**PERIODIC MICROSTRIP LEAKY-WAVE
ANTENNA WITH OPTICAL BEAM SCANNING
CAPABILITIES**

**BY
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in Partial Fulfillment of the Requirements
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ABSTRACT

This work investigates the feasibility of achieving dynamic beam scanning using optical control as applied to a periodic leaky-wave microstrip antenna. Optical control is implemented by illuminating periodic silicon elements placed along the length of the microstrip antenna. By varying the intensity of the optical illumination, induced conductivity changes in the silicon elements alter the antenna propagation constant and cause the peak of the radiation pattern to scan in space.

Leaky-wave behavior is first characterized by using a passive antenna design approach to investigate the effects of element size and periodic separation on radiation performance. Optical control is then integrated into the passive design using periodically placed silicon elements along the length of the microstrip structure. These elements are illuminated from below through a series of holes etched into the ground plane in order to avoid field obstruction in the front region of the antenna. These holes were found to introduce significant backside radiation, degrading antenna gain. Simulation results, for this structure, indicate that optical illumination with excess carrier concentration densities ranging up to 10^{17} cm^{-3} will produce beam scanning in the order of 5 degrees. Measured results were inconclusive. Several improvements are recommended for future work including, avoiding stop band regions, implementing an optically transparent ground plane and increasing the optical illumination intensity. Implementing these recommendations will lead to improved performance establishing the feasibility of the above antenna design.

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CHAPTER 1

INTRODUCTION

1.1 Motivation

As demand for radio frequency bandwidth continues to grow, an increasing number of wireless communications system planners are looking to millimeter wave frequency radio links for increased capacity [1]. The successful deployment of these higher frequency communication systems often require adaptive antenna systems capable of wide angle operation and dynamic beam scanning. In addition to wide angle scanning these antennas must also demonstrate high efficiency, low geometric profile, light weight construction and be suitable for low cost manufacturing [2]. These requirements are being driven by system issues such as coverage, frequency reuse, interference avoidance and cost reduction, which can become more difficult at higher frequencies.

An extensive overview of antenna technologies suitable for millimeter wave applications has been documented by Dybdal [3] and Schwering [4]. The principle architectures for these antennas include reflectors, lenses, reflectarrays, phased arrays and leaky-wave structures. The shortcomings associated with these architectures include manufacturing difficulties, complex beam steering systems, narrow bandwidths, inferior gain and large losses. Each of these configurations generally requires some form of phase shifter for beam shaping and steering. Phase shifting technologies currently available include semiconductor devices [5], ferrite structures

[5], ferroelectric techniques [6] [7] , microelectromechanical systems (MEMS) [8] [9] and optical devices [10]. The majority of these devices require complex biasing networks and several of them suffer from temperature sensitivity, large insertion loss and poor power handling capabilities. A summary of the above technologies as well as a discussion of their relative advantages and disadvantages is addressed later, however, it is sufficient to say that considerable progress in adaptive millimeter antennas is needed for next generation wireless communication systems.

1.2 Thesis Objectives

The primary objective of this research is to investigate the feasibility of optical control techniques to the design of a scanning millimeter wave microstrip periodic leaky-wave antenna. Although, previously optical control has been applied to other waveguide and microstrip antenna structures [11], [12] and [13], the application of such techniques to microstrip leaky-wave structures has not yet been investigated. The above objective was addressed through the design, simulation, fabrication and testing of a prototype antenna. Secondary objectives included the identification and analysis of design issues related to antenna operation and performance.

The microstrip antenna structure used for proof of concept purposes consisted of a transmission line periodically loaded with a series of microstrip and silicon elements. Variable optical illumination of the silicon elements provided a photo-induced conductivity gradient across the semiconductor in an attempt to cause the radiation beam peak to scan. This approach was selected since it allows the use of economic

batch microstrip fabrication techniques while simultaneously achieving the good side lobe performance, high efficiency and low profile characteristics associated with leaky-wave architectures. Exploitation of the photo-conductive properties of semiconductor materials enables ultra fast high power shifting without the use of complex expensive biasing networks. In principle this approach provides potential to achieve many of the desired features required by next-generation wireless systems.

1.3 Thesis Outline

This thesis is organized in the following manner. First a summary of current millimeter wave antenna and phase shifter technologies as well as a discussion of their relative advantages and disadvantages is provided in Chapter 2. This is followed in Chapter 3 by a detailed discussion regarding the radiation mechanisms of periodic leaky-wave antennas and semiconductor photoconductivity.

Leaky-wave antenna design which is an iterative process involving a large number of inter-dependent design parameters such as frequency of operation, phase constant, leakage rate, periodic spacing interval and dielectric constant of the substrate material is described in Chapter 4. In this section, leaky-wave behavior is first characterized using a passive antenna design approach to investigate the effects of element size and periodic separation on radiation performance. Design procedures as well as corresponding simulated results are then outlined.

Chapter 5 presents the measured performance of an optically controlled antenna prototype, which was fabricated to investigate feasibility. Measured performance is compared with simulated results and design issues are identified. Finally Chapter 6 presents a series of conclusions and addresses areas for future work.

CHAPTER 2

TECHNOLOGY OVERVIEW AND PROPOSED CONCEPT

This chapter provides a review of various antenna technologies suitable for next-generation millimeter-wave wireless systems. A brief overview of current antenna architectures and phase shifting technologies is presented followed by a discussion of the advantages and disadvantages in relation to the requirements of next-generation systems. A novel microstrip leaky-wave antenna with optical control is also introduced as a new candidate for millimeter wave systems.

