better insertion loss in the on state than switch topology S1. The method of employing an interleaved discontinuity reduced the losses in the active state by about 40% compared to a conventional gap. However the degree of electrical isolation for topology S2 was about 12% lower than that exhibited by S1 due to high coupling that exists across the gap.

3.7.3 S3 topology

The nature of the gap of the third switch topology is identical to that of S1 discussed in section 3.8.1, except that the corners of the gap are chamfered to an angle, $\theta = 51.5^\circ$ to reduce the effective gap capacitances. The geometry of the photoconducting switch is shown in Figure 3.10 (a). A silicon dice measuring 2 mm x 1 mm x 0.3 mm, $(L \times W \times H)$ bridged the microstrip gap discontinuity, $G$. The dice is fixed across the gap using a conducting silver epoxy placed on both ends of the dice. The micro structural details at these junctions are illustrated by a picture in Figure 3.10 (b). The material used for the substrate is Taconic substrate, TLY-5, 1.15 mm thick, and having dielectric constant of 2.18. The size of the substrate used is kept to about 27 mm x 15 mm and thus the photoconducting switch is a good candidate for incorporation into miniaturised devices.

The measurement results obtained from a switch incorporating this type of gap are shown in Figure 3.11. Measurements for the switched line under 0, 0.5, 1.2, 10, 30, 70, 100, and 200mW were carried out. The transmission and reflection coefficients measured at 1 GHz in the dark state are $-23.4 \text{ dB}$ and $-0.34 \text{ dB}$ respectively. At 1 GHz this switch gave insertion and return losses of 0.7 dB and 21.6 dB respectively when illuminated with 200 mW of optical power. These are the best results obtained from any of the switches used. The measurements show that the responses of the
switch strongly depend on the level of optical power. Measurement results indicate that one may achieve an insertion loss value as low as 1dB with only 70mW of optical power.

A summary of the comparison of the performances of the switches incorporating the three topologies is given in Table 3.1. It can be seen that the switch with S3 topology gives the best performance. The simple construction of this switch makes it easy to fabricate, and can be regarded as an alternative to the conventional optically controlled microstrip switches supported by a semiconductor substrate [1, 8]. Another important feature of the switch is the ability to control the phase up to 70° by varying the optical power incident on the silicon between 0mW and 200mW. This phenomenon is presented in Chapter 4.
Figure 3.8. The measured results for switch topology S1.
Figure 3.9. The performance of switch topology S2.
Figure 3.10(a) Illustration of the geometry of the analysed photoconductive microswitch (Not drawn to scale), (b) A picture of the active region.

<table>
<thead>
<tr>
<th>Switch topology at 1 GHz</th>
<th>S1</th>
<th>S2</th>
<th>S3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transmission Coefficient (dB)</td>
<td>Off</td>
<td>-22.4</td>
<td>-19.6</td>
</tr>
<tr>
<td></td>
<td>On (200 mW)</td>
<td>-2.7</td>
<td>-1.5</td>
</tr>
<tr>
<td>Reflection Coefficient (dB)</td>
<td>Off</td>
<td>-0.4</td>
<td>-0.5</td>
</tr>
<tr>
<td></td>
<td>On (200 mW)</td>
<td>-8.5</td>
<td>-13.8</td>
</tr>
<tr>
<td>Phase (Deg)</td>
<td>Off</td>
<td>20.7</td>
<td>11.9</td>
</tr>
<tr>
<td></td>
<td>On (200 mW)</td>
<td>-34.6</td>
<td>-51.4</td>
</tr>
<tr>
<td>Phase shift (Deg)</td>
<td></td>
<td>55.3</td>
<td>63.3</td>
</tr>
</tbody>
</table>

Table 3.1. Summary of the performances of switches incorporating S1, S2 and S3 topologies for both on and off states at 1 GHz.
Figure 3.11. Measured scattering values of topology S3.
3.8 Simulations

The optically activated switch was evaluated using a commercial electromagnetic simulation package called Microwave Office™. The EM solver in this package is based on MoM. The Electro-Magnetic (EM) analysis and linear simulations were used to refine the gap topology derived from a conventional single gap discontinuity in order to obtain high electrical isolation and low insertion loss for the OFF and ON states of the switch. In order to simulate the optical switch, the derived values for equivalent conductivity ($\sigma_e$) are used to define the properties of the silicon dice, and assuming:

1. a uniform carrier generation throughout the silicon.
2. the plasma thickness remains constant with variation in the carrier density.

However, in a practical situation, the light intensity decreases with increasing penetration depth, and less free-carriers are generated with decrease in depth of the silicon creating a conductivity gradient in the cross-section. A close approximation of this case was simulated by using a number of layers of silicon, each stacked on top of another and each with a higher conductivity than the previous. This complex design significantly increased the simulation time with no improvement in the results and so was considered to be unnecessary.

A lumped element equivalent circuit representing the silicon switch in a microstrip transmission line has been developed from first principles to fully characterise the active region within the switch architecture and to describe the device performance. A
gap in a microstrip transmission line is represented by a simple two-port model consisting of a configuration of a series gap capacitance with two shunt capacitances [34]. Building on the equivalent circuit of a gap in a transmission line (three capacitors C1, C2, C4), additional components are included to account for the photoconducting effect of the silicon, namely R2 and C3. Series inductors (L1 and L2) and resistors (R1 and R3) are added to account for losses in the switch. In order to maintain some symmetry in the switch C1 is identical to C4, R1 to R3 and L1 to L2. Lumped element circuits in the literature [35, 36] cover microstrip lines printed on a silicon substrate and so differ from this design as there is no conduction through the dielectric here. A diagram representing the layout of various elements mentioned above is shown in Figure 3.12

![Figure 3.12. An equivalent circuit of the optical silicon switch](image)

The values of the lumped elements are tuned in order to match measured S-Parameters under optical illumination varying from 0 to 200mW. An optimisation technique was adapted to derive the element values that agree with the measured frequency response of the microswitch. For illustration purposes, the comparison between the measured and equivalent circuit transmission coefficients are given in
Figure 3.13. Analysis of a number of such graphs shows with increasing optical power, the gap capacitances (C2, C3) increase whilst the gap resistances (R1, R2, R3) decrease, indicating it is a capacitive switch.

Figure 3.13. Comparison of measured, predicted (EM analysis) and equivalent circuit results for both the off and on state.
Figure 3.13. Analysis of a number of such graphs shows with increasing optical power, the gap capacitances (C2, C3) increase whilst the gap resistances (R1, R2, R3) decrease, indicating it is a capacitive switch.

![Graph of measured, predicted, and equivalent circuit results for both off and on state.](image)

Figure 3.13. Comparison of measured, predicted (EM analysis) and equivalent circuit results for both the off and on state.
3.10 Conclusion

Photoconducting switch designs have been presented. The basic structure of the switch consists of a microstrip transmission line incorporating a discontinuity and a semiconductor material across it. In the proposed designs the active region consists of a small high resistivity silicon dice. Three different topologies of the active regions have been studied and the characteristics and performance of the photoconducting switches for both the dark and optically illuminated states have been discussed. These switches are small, simple to fabricate and can be operated rapidly. There is a good agreement between the measured and predicted results. An equivalent circuit composed of resistors, inductors and capacitors has been found to accurately model the switch. The circuit has enabled the characterisation of the silicon dice incorporated in a transmission line under varying optical powers. This circuit can be a very useful tool in determining and ensuring optimum switch performance.

The optically controlled microswitch discussed in this Chapter has many potential applications. The potential applications this switch could be used for is in microstrip based devices such as filters and metamaterials, in order to alter the position of either the passband or stopband response of the device. The switch could be implemented in microstrip patch antennas to switch the radiated power as fast as 10.5 μs, and could also be incorporated as part of the phase shifting module (consisting of multiples switches) in steering the beam of an antenna. These potential uses were investigated in this thesis and are discussed in detail in Chapter 4 and Chapter 5.
3.10 References


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CHAPTER FOUR

4. Optically switchable microwave components

4.1 Introduction

This chapter presents the applications of optically activated Si microswitches discussed in Chapter three. The examples discussed are switchable parallel coupled line filter, tunable phase shifter and microstrip patch antenna (MPA). An optically tunable dipole antenna consisting of these silicon microswitches was also demonstrated by Panagamuwa et al [1]. These microwave devices are novel because they are simple to design, construct, and optically control. These devices are compact, and can also be easily integrated into microwave circuits, and would find a wide range of applications in communication systems requiring fast frequency tuning, and rapid phase change. The rate of phase change in tunable phase shifter, and the rate of frequency switching in the case of switchable filter and antenna is largely dependent upon the speed of the silicon switches. The switching speed is defined as the time
period from the moment the command is send to the switch to change state to the moment the insertion loss of the switched path reaches 90% of its final value [2].

Switchable filters can provide surprising benefits, especially for multimode portable terminals and tunable transceivers. Optically controlled phase shifters and switchable microstrip antennas with high switching speeds are required in a number of applications such as in scanning and imaging systems. However, switchability or tunability causes undesirable changes such as fluctuating of insertion loss, bandwidth and impedance matching. The optical injection technique adopted in this work to achieve switching does not lead to a severe degradation of the performance of the device.

The first section of this chapter presents a frequency agile parallel-coupled microstrip filter [3]. The center frequency of the device may be switched very rapidly between two discrete values by applying optical illumination to a pair of microswitches consisting of silicon dice, which are mounted on the printed side of the board. Increasing the level of optical power to 200 mW reduces the passband insertion loss by about 46% from 3.41 dB. The percentage bandwidth reduces by less than 2 % when tuning over the maximum frequency range.

The second section describes a study of microstrip resonators, and phase shifters both under static and optically illuminated states. This section also highlights some results obtained from 1-D (dipole) and 2-D (tripole) Metallodielectric Eletromagnetic Band Gap (MEBG. Stopband characteristics are observed when periodic structures, such as Frequency Selective Surfaces (FSS) elements are inserted between the grounded substrate and microstrip transmission line, and the resulting structure is known as
MEBG. These periodic structures can exhibit frequency regions where no electromagnetic waves are allowed to propagate [4]. The details of the applications of these MEBG structures in resonators and antennas will also appear in the second section of this chapter.

A study of a switchable microstrip antenna incorporating an optically controlled microswitch located in series with the feeding line is discussed in the third section [5]. Microstrip antennas are widely used in wireless communication systems because they are compact, lightweight, simple and cheap to manufacture using printed circuit technology. However these antennas present some disadvantages to the communication industry, and there have been several procedures developed over the past decades to overcome these limitations. An optically reconfigurable patch antenna with the aid of Si microswitch can be considered to address some limitations of the microstrip patch. The strength of the coupling across the gap of the switch is an important factor, which controls the switching performance. A good switchable antenna is characterised by a switch exhibiting low electrical coupling in the off state, and a low insertion loss when switched on. Four switches with different transmission line topologies were investigated. These topologies consisted of a single, stepped, interdigitated, and chamfered gap. Both simulations and measurements indicated that the antenna with the chamfered gap produced optimal results in terms of the best possible electrical isolation, match and insertion loss.

Some techniques using photoconductivity of silicon have been previously developed, and are introduced to these switchable components in order to connect gap discontinuities within microstrip. The switching techniques adopted in this work were
inspired by approaches used for analysis of optical control in microwave devices [6, 7]. However, the novelty of the research work mainly lies on the size and nature of the switches incorporated in these devices. The electromagnetic responses of the devices were controlled by using the photoconductive properties of the silicon embedded within the switches.

4.2 Switchable Coupled Line Filter

![Diagram of Switchable Coupled Line Filter]

Figure 4.1. Switchable first order filter using optically activated silicon dice. (not to scale) Dimensions are in millimeters.

The aim of the work presented within this section is to produce a switchable microstrip filter for microwave applications. The device presented is produced by modifying a single pole parallel-coupled line filter. This device is very simple to design, construct, and tune. Significantly, it also has the ability to be tuned very rapidly. The rate of frequency tuning is largely dependent upon the speed of the silicon switches. The subject of the switch's speed is discussed in Chapter 3. This device would find many applications in systems requiring fast frequency tuning. The
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tuning technique has also been employed to create a frequency tunable dipole antenna [1], and a reconfigurable electromagnetic bandgap structure [8].

The new switchable device is shown in Figure 4.1. A single resonator is employed in isolation in order to reduce system complexity. This device is fabricated, on a Taconic TLY-5 substrate, that is 1.19 mm thick and has a relative permittivity of 2.2 ± 0.2. The conventional device is modified by the addition of two short pieces (5.8mm long by 1mm wide) of microstrip transmission line. These sections of microstrip transmission line are located co-axially with the resonator and are displaced from its open ends by a short coupling gap. These gaps measured 0.2 mm. A single silicon dice, measuring 1 mm², is fixed above each gap [5]. The separation between the feedlines and the resonator lines is 0.5 mm.

When both silicon dice are in the off-state there is a considerable degree of electrical isolation between the transmission line sections and the resonator. Consequently, very little energy transfer occurs across the gap, and the resonant frequency of the device is approximately that of the conventional filter. When both silicon dice are illuminated, with a sufficient level of optical power, an electrical connection is formed between the transmission line sections and the resonator. The effective length of the resonator is thus extended by the sum of the lengths of the transmission line sections and two coupling gaps. Thus, the half-wavelength resonant frequency of the device now occurs at a lower microwave frequency. By switching the optical illumination on and off, the centre frequency is switched between two discrete values. The frequency value in the on state depends on the amount of the optical power supplied across the switches. A tuning range as high as 392 MHz is achieved from the above described
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device. In order to achieve the best performance each silicon dice requires a dedicated source of optical excitation. This excitation is provided by a Laser.

4.2.1 Switched First Order CLF Results

When both silicon dice are in the off-state the resonator (48 mm long by 1 mm wide) is electrically isolated from the extension sections and very little energy transfer occurs. One would expect this result because the isolation of the silicon switch is 17.12 dB at a frequency of at 1.799 GHz [1]. When both silicon dice are illuminated the extension sections are connected to the resonator, and its new length is 60 mm because the isolation of the switch is reduced. Consequently the centre frequency of the device is reduced accordingly. Measurement results [1] indicate that considerable energy transfer occurs, across the gap, at 1.799 GHz when the switch is illuminated with 200 mW of optical power because the insertion loss of the silicon switch is 0.64dB. A single infra-red (980 ± 5 nm) Laser is used to illuminate each of the two silicon dice. The optimum light delivery system employs a single dedicated Laser for each silicon dice. An alternative to this would be to employ an optical T-junction to split the light, from a single Laser, between two silicon dice.

The scattering parameter measurements presented within this section were obtained using an Anritsu 37397D Vector Network Analyser. In the following text, the device described as the reference incorporates a resonator measuring 60 mm. 60 mm is the effective length of the resonator in the switched device when both silicon dice are illuminated. The optimum level of performance for the technology employed can only be achieved by the reference device. The reference device does not contain optical switches. The reference therefore provides a standard against which to assess the
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performance of the switched device, when excited at various optical power levels. With the aid of an electronic control system it is possible to alter the amount of optical power produced by the Laser. Before making a microwave measurement both Lasers were set to produce the same level of optical power.