2.1 Antenna Architectures

Demands for wider bandwidth and increased radio frequency spectrum are prompting wireless systems to expand into the millimeter wave bands. New low-cost adaptive wideband antenna structures are required to deal with system issues such as coverage, cell and frequency reuse as well as interference avoidance. In addition to scanning capabilities, these antennas must also exhibit many of the following characteristics:

(i) light weight, (ii) good temperature stability, (iii) high radiation efficiency, (iv) low profile, (v) broadband capabilities and (vi) capability for low-cost volume manufacturing.

Potential candidates currently being considered in this work include: (i) reflector antennas, (ii) Luneberg lenses, (iii) reflectarrays, (iv) microstrip phased arrays, (v)

leaky-wave waveguide antennas and (vi) coplanar antennas. A brief description of these various technologies along with a summary of their relative strengths and weaknesses is provided.

2.1.1 Reflector Antennas

Reflector antennas are made up of a low gain radiation source that illuminates a large conducting surface (reflector), which enhances the gain and collimates the incident radiation source. Reflector antennas come in a variety of shapes including parabolic, flat, linear and corner [14]. The position of the feed antenna and the shape of the reflector determine the location of the beam. The feed configurations can vary to include offset-fed and center-fed arrangements as illustrated in Figures 2.1a and 2.1b, respectively [14].

Figure 2.1a – *Offset-fed Reflector*

Figure 2.1b – *Center-fed Reflector*

Reflector antennas use a spatial feed network and as a result are extremely low loss structures with good gain, high radiation efficiency and superior pattern shaping capabilities. Reflectors are also capable of beam steering using multiple feeds for switch beam operation or a beam forming network. These antennas are not ideally suited to next-generation systems since they are bulky and require complex

mechanical steering and beam forming networks, particularly for wide-angle applications.

2.1.2 Luneberg Lens

The Luneberg lens is an inhomogeneous spherical lens whose index of refraction varies continuously from $\sqrt{2}$ at the center to unity at the surface [15]. The antenna consists of a feed whose phase center is placed near the surface of the lens, which then collimates the incident radiation as illustrated in Figure 2.2 [15]. Scanning is accomplished by mechanically moving the feed around the surface of the lens. Due to the symmetry of the lens, scanning performance is independent of radiation direction, unlike many other antenna structures. This makes it possible to achieve wide scan angles since the pattern does not degrade with beam location [15].

Figure 2.2 – *Luneberg Lens*

The disadvantages associated with the Luneberg lens include manufacturing difficulties for a lens with a gradual index of refraction and cumbersome mechanical beam steering.

2.1.3 Reflectarrays

Reflectarrays are a class of flat reflector antennas consisting of a source antenna and a grounded dielectric substrate with printed metal patterns as illustrated in Figure 2.3 [16]. The patterns can be created using standard printed circuit etching technology and are used to transform the spherical phase of the source antenna into a plane wave [16]. The elements in the reflectarray introduce a phase shift into the reflected wave, which is a function of the individual element's dimensions. Designing the elements to produce a unique phase shift makes it possible to achieve the desired radiation pattern [16].

Figure 2.3 – *Reflectarray Geometry*

These antennas are attractive because they have a low profile design that can be surface mounted, a wide scanning range, low fabrication costs and good efficiency with large aperture [16]. One of the drawbacks associated with reflectarrays is the relatively narrow bandwidth performance of between 5% - 10%. Also, off-axis scan performance of the reflectarray is an issue since the desired patterns can quickly degrade [16].

2.1.4 Microstrip Phased Arrays

Microstrip phased array antennas are manufactured using standard etching techniques and consist of metallic elements etched onto an electrically thin, grounded substrate [14]. The elements of the array can be fed either using a single (series), a multiple line (corporate) feed or the two configurations in combination. Series and corporate feed architectures are illustrated in Figures 2.4a and 2.4b.

Figure 2.4a – *Series Feed*

Figure 2.4b – *Corporate Feed*

The benefits of microstrip phased array antennas include a compact design, convenient manufacturing and ease of integration with active devices. The primary disadvantage of these antennas is that they require a large microstrip feed network and as a result are quite lossy at millimeter wave frequencies, which limit the maximum attainable gain [14].

2.1.5 Leaky-Wave Waveguide Antennas

Leaky-wave waveguide antennas are non-resonant and consist of a modified metal or dielectric waveguide with a series of perturbations, such as holes, slots or conductive strips. Two possible configurations are illustrated in Figures 2.5a and 2.5b. The perturbations produce radiation since they cause energy to leak away along the length of the waveguide. These antennas have a variety of designs some of which include

slotted guide, periodic dielectric guide, grooved guide, non-radiative dielectric (NRD) guide and coplanar guide [17] [18] [19] and [20].

Figure 2.5a – *Leaky-Wave Inductive Grid Antenna with Metal Perturbations* [17]

Figure 2.5b – *Leaky-Wave Slotted Guide Antenna* [17]

Scanning is achieved by changing the phase of the guided waves by loading the waveguide with either electronic or ferrite phase shifters. Loading the guiding structure changes the wave's propagation constant, which in turn causes the antenna beam to scan in space.

There are several advantages associated with these antennas, such as low side lobe levels, good efficiency, low profile and ease of integration with a variety of phase shifter technologies [17]. Leaky-wave antennas are essentially modified waveguide structures and depending on the guiding medium they can be quite difficult to

fabricate. These antennas also suffer from a limited operational bandwidth since their beam peak scans with frequency.