Figure 4.2(a) $S_{11}$ and (b) $S_{21}$ The measured scattering parameters for the reference and switched first order filter.
Figure 4.2 illustrates the measured scattering parameters for the reference device compared to those obtained in the off-state and with 10, 30, and 200 mW of optical excitations. From this figure one will observe that the level of optical power has a strong affect upon the losses incurred within the device, but less affect on the frequency shift. An increase in the optical power reduces the microwave signal losses incurred within the silicon dice. The bulk of the frequency shift occurs upon increasing the optical power from zero to 10 mW (299 MHz). Increasing the optical power from 10 mW to 30 mW increases the frequency shift by a further 82 MHz. Increasing the optical power from 30 mW to 200 mW results in only a marginal increase in the frequency shift (11 MHz). These measurement results indicate that one could almost achieve the maximum frequency shift of 382 MHz by using only zero and 30 mW of optical power. This suggests the possibility of utilising low power light sources such as Light Emitting Diodes (LEDs') for switching such a device. Technical difficulties may be encountered when trying to replace the Laser with and LED due to the divergent nature of the LED's light beam and because it emits a broader spectrum of wavelengths of light.

<table>
<thead>
<tr>
<th>Device/State</th>
<th>Insertion Loss (dB)</th>
<th>Return Loss (dB)</th>
<th>Centre Frequency (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Reference</td>
<td>0.394</td>
<td>25.94</td>
<td>1.806</td>
</tr>
<tr>
<td>200 mW</td>
<td>1.847</td>
<td>14.66</td>
<td>1.799</td>
</tr>
<tr>
<td>30 mW</td>
<td>3.407</td>
<td>10.04</td>
<td>1.810</td>
</tr>
<tr>
<td>10 mW</td>
<td>4.841</td>
<td>7.48</td>
<td>1.892</td>
</tr>
<tr>
<td>Off</td>
<td>0.601</td>
<td>23.13</td>
<td>2.191</td>
</tr>
</tbody>
</table>

The passband insertion loss, return loss and centre frequency for the reference and the switched device are shown in Table 4.1. The switched device is excited with different
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levels of optical power. Figure 4.3 shows the measured and simulated scattering 
parameters for the switched filter in both the on- (illuminated with 200 mW of optical 
power) and off-states. The simulations were performed using Microwave Office (from 
Applied Wave Research) Version 6.03. The 2.5D electromagnetic simulation engine 
within this package employs the Method of Moments technique.

![Graph showing measured and simulated scattering parameters for the switched filter](image)

Figure 4.3(a) and (b). The measured and simulated scattering parameters for the switched first order filter in both the off- and on-states respectively.
The percentage bandwidth (-3 dB level) of the switched first order filter in the off-state is 18.76%, whilst in the on-state it has decreased by 1.89%. The percentage bandwidth of the reference device is 13.04%. These results indicate that the bandwidth is relatively invariant under tuning. Figure 4.4 compares the power loss, as a function of frequency, within

the switched (illuminated with 200 mW of optical power) and reference devices. Figure 4.5 illustrates the power loss, as a function of frequency, within the switched device in the off-states. From Figure 4.5, it is seen that the on-state power losses, at the fundamental resonant frequency, exceed those associated with the off-state. In the off-state the peak power loss, within the vicinity of the fundamental resonant frequency, is 13.36% and occurs at 2.03 GHz.
An optically switchable microstrip filter has been demonstrated, both through measurements and simulations. The agreement between the measured and simulated results is good. The switching technique adopted in this work is novel and can be regarded as a good alternative to the conventional electrical and mechanical methods. The optical switches, presented here, are easily integrated within the device after manufacturing the circuit, using standard techniques. Laser diodes were employed as the source of the optical excitation. Measurement results indicate that the maximum frequency shift could almost be achieved by using a level of optical power as low as 30 mW. This suggests the possibility of utilising low power light sources such as Light Emitting Diodes (LEDs) for switching such a device. These results demonstrate the ability of the device to provide effective frequency tuning whilst maintaining a reasonably constant percentage bandwidth.
4.3 Resonators

Stopband characteristics are observed when periodic elements are inserted between the grounded substrate and microstrip transmission line. The two types of elements chosen to create a band gap are dipoles and tripoles. Figure 4.6 depicts a two-dimensional (2-D) structure which can produce a one-dimensional (1-D) Electromagnetic Band Gap (EBG) characteristics, comprising a 50 Ω microstrip line printed above a periodic array of dipoles which were in turn etched onto a RT-Duroid 5880 substrate having dielectric constant of 2.2 and a thickness, (S) of 1.125 mm. The dipole length (L), and width (W), are 15.5 mm and 1 mm respectively. The periodicities (Dy and Dx), rectangular lattice were 5 mm and 17.5 mm respectively. The band gap arising from such a planar array of conducting dipole elements printed onto a grounded dielectric substrate is shown in Figure 4.7, which clearly demonstrates that the dipole array can be used as a stop band structure, albeit one dimensionally. There is close agreement between the location of the band gap indicated by the measured and predicted results.

Tripoles arrays may be used as a two-dimensional MEBG because they are not linear in nature and therefore inherently prevent the propagation of surface wave modes in all directions. From the dispersion curves, it is noted that a tripole array based upon a triangular lattice exhibits good performance within the common TE and TM band gap regions, and produces an absolute band gap at a frequency determined by the dimensions of the array and unit element. An experimental structure was fabricated incorporating a microstrip transmission line printed above a tripole array (see Figure 4.8). The tripole array considered has its element spaced out periodically on two axes.
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separated by an angle $\alpha = 60^\circ$. The lattice periodicity $D = 12 \, \text{mm}$, tripole's arm length $L = 5 \, \text{mm}$ and the width $W=0.6 \, \text{mm}$. Scattering parameters measurements of the fabricated prototype indicate that the structure exhibits a bandgap in excess of 20 GHz with a cut-off frequency of around 9 GHz. The microstrip/tripole array structure along with the resulting bandgap are shown in Figures 4.8 and 4.9.

Figure 4.6. Cross sectional (top) and aerial (bottom) views of a microstrip transmission line printed above dipole MEBG.

Figure 4.7. Predicted and measured insertion loss for structure shown in Figure 4.6.
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separated by an angle \( \alpha = 60^\circ \). The lattice periodicity \( D = 12\, \text{mm} \), tripole’s arm length \( L = 5\, \text{mm} \) and the width \( W = 0.6\, \text{mm} \). Scattering parameters measurements of the fabricated prototype indicate that the structure exhibits a bandgap in excess of 20 GHz with a cut-off frequency of around 9 GHz. The microstrip/tripole array structure along with the resulting bandgap are shown in Figures 4.8 and 4.9.

![Microstrip Transmission Line](image)

**Figure 4.6.** Cross sectional (top) and aerial (bottom) views of a microstrip transmission line printed above dipole MEBG.

![Measured vs Predicted Insertion Loss](image)

**Figure 4.7.** Predicted and measured insertion loss for structure shown in Figure 4.6.
Figure 4.8. Microstrip transmission line printed above a tripole array to from an MEBG.