2.1.6 Coplanar Leaky-Wave Line Antennas

Recently a linear co-planar leaky-wave antenna has been proposed for millimeter wave applications [19]. The antenna is illustrated below in Figure 2.6 and its radiation is caused by periodic perturbations along the length of a coplanar waveguide. The advantages of this antenna design include reduced conductor loss, capacity to incorporate shunt devices without the need for drilling via holes and straightforward fabrication using standard etching procedures. However, due to their geometry coplanar antennas intrinsically excite spurious modes and as a result often require complex designs to provide sufficient radiation efficiency.

Figure 2.6 – *Coplanar Antenna Configuration and Dimensions* [19]

2.2 Phase Shifting Devices

Beam steering requires the relative excitation phase between the elements of an array to be dynamically changed typically using phase shifters. There are a variety of phase shifting devices, which can be incorporated to provide beam scanning including, semiconductors, ferrites, ferroelectrics, microelectromechanical systems

(MEMS) and optical phase shifters. A brief survey of the most common phase shifter technology is provided in the subsequent sections.

2.2.1 Semiconductor Phase Shifters

Semiconductor phase shifters consist of semiconductor junction devices such as diodes or transistors, and circuitry to electronically control the signal using either digital or analog techniques [5]. P-i-n diodes and GaAs FETs are two-terminal voltage controlled off-on switches, which are frequently incorporated into a variety of phase shifting circuit topologies such as switched lines, hybrid couplers and loaded lines designs as illustrated in Figure 2.7a, 2.7b and 2.7c respectively [5].

Figure 2.7a – *Switched Line p-i-n Diode Phase Shifter* [5]

Figure 2.7b - *Hybrid Coupled p-i-n Diode Phase Shifter* [5]

Figure 2.7c – Loaded Line *p-i-n* Diode Phase Shifter [5]

The switched line circuit obtains a phase shift by using a semiconductor device to switch the signal between different path lengths, thus realizing a shift that corresponds to the additional path delay. The hybrid coupler phase shifter employs two identical reflective terminations, each with two switched states, at the coupled ports of the hybrid device. In this device, the phase shift corresponds to the additional path length traversed by the signal in the two switched states. The loaded line phase shifter employs diodes to switch between shunt capacitive and shunt inductive loads. This switch has the same effect as increasing and decreasing the electrical length of the transmission line and causes a phase shift.

The benefits of using these types of phase shifters include good temperature stability, easy integration with most antenna technologies and the capacity to be incorporated in both digital and analog designs. These devices however are generally not designed to for high power handling and are expensive [5].

2.2.2 Ferrite Phase Shifters

A simple ferrite phase shifter consists of a waveguide, a drive coil and a ferrite rod as illustrated in Figure 2.8 [5]. These shifters rely on the interaction between

electromagnetic waves and the spinning electrons in a magnetic ferrite, which can be controlled by an applied magnetic field to induce a phase variation [5]. The permeability of the ferrite material is directly proportional to the magnitude of the applied magnetic bias field. The propagation constant of the electromagnetic waves traveling within the waveguide is a function of the permeability of the medium within the guide; therefore, a change in permeability will result in a change in the propagation constant [5]. It is possible to electronically control the phase of the electromagnetic energy by altering the magnitude of the applied magnetic field.

Figure 2.8 – Ferrite Rod Phase Shifter [5]

The advantages associated with these phase shifters include quick switching times and good power handling capabilities [5]. However, the disadvantages relate to temperature sensitivity and incompatibility with most microstrip and planar devices [5].

2.2.3 Ferroelectric Phase Shifters

Ferroelectric materials experience a change in relative permittivity when a dc bias voltage is applied [21]. Ferroelectric phase shifters exploit this property to alter the

effective electrical length of transmission lines and thus create a delay line phase shifter as illustrated in Figure 2.9 [22].

Figure 2.9 – *Thin Film Ferroelectric Coplanar Waveguide Phase Shifter* [22]

Currently, there are two types of ferroelectric phase shifters, thick and thin film [21]. The thickness of thick film phase shifters measures in the order of 0.5 mm to 2.5 mm, whereas thin film devices are in the order of 0.3 μm to 2 μm [21]. Thick film devices exhibit low loss, however they require large dc bias voltages. Thin film devices, on the other hand, experience large loss but require a lower dc bias voltage on the order of 10^4 V/cm [8]. There are several advantages to these phase shifters such low cost, good power-handling capabilities and compatibility with most planar technologies [8]. The disadvantages associated with these devices include high insertion loss and large dc biasing voltage requirements [21].

2.2.4 Microelectromechanical Systems (MEMS) Phase Shifters

Microelectromechanical systems are devices that use electrostatic, piezoelectric, magnetostatic or thermal material properties act as a mechanical switch in phase shifting circuitry [9]. These switching devices can be fabricated using either a fixed membrane or a floating cantilever and are modeled as mechanical springs with an

equivalent spring constant [9]. The switches can be integrated in either a series or shunt configuration to produce a variety of phase shifter designs. Figures 2.10a and 2.10b depict series and shunt MEMS switch, respectively.

Figure 2.10a – *Cross Sectional View of a Series MEMS* [9]

Figure 2.10b – *Cross Sectional View of a Shunt MEMS* [9]

The advantages associated with these phase shifters include near zero power consumption, high isolation and low insertion loss [9]. The disadvantages include high voltage drives, and long term reliability and packaging issues [9].

2.2.5 Optical Phase Shifters

Optical phase shifters consist of a rectangular semiconductor waveguide with tapered ends as illustrated in Figure 2.11 [10]. Optical control is realized by illuminating a region of the broadband wall of the semiconductor guide to form an area of plasma. The plasma effectively loads the guide by introducing a region of high conductivity, which changes the propagation constant of the guided wave [10]. The magnitude of the phase shift varies with the intensity of optical illumination, by altering the

propagation constant, making it possible to achieve a range of phase shifts on the order of $300^\circ/\text{cm}$ with a single device [23].

Figure 2.11 – Optically Controlled Phase Shifter [10]

The benefits of this phase shifter include ultra fast switching, good power handling capability and no biasing requirements. However, these devices closely resemble antenna structures, making them prone to spurious radiation, which can result in relatively high losses.

2.3 Microstrip Leaky-Wave Antenna with Optical Beam

Scanning

The previous sections discussed a variety of antenna and phase shifting technologies suitable for application in millimeter-wave wireless communications systems. All of these alternatives exhibit both advantages and disadvantages a summary of which is provided in Tables 2.1 and 2.2, with respect to a list of next generation system requirements.

Antenna Architecture	High Volume Manufacturing	Low Profile	Low Cost	Efficient Radiator	Wide Angle Scanning
Reflector	X	X	X	✓	X
Luneberg Lens	X	X	X	✓	✓
Reflectarray	✓	X	✓	✓	X
Microstrip Phased Array	✓	✓	✓	X	X
Leaky-Wave Waveguide	X	✓	X	✓	✓
Coplanar Line	X	✓	X	✓	✓

Table 2.1 – Summary of Millimeter Wave Antenna Architectures and their Respective Capabilities

Phase Shifting Device	Fast Switching Time	High Power	Temperature Sensitivity	Low Loss	Long Term Reliability
Semiconductor	✓	X	✓	✓	✓
Ferrite	✓	✓	X	✓	✓
Ferroelectric	✓	✓	✓	X	✓
MEMS	✓	✓	X	✓	X
Optical	✓	✓	✓	✓	✓

Table 2.2 - Summary of Phase Shifting Devices and their Respective Capabilities

Examining Table 2.1 it can be seen that all of the antenna architectures suffer from at least one shortcoming. An interesting combination that merits further investigation involves the integration of microstrip and leaky-wave designs. Microstrip antennas suffer from poor radiation efficiency and large losses. Conversely, leaky-wave antennas are efficient low loss structures. A microstrip leaky-wave antenna can be realized by replacing the waveguide with a microstrip transmission line. This approach exploits the simple and economic fabrications techniques of microstrip technology while simultaneously achieving the beam scanning advantages of a leaky-wave design.

Table 2.2 emphasizes the multiple benefits associated with optical phase shifting devices. Incorporating optical phase shifting techniques with a microstrip leaky-wave architecture permits low cost, fast switching and wideband beam scanning without the use of complex biasing networks.

Although, the application of optical control to antennas is a relatively new concept, it has, in the past, been integrated into various microwave devices such as a solid state microstrip switches [23] and phase shifters [24]. More recently, several antenna architectures have employed optical control techniques, including among others, a millimeter wave dielectric waveguide antenna [11], a semiconductor slab waveguide antenna with periodic photo-excitation [12] and most recently a reconfigurable bowtie antenna [13]. The above mentioned dielectric and semiconductor slab antennas both employ optical illumination to induce a series of photoconductive regions along the length of a semiconductor waveguide to alter the structure's phase constant. These waveguide antennas, however, are bulky and cannot be manufactured using low cost batch techniques. The bowtie antenna uses optical illumination to create a reconfigurable metalized antenna pattern on a semiconductor substrate material, however, is not intended for beam scanning operation. The proposed microstrip leaky-wave antenna incorporates both of these concepts to implement optical beam scanning using microstrip technology.

Verification and assessment of the practical feasibility of this novel millimeter wave antenna design is pursued as part of this research activity. The proposed microstrip

leaky-wave antenna structure consists of a transparent fused silica substrate, gold ground and conductor planes and pieces of high resistivity silicon as illustrated in Figure 2.12. The microstrip structure is perturbed with pieces of high resistivity silicon and then illuminated creating local plasma regions. The optical intensity is varied causing the conductivity of the silicon elements to change, which in turn change the structure's propagation constant causing the beam peak to scan in space. The semiconductor is illuminated from the underside by means of small holes etched in the ground plane to prevent the optical source from blocking the forward radiation pattern. The design and performance of this novel antenna architecture is further discussed in Chapters 3, 4 and 5.

Figure 2.12 – *Novel Microstrip Leaky-Wave Antenna with Optical Control*

CHAPTER 3

LEAKY-WAVE ANTENNA AND PHOTOCONDUCTIVITY THEORY

The purpose of this chapter is to provide an overview of leaky-wave antenna theory and the photoconductive properties of semiconductors as applied to the design and analysis of the proposed millimeter wave microstrip antenna introduced in Chapter 2.

3.1 Leaky-Wave Antenna Theory

Antennas can be classified as either resonant or non-resonant. Resonant antennas, such as microstrip patches and dipoles, resonate at discrete frequencies and radiate strongly. Their radiation efficiency, however, rapidly degrades with frequency producing a relatively narrow band design. Non-resonant antennas or traveling wave antennas do not resonate but rather leak energy along their length, thereby producing more efficient broadband structures. A leaky wave antenna is a type of traveling-wave structure that radiates continuously along its length [25]. Generally, leaky-wave structures are designed to be approximately 20 wavelengths in length and to radiate 90 % of the input power [15].