Figure 4.9. Insertion loss of MEBG shown in figure 4.8.
In addition to the creation of stopband characteristics phenomenon, a dipole MEBG may also be employed as a mechanism for the suppression of harmonics frequencies generated by a microstrip resonator. When a 5 GHz resonator replaces the microstrip line shown in Figure 4.6, its harmonic frequencies can be suppressed. The objective here is to remove the first harmonic at 10 GHz. The 5 GHz resonator is 19.25 mm long, 3.5 mm wide. The resonator is a series type, separated from the feeder lines by a similar gap size employed in switches (Chapter 3) of 0.25 mm. Two different topologies of the dipole array were investigated. The first MEBG resonator (Figure 4.10a) features an identical and symmetrical dipole array (L=15.5 mm) at each end of the resonator. The second MEBG resonator (Figure 4.10b) consists of a pair of unequal length of dipole arrays, and the dimensions of the dipoles on one end being reduced about 15% (i.e. L1=15.5 mm, and L2=13.18 mm). The most significant effect on the stopband of incorporating different sizes of dipole is the enhancement of the bandgap at 8 GHz. The dipoles underneath the resonators have been deliberately removed in both fabricated devices, in order to alleviate any strong coupling that may alter the resonant frequency and harmonic response. However, the objective and performance of the dipoles underneath the resonator will be discussed in the next paragraph. In Figure 4.11 (measured) and Figure 4.12 (predicted), reveal that an equal dipole MEBG eliminates the first harmonic while maintaining the resonance. However, this is accompanied by 2 dB increase in the insertion loss at 5 GHz. The discrepancies between the measured and predicted results primarily occur at higher frequencies, above about 6 GHz (this includes the harmonic at 10 GHz), and are largely due to spurious effects from the connectors as well as discontinuities at the edge of the prototype. Nevertheless, the suppression of the harmonic is clearly demonstrated. The resonator incorporating a dipole array has Q factor over three.
times that that of the conventional resonator. In the case of the resonator with unequal
dipole array situated either side of the resonator, two distinct but closely spaced
resonances appear and thus an increase of the bandgap is obtained. The Q factor is
also improved, with about a 3 dB reduction in the insertion loss at 5 GHz.
The objective of dipoles below the resonator is to shift downwards the fundamental
resonance in frequency by loading and packing the structure with the periodic array.
This evidently introduces slow wave propagation and creates a slow wave structure.
The performance of this device in terms of its insertion loss and group delay are
shown in Figure 4.13 and 4.14 respectively. Group delay is defined as the rate of
change of phase with frequency \((-\frac{d\phi}{d\omega})\) and is inversely proportional to the speed of
signal propagation [9]. It can be observed that a 1 GHz shift in the resonant frequency
increases the Q from 3.3 to 7.4. The group delay of an ordinary microstrip line
resonator is 1640 ns. This is increased to 2950 ns when an MEBG is introduced below
the transmission line. This is a similar behaviour to the velocity of wave propagation
on meander line CPW [10]. It was observed that packing the dipoles more closely
and/or increasing their lengths, the resonance shifts further. Although the results were
limited to a few cells in the direction perpendicular to the propagation, simulations

Figure 4.10. Single resonator geometry with (a) single and (b) dual dipole MEBG.
and further tests have shown that one set of dipoles is sufficient to obtain the same effects. This will have an impact on the size of the device when realised in practice.

Figure 4.11. The measured insertion loss for the enhanced microstrip line resonator.

Figure 4.12. Predicted insertion loss for microstrip line resonator with and without MEBG.
Figure 4.13. Insertion loss of resonator/MEBG device.

Figure 4.14. The group delay measurements of the resonators.
4.4 Switchable EBG Phase Shifter

Figure 4.15 shows an EBG resonator similar to the one discussed in the previous section. The central portion of the horizontal microstrip line represents the resonator etched onto a thin dielectric laminate. This laminate is mounted above a dipole EBG array, and a couple of optically controlled switches are mounted above the resonator discontinuities. The purpose of these micro-switches is to control the phase delay of a signal propagating along it. As is well known, RF energy is coupled into and out of the resonator by means of two feeding microstrip lines. Each line is separated from the resonator by a gap size of 0.25 mm. Switches using optical control made up of silicon dice are used to bridge the gaps. Silicon dice placed above resonator gaps act as the two phase shifting elements controlled by optical illumination. These micro-switches are used to alter and control the propagation constant ($\beta$) of the microwave signal along the transmission line, and the value of $\beta$ is almost double that of a normal microstrip line by incorporating periodic elements (EBG) underneath the transmission line [11].

The analysis presented here suggests the structure in Figure 4.10 may find application as a novel optical phase shifter. The functionality of this optically controlled phase shifter is based on the dielectric properties of silicon substrate incorporated in microstrip resonators. Wireless communication systems may benefit from EBG device development in a number of different scenarios. The presence of the EBG within the structure makes it a slow wave device, improves the Q factor of the resonator, suppresses harmonics of the fundamental frequency, and also suppresses surface waves. The first three advantages of implementing EBG have already been discussed, and the surface wave propagation in EBG materials is discussed later in
this chapter. The purpose of separating the resonator from the EBGs is to prevent shorting the conducting dipole array from the resonator. The EBG layer is made up of a periodic array of dipoles etched onto a Taconic (TLY - 5) substrate. This is an alternative technique to the conventional method of either drilling holes, or etching pattern in the ground plane. Prototypes consisted of three rows of the EBG, and the resonator is placed perpendicular to the middle row, as this is the most influential row. The resonator is etched onto a thin dielectric laminate, and is mounted on the Taconic substrate using non-conductive glue. Experiments revealed that the intensity of the optical source determines the level of the switching and the corresponding delay change. Another factor that influences the switching level is the nature of the switch topology. Even low power optical sources such as LEDs can be used to effect a significant phase and delay change.

Figure 4.15. Photograph of resonator with EBG
4.4.1 Optical switch topologies in optically tunable phase shifters

The topologies of the types of microstrip gaps incorporated in optical phase shifters that have been investigated are illustrated in Figure 4.16. Figure 4.16(a) depicts a straight gap with chamfered ends. This type of microstrip gap has been implemented in microswitches discussed earlier in the preceding chapter, and it has been established that this type of gap has superior scattering characteristics over the commonly used simple straight gap or discontinuity. The performance of the chamfered gap was investigated using two phase shifters assigned with the serial numbers PH025 and PH04. Devices PH025 and PH04 were identical apart from the gap size along the straight portion. The gap sizes of PH025 and PH04 were 0.25 mm and 0.4 mm respectively. Silicon dices measuring 2.0 mm x 1.0 mm x 0.3 mm, (LxWxH) was bonded over these gaps using silver epoxy.

Figure 4.16. Illustration of (a) chamfered and straight gap, (b) chamfered and interdigital, and (c) straight and interdigital gap. (not to scale)
Another batch of PH025 and PH04 were designed and built, but the technique of bonding the silver epoxy was different. The technique of bonding the epoxy was not only limited to the sides of the silicon dice, but it was extended over the silicon as well. Figure 4.17 illustrates how this was achieved, and thereby creating two gaps $W_s$ and $W_g$, one above the other. The purpose of making $W_g > W_s$ is to reduce the surface area of illumination in order to reduce the insertion loss when the switch is activated. $W_g$ in two prototypes built, B2025 and B204 were 0.25 mm and 0.4 mm respectively. The width of the gap above the silicon dice $W_s$ was about 1.0 mm wide.

![Silver epoxy overhanging](image)

Figure 4.17. Illustration of the silver epoxy overhanging the silicon dice, $W_g > W_s$.

The type of discontinuity illustrated in Figure 4.16(b) can be described as chamfered and interdigital. The interdigital gaps (IDGs) exists only across the non-chamfered region. Prototypes built having interspacing sizes of 0.25 mm and 0.40 mm were investigated, and these devices bear the serial numbers IDG2 and IDG4 respectively. The other analysed gap type consisted of a simple interdigital gap, IDG3, see Figure 4.16(c). IDG3 has an interdigital gap size of 0.25 mm. The dimensions of the silicon dice placed above the IDGs measures 2 mm x 2 mm x 0.3 mm (LxWxH).
4.4.2 Optical sources and results

The performances of the phase shifters were evaluated using illumination from laser diodes (LDs) and Light Emitting Diodes (LEDs) running at various optical power levels. The use of LDs require safety precautions in place, whereas LEDs do not demand stringent safety measures. Examples of the measured results from PH025 and IDG3 using these light sources are discussed in this section.