Depending upon their architecture, leaky-wave antennas can be categorized as either uniform or periodic. Uniform antennas are homogeneous along the length of the guiding structure. An example of a uniform leaky-wave antenna is a slitted circular waveguide whose basic geometry is illustrated in Figure 3.1 [25].

Figure 3.1 – *Example of a Uniform Leaky-Wave Antenna, a Slitted Circular Waveguide [25]*

Periodic leaky-wave antennas have regular perturbations along their length [15]. These antennas can be designed to radiate from backward end fire into the forward quadrant with the exception of a narrow region around broadside. A periodic holey waveguide is a periodic leaky-wave radiator whose geometry is illustrated below in Figure 3.2 [25].

Figure 3.2 – *Example of a Periodic Leaky-Wave Antenna, a Holey Waveguide [25]*

This section will deal only with leaky-wave theory as it applies to periodic architectures since the periodic antennas have superior scan performance [15].

3.1.1 Radiation Behavior of Leaky-Wave Antennas

The radiation behavior of periodic leaky-wave antennas can be characterized using a complex wave number of the perturbed guiding structure as follows [15]:

$$\kappa = \beta + j\alpha \quad (\text{m}^{-1}) \quad (3.1)$$

where, β and α are the phase and leakage constants, respectively. The leakage constant is similar to the attenuation constant of a transmission line, however this case accounts for the loss due to radiation. The leakage constant is related to the rate at which energy is radiated from the antenna.

The phase constant describes wave propagation in the guiding medium and can be expressed as [25]:

$$\beta = \frac{2\pi}{\lambda_g} \quad (\text{rads/m}) \quad (3.2)$$

where λ_g is the guided wavelength.

The relationship between the free-space wave number k_o , which is equal to $2\pi/\lambda_o$, and the phase constant β can be graphically illustrated using k_o vs. β diagram as illustrated in Figure 3.3 [25] for the dielectricly loaded uniform quasi-TEM line depicted in Figure 3.4.

Figure 3.3 – k_0 vs. β Diagram of a Dielectric Loaded Uniform Quasi-TEM Line [25]

Figure 3.4 – Dielectric Loaded Uniform Quasi-TEM Line

The two solid straight lines in Figure 3.3, represent the forward and backward traveling waves, in guiding structure. The slopes of these lines are inversely proportional to the relative permittivity of the guiding material and can be expressed as follows [25]:

$$\frac{k_o}{\beta} = \sqrt{\frac{\epsilon_o}{\epsilon_{eff}\epsilon_o}} = \sqrt{\frac{1}{\epsilon_{eff}}} \quad (3.3)$$

$$-\frac{k_o}{\beta} = -\sqrt{\frac{\epsilon_o}{\epsilon_{eff}\epsilon_o}} = -\sqrt{\frac{1}{\epsilon_{eff}}} \quad (3.4)$$

where ϵ_o is the permittivity of free space and ϵ_{eff} and is the effective permittivity of the dielectric guiding material.

The dashed lines represent forward and backward traveling waves with a phase velocity equal to the speed of light. A wave with a phase velocity, $v_p = \omega/\beta$, that is greater than the speed of light is considered to be fast ie. $|\beta| < k_o$; whereas, a wave with a phase velocity slower than the speed of light is said to be slow ie. $|\beta| > k_o$. In all traveling wave antenna structures slow waves are bound and do not radiate while fast waves are unbound and radiate [15]. Fast and slow wave regions can be graphically described on a k_o vs. β as shown in Figure 3.5 [25].

Figure 3.5 – Fast and Slow Wave Regions [25]

3.1.2 Periodic Leaky-Wave Antennas

Uniform leaky-wave antennas such as the slit cylinder illustrated in Figure 3.1 support fast wave modes and inherently radiate. Periodic antennas, on the other hand, are based on slow-wave guiding structures and rely on periodic perturbations to excite fast-wave modes. These periodic perturbations introduce an infinite but discrete number of spatial harmonics each of which can be characterized by the following unique phase constants [15]:

$$\beta_n = \beta_o + \frac{2n\pi}{d} \quad (\text{rad/m}) \quad (3.5)$$

where β_o is the fundamental space harmonic, d is the period of a modulating structure and n is an integer representing the space harmonic number.

A simplified dispersion relation for each of these harmonics are a function of frequency and the spacing d and can be graphically represented using a Brillouin Diagram as illustrated below in Figure 3.6 [25] for the axial periodically loaded quasi-TEM structure depicted in Figure 3.7.

Figure 3.6 – Brillouin Diagram of an Axial Periodically Loaded Quasi-TEM Structure [25]

Figure 3.7 – Axial Periodically Loaded Quasi-TEM Structure

The Brillouin diagram is periodic over intervals of 2π and contains a series of straight lines with slopes of $\pm(1/\epsilon_{eff})^{0.5}$ starting at $\beta d = 2n\pi$ (where $n = 0, \pm 1, \pm 2 \dots$) [25].

Inspection of the Brillouin diagram reveals that traveling wave space harmonic with can now enter the fast wave region and will therefore radiate [15]. Brillouin diagrams are used to theoretically determine the fast-wave region for antenna design. The frequency and spacing are often chosen such that only the $n = \pm 1$ space harmonics radiate producing single radiation lobe at the desired operating point [15].

A more accurate representation of the above Brillouin diagram consists of a series of pass bands and stop bands due to coupling between forward and backward wave modes as illustrated below in Figure 3.8 [25]. In the stop band regions the mode is attenuated along the guiding structure resulting in poor radiation.