The illumination from the LDs was focused above the silicon dice by using 1.0 mm diameter fibre optic cable. The use of fibre optic cable allows an efficient light delivery to the silicon switch. The fibre optic cable does not interfere with the performance of the device. The amount of optical power was varied by changing the driving current to the LDs. Measurements were carried out when the LDs were emitting power between 0 and 200 mW. The transmission coefficient results from Figure 4.18 indicate that the application of this device is not only limited to a phase shifter, but it's uses can be extended to act as an optical microwave switch. The region it can be regarded as a switch is between 1 GHz and 3 GHz. The insertion loss at 2 GHz in the ON and OFF states are 1.5 dB and 42 dB. The insertion loss with the LDs ON can almost be maintained near the resonance frequency, and can be as low as 0.7 dB at 4.2 GHz. The relationship between the insertion loss and optical power from LDs at various frequencies is shown in Figure 4.19. The insertion loss values of various devices lie between 0.4 dB and 3.0 dB for the optical illumination of at least 50 mW. The device, B204, with a switch consisting of two stacked gaps gives the least insertion loss as predicted. The corresponding differential phase of the magnitude between 40 degree and 300 degree is given in Figure 4.20.
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Figure 4.18. Measured transmission responses at different optical LD power levels of phase shifter PH025.

Figure 4.19. The relationship between insertion loss and the controlling optical power from LDs at different frequencies.
Figure 4.20. The relationship between phase change and the controlling optical power from LDs at different frequencies.

Figure 4.21(a). Variations of the transmission coefficients of phase shifter IDG3 with changes in the LEDs supply voltage.
Figure 4.21(b). Variations of the phase of IDG3 with changes in the LEDs supply voltage.

Figure 4.21(c). The variation of insertion loss and phase of IDG3 at different frequencies.
The LEDs in a TO-39 package were placed directly above the silicon dice. It can be observed from Figure 4.21 that the presence of LEDs does not adversely affect the static state of IDG3. The LED illumination was controlled by changing its supply voltage, between 0 and 1.65 V and the performance of the manufactured devices with respect to the LED supply voltage was investigated. The typical forward voltage, $V_F$, and forward current, $I_F$ of this type of LEDs are 1.65 V and 500 mA respectively. These figures suggest the possibility of employing portable and compact power sources. The peak emission wavelength of this type of LEDs is 880 nm.

The manner in which the insertion loss changes with voltage driving the LEDs is indicated in Figure 4.22 shows. The insertion loss increases with an increase in supply voltage until 1.30 V, and then starts to gradually reduce as the voltage increases. The relationship between insertion loss and the controlling optical power from LEDs at different frequencies from the value where the loss starts to decrease is shown in Figure 4.23. These results indicate that it is possible to control the phase linearly with an increase in the supply voltage.

We also investigated the performance of these phase shifters with LEDs being driven with a supply voltage below 1.30 V. It was observed that it is again possible to operate the phase shifter linearly with the LEDs running even at low voltages. A phase change of more than 120 degrees, (sufficient to steer a beam 45 degrees), and with insertion loss as low as 3.5 dB was measured at 3.37 GHz.
Figure 4.22. The relationship between insertion loss and the controlling optical power from LEDs at different frequencies.

Figure 4.23. The relationship between insertion loss and the controlling optical power from LEDs at different frequencies.
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4.5 Microstrip Patch Antenna (MPA)

4.5.1 General properties of MPA

A MPA is a planar antenna fabricated using "Microstrip." MPAs are relatively easy to design and fabricate using standard printed circuit technology, and thus have made it possible to integrate active microwave components and antennas onto a single microwave printed circuit. This has led not only to a compact structure, but has also given rise to the development of a new class of components known as active antennas. The objectives of this section are twofold: to demonstrate the ability to integrate the MPAs with EBGs to solve some of the inherent problems of MPAs, and to investigate the performance trend of the antenna with optical switches. The principle of operation of MPAs is simple. Portions of the patch act like slots, with respect to the ground plane. The 2-D patch resonates in one dimension, and at a half wavelength, and radiates in the other dimension. The radiation is due to the strong electric fields and high ground plane currents that exist at the edges of the patch. The patch resonates along its length, and radiates along its width [12-14].

4.5.2 Advantages and disadvantages of microstrip antennas

MPAs are employed in a wide variety of communication systems due to their numerous advantages, including low geometrical profile, light weight and ease of integration with passive and active microwave devices. Various forms of microstrip patch antenna are used for applications ranging from satellite antennas to small handset terminal antennas for mobile communication systems. However, despite its
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acceptance in many main stream applications, the conventional form of this printed antenna still suffers from several important limitations. These limitations include its relatively narrow impedance bandwidth, antenna size, spurious feed radiation and problems associated with the excitation of surface waves. Over the years researchers throughout the world have developed several techniques with varying degrees of success to overcome these disadvantages. In particular, researchers have investigated ways of improving the bandwidth, reducing the size of the device, efficiently integrating the antenna with MMIC technologies and also enhancing the efficiency of MPAs [15-17]. Due to the inherently narrow band characteristics of microstrip patch antennas, a considerable amount of effort has been expanded in order to reconfigure microstrip patch antennas [18, 19]. A reconfigurable microstrip patch antenna allows not only for relatively wide-band frequency diversity but also for the electronic correction of errors introduced by changes in temperature, environment, and manufacturing imperfections.

4.6 Practical applications of EBGs in antennas

It is well known that one of the main mechanisms for loss within a microstrip patch antenna is surface waves [4]. This surface wave loss is mainly due to TM₀ surface...
wave. This particular mode has no cut-off frequency. Surface waves are modes of propagation supported by a grounded substrate. Generally in antennas, surface waves spread out in a cylindrical pattern around the point of excitation, with field amplitudes decreasing with distance. Surface waves are reflected within the grounded dielectric substrate as shown in Figure 4.24. The fields remain trapped within the dielectric and take up part of the energy of the transmitted signal, thus decreasing the desired signal amplitude and contributing to degradation in the antenna efficiency. These waves propagate under the surface of the substrate and are radiated into free space at the truncation of the substrate. These radiated waves can significantly distort the antenna pattern by producing ripples in the radiation pattern, raising sidelobe and cross-polarisation. The objective of this work was to enhance the performance of a patch antenna by surrounding it with an EBG array that exhibits an absolute band gap (forbid the propagation of electromagnetic waves) in the region of the operating frequency. This 2-D periodic array will prevent the TM₀ mode surface wave from propagating in any direction within the structure. Therefore, as desired the patch antenna radiates more energy. To test this concept a microstrip patch antenna surrounded by a tripole EBG array was fabricated on FR4 substrate as shown in Figure 4.25. The measured results revealed return loss improvement, Figure 4.26, and smoother radiation pattern, Figure 4.27 than the patch without the MEBG due to the suppression of surface waves being more prominent as a result of higher dielectric constant. Simulated results in Figures 4.28 and 4.29 show a comparison between a rectangular microstrip patch antenna without and with tripole MEBG in terms of the currents displayed on the surface of the microstrip substrate. The reduction in surface waves is clearly visible. It can be seen that the presence of the tripole MEBG around
the patch suppressed most of the surface currents, which is a desirable feature in the reduction of the patch antenna losses.

Figure 4.25. Patch antenna and tripole elements printed on the same substrate layer.

Figure 4.26. Return loss for patch antenna printed on FR4 with and without tripole array.
Figure 4.27. H-plane radiation pattern measurements for patch antenna on FR4 with and without MEBG.

Figure 4.28. Surface waves display surrounding a rectangular microstrip patch antenna alone.
Figure 4.29. Surface waves display surrounding a rectangular microstrip patch antenna with tripole MEBG.
4.7 Optically Switched Patch Antenna

This section describes the performance of the microstrip patch antenna when connected in series with a photoconductive switch. The layout of the structure is shown in Figure 4.31. The performance of the switch alone in isolation has been evaluated in chapter three. The performance of the active antenna depends mainly on the impedance match between the switch and the transmission line. The switch, fabricated on an RT Duroid 5880 substrate, consists of a 50 Ω microstrip transmission line to facilitate maximum power transfer from the antenna. High resistivity silicon dice bridges the microstrip gap as shown in Figure 4.30. The dimensions of this silicon dice were such that the gap on the line was either partially covered or completely overlapped depending on the topology of the switch. The size of dice was governed by dimensions relating to a 50 Ω impedance across the switch. Three switch topologies shown in the inset of Figure 4.31 were investigated. It is possible to reduce the value of gap capacitance and also to increase the dark state isolation by chamfering and notching the microstrip line at the gap. For the purpose of experimentation, each switch was placed in a brass box, (Length $L = 30$ mm width $W = 15$ mm, and height $H = 10$ mm). This switch module was connected in series with microstrip feed of an antenna in order to realise a smart antenna that could be optically switched.

The performance of the reference, offset fed and inset fed microstrip patch antennas is illustrated in Figures 4.32 and 4.34 respectively. The values of the return loss at resonance were below 20 dB. A good agreement between the measured and predicted results was noted. The microstrip patch antennas were fabricated on a 1.125 mm thick RT Duroid substrate having a dielectric constant of 2.2. These planar antennas are
depicted in Figures 4.33 and 4.35. Each antenna was 50.0 mm long, and 59.3 mm wide. The resonant frequency of the offset-fed antenna was 1.68 GHz, and the feed line was located 16.3 mm away from the radiating edge. This technique for feeding the antenna yields multiple resonances between 1 and 3 GHz. The second and third harmonic frequencies were located at 2.0 GHz, and 2.63 GHz respectively.

The second technique for feeding the microstrip patch antenna was to recess the transmission line to a distance of 14.0 mm as shown in Figure 4.35. Unlike the antenna incorporating an offset feed, the design featuring an inset feed produced a single resonance at around 2.0 GHz. The inset-fed antenna was also employed to investigate the possibility of switching on and off the patch optically. This optically controlled antenna was realised by placing the patch in series with the microswitch discussed in chapter three.

Figures 4.36(a) and 4.36(b) show both the predicted and measured results for the patch antenna under dark and optically illuminated states. A conventional microstrip patch antenna (identical patch antenna without the switch) was used as a reference antenna. This reference antenna was considered as a benchmark radiator. At resonance the value of return loss for the patch and switch under the dark state is almost 0 dB, but two minima at 1.92 GHz and 2.1 GHz exhibit return loss values of -2.5 dB. The patch incorporating the switch under the illuminated state exhibits a similar return loss to the reference patch, and both are resonant at 2.0 GHz.

The boresight gain measurements for the optically controlled antenna are compared to the reference patch in Figure 4.37. As expected, noticeable switching takes place at the operating frequency of 2.0 GHz and the on/off ratio is about 15 dB. Power dissipation within the silicon reduces the amount of power radiated from the
illuminated compared to the reference device. In the off state, the patch is not completely turned off due to coupling across the gap of the switch. This effect leads to radiation 1.92 GHz, and 2.1 GHz.

Figure 4.30. Photograph of a microwave (Si) microswitch.

Figure 4.31. Schematic diagram of a patch antenna. The insert shows the three switch topologies.
Figure 4.32. $S_{11}$ for offset fed microstrip patch antenna.

Figure 4.33. A picture of the offset-fed microstrip patch antenna fabricated on RT Duroid substrate for the purpose of experimentation.
Figure 4.34. S11 for inset fed microstrip patch antenna.

Figure 4.35. A picture of inset-fed microstrip patch antenna fabricated on RT Duroid substrate for the purpose of experimentation.
Figure 4.36. Return loss for the patch antenna under the dark and optically illuminated states.
Figure 4.37. Measured boresight gain for the optically switched antenna and reference antenna.
4.8 Conclusion

A simple optically switchable microstrip filter has been demonstrated, both through measurement and simulation. There is good agreement between the measured and simulated results. The center frequency of the device may be switched very rapidly between two discrete values by applying optical illumination to a pair of microswitches consisting of silicon dice, which are mounted on the printed side of the board. Measurement results indicate that one may achieve a frequency shift of 392 MHz by illuminating both dice with about 30 mW of optical power leaving the source. Increasing the level of optical power to 200 mW reduces the passband insertion loss by about 46% from 3.41 dB. The measured return loss values when both dice are illuminated with 30 mW and 200 mW are 10.04 dB and 14.66 dB respectively. The percentage bandwidth reduces by less than 2% when tuning over the maximum frequency range, stated above. The switching technique adopted in this work is novel and can be regarded as a good alternative to the conventional electrical and mechanical methods. The filter may easily be patterned using standard pcb fabrication techniques.

A periodic dipole array can be effectively used as 1-D MEBG, whereas a tripole array can be used as 2-D MEBG, for use in resonators, antennas and filtering applications based upon microstrip technology. It is known that for a microstrip patch antenna on high dielectric constant (> 10) substrates, the surface wave losses become even more prohibitively large and thus the antenna becomes less efficient. The MEBG structures shown in this chapter can be used to improve the performance of a patch antenna.
A novel optically tunable phase shifter has also been demonstrated, achieving more than 120° phase change between the on and off states. Optical switching of a patch antenna has been demonstrated by employing a microswitch based upon a silicon dice. 15-dB isolation of has been achieved while the antenna remained matched to the feed line impedance. At resonance, the reduction in the boresight gain is primarily due to the dissipation of energy within the switch, which can be improved by enhancing the better ohmic contact between the silicon and microstrip lines. The EM model used to describe the optical excitation of the devices, represents switching action well since there is a good agreement between simulation and measurements.
4.8 References


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CHAPTER FIVE

5 Tunable Metallodielectric Electromagnetic Band Gap (MEBG) Based Structures with Defects

5.1 Introduction

This chapter is concerned with the integration of Si micro-switches investigated in Chapter Three directly into EBG structures in order to control the band gap. Defects were introduced within a periodic arrangement of EBG elements. The ability to control the propagation of electromagnetic waves within the band gap by introducing appropriate defects into the structure makes the EBG very attractive for the fabrication of various devices. The work on EBG structures has been focused on investigation of steady state structures [1-3], however in this chapter a study into a tunable planar EBG based structure, including incorporating defects within the structures is discussed. EBG structures exploiting fixed or tunable defects have received much attention because of their ability to operate as narrow or wide band filters. With the recent interest in reconfigurable wireless devices, the demand for controllable narrow or wide band filters is increasing. By altering the defects in an
EBG structure, the transmission properties of the device can be adjusted. Microwave band gap devices incorporating defects have been simulated, and to verify the conclusions drawn from the numerical computations various prototypes were fabricated, and tested [4]. Furthermore the tunability of the structure has been studied. Tuning was achieved by controlling with the aid of Si micro-switches the dimensions of the dipole elements within an EBG structure. These techniques yield an improved stop band characteristics in terms of its width and depth. The tunability is demonstrated using an optical injection technique on high resistivity (typically \( \rho > 6 \text{k}\Omega \text{cm} \)) silicon switches placed across regions with defects. By controlling intensity of optical illumination focused on a region of silicon above the defects, the dimensions of the elements within the EBG structure can be adjusted. This leads to changes in the transmission properties of the structure. When silicon is illuminated by a light source, a plasma of electron hole pairs is generated of a certain depth and concentration. The denser the plasma is, the closer to a metallic state the semiconductor gets. The extent to which this transformation occurs depends primarily on the carrier lifetime and light intensity. These photoconductive properties of silicon are utilised to alter the dimensions of the elements resulting in the reconfigurable stopband.
5.2 Excitation of MEBG