Figure 3.8 – Modified Brillouin Diagram [25]

The angle of radiation can be analytically predicted to assess the antenna's theoretical beam scanning performance. The peak of the radiation lobe is measured from broadside and is directly related to the frequency of operation [25].

Within the fast wave region, the direction of the beam peak of a leaky-wave antenna can be graphically determined for any arbitrary operating point P using the following procedure with the aid of the illustration shown in Figure 3.9: (i) draw a circle centered at the origin O with a radius $k_e d$ and (ii) drop a perpendicular which intersects the circle (at point C) from point P to the $\beta_n d$ axis [25]. The angle between the OC segment and the $k_o d$ axis is related to the radiation angle of the beam with respect to broadside through following equation [25].

$$\theta_n = \sin^{-1} \left(\frac{\beta_n}{k_o} \right) \text{ (Degrees)} \quad (3.6)$$

where θ_n is the angle of radiation and β_n is the phase constant of the n^{th} radiating spatial harmonic.

Figure 3.9 – Graphical Analysis of Beam Peak Direction [25]

3.2 Photoconductive Semiconductor Theory

Semiconductors are a group of materials whose conductivity can be altered among other means, through optical excitation [26]. Optical excitation occurs when the incident photon energy is sufficient to excite a valence band electron into the

conduction band. Photon energy is directly related to optical wavelength and in general semiconductors are most sensitive to photons with energies equal or slightly more energetic than the band gap [26]. Less energetic photons are not absorbed and very high energy photons are only absorbed at the surface and contribute negligibly to bulk conductivity. The relative spectral response of a slab of silicon uniformly illuminated over an optical spectrum ranging from 0.3 μm to 1.2 μm is illustrated in Figure 3.10 [24]

Figure 3.10 – *Relative Spectral Response of Silicon* [24]

3.2.1 Optically Induced Conductivity

Photons absorbed in a uniform piece of semiconductor material increase the free charged carrier concentration creating a plasma layer and raising the surface

conductivity. The relationship between conductivity and carrier concentration is described by [26]:

$$\sigma = q(z\mu_n + p\mu_p) \text{ (S/m)} \quad (3.9)$$

where σ is conductivity, q is the magnitude of the charge, z is the electron concentration in the conduction band, p is the hole concentration in the valence band and μ_n and μ_p are the electron mobility and hole mobility respectively.

The change in conductivity due to optical excitation can be quantified by examining excess carrier concentrations. In equilibrium the only source of carrier generation is thermal excitation and the equilibrium generation rate can be expressed as [26]:

$$gT = \alpha_r z_o p_o \text{ (cm}^{-3}\text{)} \quad (3.10)$$

where gT is the equilibrium generation rate, α_r is a constant of proportionality for recombination, z_o is the equilibrium concentration of electrons and p_o is the equilibrium concentration of holes.

Optical excitation increases the excess generation rate by a factor of g_{op} enabling equation (3.10) to be rewritten as [26]:

$$gT + g_{op} = \alpha_r (z_o + \delta z)(p_o + \delta p) \text{ (cm}^{-3}\text{)} \quad (3.11)$$

where g_{op} is the optical generation rate in cm^{-3} , δz and δp are the changes in electron and hole concentration, respectively from thermal equilibrium.

Assuming steady state recombination and no trapping due to impurities or defects [20], the changes in electron and hole concentrations are equivalent ($\delta z = \delta p$).

Therefore, (3.11) can be expanded and rewritten as follows [26]:

$$gT + g_{op} = \alpha_r z_o p_o + \alpha_r [(z_o + p_o)\delta z + \delta z^2] \text{ (cm}^{-3}\text{)} \quad (3.12)$$

Examining (3.12) it can be seen that the first term is equal to the equilibrium generation rate gT . Neglecting second order effects due to low-level excitation the optical generation rate can be rewritten as follows [26]:

$$g_{op} = \alpha_r (z_o + p_o) \delta z = \frac{\delta z}{\tau_z} \text{ (cm}^{-3}\text{)} \quad (3.13)$$

where τ_z is the carrier lifetime.

A general expression describing the optically induced conductivity can be obtained by substituting (3.13) into (3.9) and is expressed below as [26]:

$$\Delta\sigma = q(\delta z \mu_z + \delta p \mu_p) = q g_{op} \tau_z (\mu_z + \mu_p) \text{ (S/m)} \quad (3.14)$$

3.2.2 Complex Permittivity

The behavior of electric fields within the plasma region of a semiconductor can be characterized by a complex permittivity thus allowing modeling and simulation of antenna performance. The complex permittivity can be derived from first principles using Maxwell's equations as follows [27]:

$$\nabla \times \vec{H} = \sigma \vec{E} + \frac{\partial \vec{D}}{\partial t} \quad (3.15)$$

where $\vec{H}$ is the magnetic field intensity vector, $\vec{E}$ is the applied time-dependant electric field vector and $\vec{D}$ is the electric displacement vector.

The displacement vector can be expressed in terms of the applied electric field vector [28]:

$$\vec{D} = \epsilon_r \epsilon_o \vec{E} \quad (\text{Coul/m}^2) \quad (3.16)$$

where ϵ_r is relative permittivity and ϵ_o is the permittivity of free space.