![Diagram of EBG structure with dipoles](image)

Figure 5.1. A structure of the microstrip EBG line with dipoles etched into the copper layer above the substrate. (This figure is not to scale)

Two main approaches for realising EBG structures, based on microstrip lines or metallic elements (hence the term Metallo-dielectric Electromagnetic Band Gap, MEBG) have been published. The first method is achieved by etching periodic patterns into the ground plane [5], and the second technique is the approach of drilling holes into the substrate layer [6, 7]. In this work the structures have been realised by placing a transmission line above a periodic array of dipoles, or tripoles, printed onto grounded dielectric substrate. The periodic dipole and tripole arrays produce a 1-D and 2-D EBG characteristics respectively. This is as an alternative to conventional methods of manufacture [8]. The conventional EBG manufacturing methods have the drawback that the manufacturing processes are not easy (vias are necessary) and incompatible with those used to produce monolithic microwave integrated circuits manufacture. Figure 5.1 shows the structure of a microstrip line with an array of dipoles etched onto the substrate. The transmission coefficient of such a device is characterised by a band gap or stop band. In the limiting case where the dipole length is zero, there is no band gap and the structure is a standard microstrip line.
5.3 Propagation in a microstrip line

Figure 5.2 depicts the electric fields in the air region above a microstrip line. The majority of the Electric (E) and Magnetic (H) fields are confined in the dielectric substrate between surface conductor and the ground. Under normal operating conditions (in the absence of the EBG array) the transmission line exhibits an (S21) of approximately 0 dB. This indicates that an input signal will propagate along the line with very little loss or reflection. This situation can only occur when the characteristic impedance of the line is matched to the input and output impedances. The presence of the EBG array causes a change in the characteristic impedance of the microstrip line and introduces a stopband due to the filtering action of the array. EBG structures have been realised adopting the microstrip transmission line as the excitation of electromagnetic waves. In EBG substrates electromagnetic waves behave in the same way that electrons behave in semiconductors. Periodic arrays placed over a ground plane can exhibit very high surface impedance values for certain frequencies and hence can be used as artificial magnetic conductor (AMC) surfaces [9].
5.4 EBG structures with defects and fabrication process

The periodic arrangement of the EBG elements create a frequency band gap [10]. An example of an EBG structure is illustrated in Figure 5.3, and consists of an array of 10.5 mm long, and 1 mm wide dipoles having a periodicity of 5 mm. These dipoles are etched on TLY-5 Taconic substrate, \(\varepsilon_r = 2.18, h = 1.17\) mm. A 50 \(\Omega\) transmission line etched onto a 0.07 mm high dielectric of was placed directly above a centre line of the array. The structure was mounted onto an aluminium base. The aluminium provided mechanical support and held the connectors firmly to the circuit. In order to verify the performance of EBG with defects, configurations of this kind were analysed using a commercial software package Microwave Office. The performance of these configurations were compared to the behaviour of normal EBG structures. The following different topologies of defects were investigated:

5.4.1 Defects type A and B

![Figure 5.3. The construction details of the prototype. The insert picture shows the microstrip line and dipole etched on substrates.](image-url)
The purpose of the defects introduced into the EBG structure was to control or tune using optical techniques the position of the stop band region. The approach of line extension or switched dipole was adopted to reconfigure the EBG device. The first form of defect considered, (type A), is shown in Figure 5.4. In this structure, adjacent dipoles alternate in length between 14 mm and 10.5 mm and have a periodicity of 5 mm. The surface area of each defect is 1.75 mm$^2$. This area corresponds to the space left after reducing the length of the other dipole by 3.5 mm.

The second defect, (type B), to be investigated consists of discontinuities or gaps in the dipole. Figure 5.5 show an aerial view of an EBG structure illustrating a 0.25 mm wide gap (Surface area of the defect or gap is 0.25 mm$^2$). Each defect is located 1.5 mm from the end of the dipole.

### 5.4.2 Defects type C1 and D1

The device incorporating defects type C1 is identical to the structure featuring with defects type A (discussed above), with an additional dipole of length 12 mm placed between the 10.5 mm and 14 mm dipoles. The dipole lengths of the structure incorporating defects type D1 were 7.5 mm, 12.0 mm and 14.0 mm. Figure 5.9 shows an undulating envelope formed by the dipoles. The outline of these defects is symmetrical. The depth of the troughs and peaks of the outline are made equal in order to achieve the desired symmetry. Unlike the previous work, [8, 11] which employed sinusoidal patterns etched into the ground plane, the approach taken here is unique in the sense that the structure is reconfigurable by virtue of lengthening the dipole elements. This can be achieved by optical excitation.
5.4.3 Defects type E1
The arrangement of dipoles in the structure containing defect type E1 is displayed in Figure 5.10. The device consists of dipoles of lengths 7.5 mm and 14 mm, and width 1 mm. A pair of each length is arranged in succession starting with the smaller dipole.

5.5 Static case results

Comparisons between EBG devices with and without defects were performed in order to gain insight into the effect(s) of the presence of the defects. The performance of the normal structure was regarded as the reference and benchmark of the reconfigurable device. The predicted and measured results for a structure incorporating six dipoles and 5 mm apart are shown in Figure 5.11. Each dipole is 10.5 mm long and 1.0 mm wide. A 2-D MoM method from Microwave Office.

Figure 5.11. Predicted and measured transmission response of a 10.5 mm EBG
simulation package was employed in order to get the predicted results. The transmission response of this structure exhibits a band gap between 12.5 GHz and 17.5 GHz when measured at -10 dB level. A similar structure consisting of 14.0 mm long dipole elements was also fabricated and measured. The results of the 10.5 mm and 14.0 mm EBG are compared in Figure 5.12. The 14.0 mm EBG exhibited a band gap at a lower frequency than the 10.5 mm EBG. At -10 dB, the band gap lies between 10.25 GHz and 14.0 GHz.

The predicted and measured results for the structure incorporating defects type A and B are shown in Figures 5.13 and 5.14 respectively. These structures exhibited wider band gaps than either those measured for 10.5 mm or 14.0 mm EBG. The pictures of the structures illustrate the strips of silicon inserted within the regions incorporating defects. The objective is to generate plasma within these regions by means of optical illumination. The reference structure is an EBG constructed from uniform dipole array of 14.0 mm long.

![Figure 5.12. EBG transmission responses for a uniform dipole array featuring dipoles of different lengths.](image-url)
Figure 5.13. EBG transmission responses for a device incorporating Defect type A.