In a dielectric, an applied electric field causes the polarization of atoms or molecules of the material to create electric dipole moments, which create a total displacement flux [27]. The additional polarization vector can be added to the displacement vector and written [27]:

$$\vec{D} = \epsilon_o \vec{E} + \vec{P} \quad (\text{Coul/m}^2) \quad (3.17)$$

where $\vec{P}$ is the electric polarization vector.

In a linear medium, the electric polarization vector is linearly related to the applied electric field vector through the following equality [27]:

$$\vec{P} = \epsilon_o \chi_e \vec{E} \quad (\text{Coul/m}^2) \quad (3.18)$$

where χ_e is the electrical susceptibility and is defined as [29]:

$$\chi_e = \chi_f + \chi_b \quad (3.19)$$

where χ_f and χ_b are the susceptibilities of the free and bound charges respectively.

Substitution of (3.18) into (3.17) introduces the following equality [29]:

$$\vec{D} = \epsilon_o \vec{E} (1 + \chi_b + \chi_f) \quad (\text{Coul/m}^2) \quad (3.20)$$

Equating (3.20) to (3.16) results in the following relationship for the relative permittivity [29]:

$$\epsilon_r = 1 + (\chi_b + \chi_f) = \epsilon_L + \chi_f \quad (3.21)$$

where ϵ_L is the dielectric constant of the host lattice and is given by [29]:

$$\epsilon_L = 1 + \chi_b \quad (3.22)$$

The motion of charged carriers within the plasma under the action of an applied electric field can be described as [30]:

$$m^* \left(\frac{d^2}{dt^2} + \frac{1}{\tau} \frac{d}{dt} \right) \delta \vec{r} = -q \vec{E} \quad (\text{N}) \quad (3.23)$$

where m^* is the effective mass of the charged carrier, $\delta \vec{r}$ is the spatial displacement vector, q is the carrier charge and τ is the collision time.

The collision time is a function of the effective mass of the charged carriers and is expressed as [28]:

$$\tau = \frac{\mu}{m^* q} \quad (\text{s}) \quad (3.24)$$

where μ is the mobility of the charged carriers.

The polarization vector due to free charges is described by the following equation [29]:

$$\vec{P}_f = -zq\delta \vec{r} \vec{E} \text{ (Coul/m}^2\text{)} \quad (3.25)$$

where z is the carrier density in the conductive region.

The polarization of the free carriers can be further described by substituting (3.23) into (3.25) to produce [29]:

$$\vec{P}_f = \frac{zq^2}{m^*} \left(\frac{d^2}{dt^2} + \frac{1}{\tau} \frac{d}{dt} \right)^{-1} \vec{E} \text{ (Coul/m}^2\text{)} \quad (3.26)$$

Performing a Fourier Transform to convert to the frequency domain, equation (3.26) can be rewritten as [29]:

$$\vec{P}_f = \frac{zq^2}{m^*} \left(-\omega^2 + j \frac{\omega}{\tau} \right)^{-1} \vec{E} \text{ (Coul/m}^2\text{)} \quad (3.27)$$

where ω is equal to $2\pi f$ and f is the frequency of the electromagnetic waves.

Finally, an expression for the complex permittivity in the semiconductor plasma region can be obtain by equating (3.27) to the free charge component of (3.18) and using (3.22) results in the following expression [29]:

$$\varepsilon_r = \varepsilon_L - \frac{zq^2}{\varepsilon_o m^*} \left(\frac{\tau^2}{1 + \omega^2 \tau^2} \right) - j \frac{zq^2}{\varepsilon_o m^* \omega} \left(\frac{\tau}{1 + \omega^2 \tau^2} \right) \quad (3.28)$$

The general expression for complex permittivity is given as follows [27]:

$$\varepsilon_r = \varepsilon' - j\varepsilon'' \quad (3.29)$$

where ε' and ε'' refer to the real and complex components of the permittivity, respectively.

Equation 3.28 is analogous to 3.29 and defines the real and imaginary components in terms of excess carrier concentration, plasma frequency and carrier lifetime.

3.2.3 Plasma Penetration Depth

The plasma region of the semiconductor has a finite penetration depth. This depth is required to accurately simulate the antenna performance and has been linearly approximated by Platt [29] as follows:

$$d_e = \frac{1}{\psi_r} \left[\frac{L_a(1 + \psi_r L_a) + v_s \tau}{L_a + v_s \tau} \right] \left[\frac{1}{\psi_r L_a} \left(\frac{\psi_r L_a^2 + v_s \tau}{L_a + v_s \tau} \right) \right]^{\frac{\psi_r L_a}{1 - \psi_r L_a}} \quad (\text{m}) \quad (3.30)$$

where L_a is the diffusion length defined as,

$$L_a = \left[\left(\frac{2\mu_z \mu_p \tau T k_b}{q} \right) (\mu_p + \mu_z)^{-1} \right]^{\frac{1}{2}} \quad (\text{m}) \quad (3.31)$$

where ψ_r is the radiation absorption coefficient, v_s is the surface recombination velocity, τ is the excess carrier lifetime, T is the absolute temperature, k_b is Boltzmann's constant, q is the electronic charge and μ_z and μ_p are the electron and hole mobilities respectively.

CHAPTER 4

LEAKY-WAVE ANTENNA DESIGN

This chapter provides a detailed design and analysis of leaky-wave antenna behavior. An initial assessment of critical design parameters is carried out for an optically controlled microstrip leaky-wave antenna which meets the following criteria: (i) low cost planar substrate, (ii) operating frequency of 32 GHz and (iii) gain of 10 dBi. Performance is assessed based upon theoretical considerations and design feasibility. From this assessment, a practical architecture for the antenna is derived. A detailed design for a passive antenna without optical control is then undertaken to establish basic antenna geometry. Fabrication constraints are also taken into account. Optical control design elements are subsequently integrated into the passive design for final fabrication and testing. Simulations are performed throughout the design process.