Figure 5.14. Transmission coefficient of the structure incorporating type B defects, and the reference device.
Figures 5.15 and 5.16 show the surface wave response of defected EBG structures comprising of two or three different length dipoles. It was discovered that reducing the length of some of the elements in the array it was possible to control the size of the band gap. For example the structure incorporating dipoles 7.5 mm, 12.0 mm and 14.0 mm long suppressed frequency harmonics above 17 GHz generated by a similar structure made up of dipoles 10.5 mm, 12.0 mm and 14.0 mm long.
5.6 Reconfigurable EBG Structures with defects

Optical tuning of an EBG device was demonstrated experimentally using pieces of silicon die placed within the defected regions. The position of the bandgap was altered by illuminating silicon die. The optical excitation controls the complex permittivity, and hence the complex impedance according to the strength of the applied optical energy. This tuning technique takes advantage of the photoconductive properties of the semiconductor, an area that has been well-researched [12]. The silicon semiconductor used exhibits a high value of dark resistivity. 5 mm long strips of silicon equal in width to the dipole were placed in the region of both type A and B defect, see, for example Figures 5.13 and 5.14. The objective is to change the physical properties of silicon from semiconductor to a pseudo-metallic state under optical excitation. The EBG structures incorporating defects type A and B were selected to demonstrate the optical control of the frequency band gap. An array of six

![Figure 5.17. EBG tuning using LEDs as the source of optical illumination.](image-url)
low power LEDs constituted the source of optical illumination, each LED was placed directly above a defect, and was enclosed, within a standard TO-39 package. The LEDs were driven with 0.5 Amps.

The effect of switching the type A defect on and off is shown in Figure 5.17. The graph indicates a shift of about 45% and 30% measured at -15 dB level (compared to the target) when the LEDs were driven using 0.6 Amps and 3 Amps respectively at the lower -10 dB level. The target was the performance measured from a structure consisting of a uniform dipole array having a length of 14 mm. The measurements reveal an increase of more than 30% in the depth of the band stop when the defects are activated. This type of defect mainly affects the falling edge of the band gap. The measured results for the structure incorporating defects type B (see Figure 5.18) demonstrate the reconfigurability of the device under optical illumination. The device delivers a shift of the bandgap similar to that of the structure incorporating type A. If the LEDs are excited with a current of 0.6 Amps the frequency shift is reduced suggesting that most of the energy is absorbed in silicon.
5.7 Conclusion

A theoretical and experimental study has been conducted to investigate EBG structures incorporating defects in both the static and optically illuminated cases. The defects considered are in form of element discontinuities or varying element dimensions. It was noted that the band gap position can be manipulated by the use of defects. Some simple EBG array configurations incorporating these defects have been proposed. The technique of implementing EBG with defect can be considered as an alternative method of tuning EBG structures, and a novel approach for the control of wave propagation in EBG structures. The performance of some defected structures were compared to their counterparts structures without defects. These conventional structures were regarded as reference structures. It was noted from measured results that the width of the stopband of a structure incorporating defect Type A was about 1.0 GHz wider than the reference device, whilst defect Type B produced a stopband width about 1.5 GHz wider than the reference. Defects Type A and B gave a stopband shift of about 15% and 13% respectively. It was also established that wide stopband greater than 1.5 GHz can be achieved using defect Type E. This defect produced a stopband width as high as 8 GHz. Simple reconfigurable EBG devices using illuminated silicon pieces were demonstrated. These silicon pieces implemented were either used as line extensions or switches in order to change the configuration of the array pattern resulting in the reconfigurable stopband. The EBG architecture consisted of a transmission line above a periodic array of conducting dipoles on grounded dielectric substrate. Low power six LEDs were employed as the light source, and these LEDs were capable of producing stopband shifts of more than 500 MHz. The main factors, which affect the extend of stop band shift, are the
number of defects within the structure, the optical power intensity employed, and the configuration of the defects.

The results presented in this Chapter suggests that EBG structures with defects have promising applications as novel reconfigurable EBG devices for use in various microwave circuits such as filters, microstrip antennas and phase shifters. Exploitation of the EBG structures is becoming attractive in communication systems.
5.8 References


CHAPTER SIX

6.0 Discussions and Future Work

6.1 Review of the research study

This thesis has described the theory and practical analysis of a novel type of a photoconductive switch. The main contributions of this work are an investigation of a small optically activated switch, and the employment of this microswitch in the control of communication systems. Potential applications for the switch have been presented, and examples have been demonstrated experimentally in the following devices:-

1. frequency switchable microstrip filter
2. optically controlled resonator and switchable phase shifter
3. reconfigurable microstrip patch antenna, and
4. tunable Electromagnetic Band Gap (EBG) structures
This research has shown that a pair of optically controlled microswitches can be employed in a microstrip coupled line filter to switch its resonant element. This simple optically switchable microstrip filter may easily be patterned using standard PCB fabrication techniques. The optical switches are incorporated following this production stage. The optical excitation could be provided using optic fibre cable that offer low electrical interference. It was found out that one may achieve a shift of 359 MHz in the centre frequency by illuminating both dice with only 30 mW of optical power.

By cascading two identical microswitches along a single section resonator above a metamaterial surface, it is possible to produce an optically controlled EBG phase shifter. This device would be very desirable in many various microwave applications such as phased-array antennas, cameras, and scanners [1, 2]. Phased-array antennas often require phase shifters, which are continuously variable and that offer fast speed in order to optimally control the main beam and null directions [3, 4]. It has been assumed that the operational speed of the optically controlled EBG phase shifter depends on the speed of the microswitches that has been discussed in detail in Chapter 3. Two optically activated switches within the phase shifter can be operated with optical sources emitting low optical intensity such as LEDs.

A reconfigurable patch antenna was realised by incorporating a photoconductive switch along a recessed transmission line. In this study, the switch performance was initially evaluated in isolation, as well as investigating the practical mechanisms for minimising losses, and coupling effects across the gap in the switch. It was found out
that the electrical isolation provided by the switch in the static state can be increased by chamfering and tapering the gap. By interleaving or creating interdigital fingers within the gap made it possible for the switch to achieve low values of insertion loss at low optical power in the active state.

It has also been shown EBG structures can be realised by fabricating periodic array of metallic elements on a ground plane. This method is compatible with Integrated Circuit (IC) fabrication techniques [5, 6]. The metallic elements prohibit the propagation of electromagnetic fields within the bandgap frequencies. EBGs for dipole and tripole arrays were investigated. It was discovered that the dipole array exhibits a stop band across a large range of propagation direction but does not provide an absolute band gap due to the polarisation of the dipole elements. The physical shape of the structure does however enable the array to be packed closely together and thus increasing the packing density. The dipole array can be used as a planar 1-dimension unit cell where the propagation is strictly perpendicular to the length of the dipole, whilst the tripole array can exhibit an absolute, and stable band gap.

An analysis of the performance of EBG structures with and without optical control has discussed. Simulations and measurements were performed and the correlation between the measured and predicted results was good.
6.2 Future research

A pilot study has been undertaken to investigate the design and potential applications for using photoconductive switches to optically control microwave devices. Although potential applications of the microswitch have been presented, further investigations are still required in order to fully test its capabilities. The areas that could be researched are as follows:

1. Power handling capability
2. Mean Time Before Failure (MTBF)
3. Scaling current design for operations above 3 GHz
4. Improving isolation and insertion loss

MEMS exhibit high levels of electrical isolation and low insertion loss. In order for the switch to be a serious contender in the market it would need to offer equal or better performance than traditional ones.

The microswitch has the capability of being a multiple product enabler. A commercialisation driven approach could be to investigate how the microswitch can be packaged. When the microswitch exists as a module, and off the shelf item, it can be employed as a multipurpose device in various systems.

Further work is required to optimise the ripples, which occur just before the lower cut-off frequency of the EBG structure. It may also be possible to achieve sharp cut-offs from these structures could be achieved by adding various sinus or perturbation [7 - 9].
6.3 References


Discussions and Future Work

Chapter Six