4.1 Concept Feasibility Assessment

The initial design concept involves a microstrip periodic leaky-wave antenna printed on a silicon substrate material. Beam scanning is achieved by selectively illuminating the semiconductor with variable optical intensities as illustrated in Figure 4.1. The intent is to vary the electrical properties of the leaky-wave antenna's perturbations to produce beam scanning. The feasibility of this initial design concept is assessed in the following areas: (i) the effect of substrate material on insertion loss, (ii) the effect of dielectric material on radiation performance, (iii) the impact of the dielectric

material on the characteristics of the fast-wave region and (iv) the overall practicality of implementing the proposed architecture.

Figure 4.1 – Initial Antenna Concept and Illumination Technique

4.1.1 Substrate Insertion Loss

The effect of substrate material on insertion loss was assessed. For the purpose of this assessment, dielectric and conductor losses are calculated for 50 ohm microstrip lines on both high and low dielectric constant substrate materials. Silicon and fused silica substrates with a thickness of 0.5 mm were selected as the high and low dielectric constant materials respectively. The silicon has a resistivity larger than $400 \Omega/\text{cm}$, a dielectric constant of 11.7 and a loss tangent of 0.0012, while fused silica has a dielectric constant of 4.0 and a loss tangent of 0.003. The transmission lines are assumed to be operating at 32 GHz and the corresponding line widths for silicon and fused silica are 0.4 mm and 1.02 mm respectively. Microstrip width and loss calculations are performed using Matlab software [31] with the simulation code located in sections A.1 and A.2 of the Appendix.

Dielectric loss for both the silicon and fused silica substrates are calculated using the following equation [32]:

$$\alpha_d = 27.3 \frac{\epsilon_r}{\epsilon_r - 1} \left(\frac{\epsilon_{eff} - 1}{\sqrt{\epsilon_{eff}}} \right) \frac{\tan \delta}{\lambda_o} \quad (\text{dB/m}) \quad (4.1)$$

where ϵ_r is the relative permittivity of the substrate material, ϵ_{eff} is the effective permittivity, $\tan \delta$ is the loss tangent and λ_o is the free space wavelength of operation.

The silicon and fused silica losses were found to be 0.09 dB/cm and 0.13 dB/cm, respectively. Although the fused silica dielectric losses are 0.04 dB/cm larger than those of the silicon the difference is not significant.

Conductor loss for the silicon substrate transmission line can be calculated using the following equation [32]:

$$\alpha_c = \frac{8.68 R_s}{2\pi Z_o h} \left[1 - \left(\frac{w'}{4h} \right)^2 \right] \left[1 + \frac{h}{w'} + \frac{h}{\pi w'} \left(\ln \frac{4\pi w}{t} + \frac{t}{w} \right) \right] \quad (\text{dB/m}) \quad (4.2)$$

where R_s is the surface resistivity of the conductor, Z_o is the impedance of the transmission line, h is the thickness of the substrate material, w' is the effective width of the transmission line, w is the width of the transmission line and t is the thickness copper of the transmission line chosen to be 1 μm . Equation 4.2 is only valid for the structures with $w/h \leq 1/2\pi$.

Since the transmission line on the fused silica substrate has a larger width w the following expression is used to determine it's conductor loss [32]:

$$\alpha_c = \frac{8.68R_s}{2\pi Z_o h} \left[1 - \left(\frac{w'}{4h} \right)^2 \right] \left[1 + \frac{h}{w'} + \frac{h}{\pi w'} \left(\ln \frac{2h}{t} - \frac{t}{w} \right) \right] (\text{dB/m}) \quad (4.3)$$

Equation 4.3 is only valid for structures with w/h ratios in the following range, $1/2\pi < w/h \leq 2$. For the particular transmission lines in question, the silicon and fused silica conductor losses are determined to be 0.12 dB/cm and 0.06 dB/cm, respectively. The results again reveal only a small difference of 0.06 dB/cm between the conductor loss of silicon and those of fused silica.

Assuming a transmission line length of 10 cm, the overall insertions loss for silicon and fused silica substrates will be 2.1 dB and 1.9 dB respectively. Since both fused silica and silicon do not incur significant losses both materials appear to be feasible substrate candidates.

4.1.2 Substrate Material Radiation Characteristics

The electric field intensity within a dielectric material will vary according to the material's relative permittivity or dielectric constant as given in (4.4) [33]:

$$\vec{D} = \epsilon_o \epsilon_r \vec{E} \quad (\text{Coul/m}^2) \quad (4.4)$$

where, D is the electric flux density vector, ϵ_o is equal to 8.854×10^{-12} F/m, ϵ_r is the relative permittivity and E is the electric field intensity vector.

The electric fields of a microstrip antenna constructed on a high dielectric constant material are highly concentrated within the substrate. Conversely, the electric fields of a similar antenna constructed on a low dielectric constant substrate are less

confined to the substrate facilitating radiation. The high dielectric constant of silicon is therefore less conducive to radiation and is consequently a poorer substrate material. On the other hand, fused silica, with its lower permittivity would appear to be better suited for such applications.

4.1.3 Effect of Dielectric Constant on Fast-Wave Region

The dielectric constant of the substrate material will affect a variety of the periodic leaky-wave antenna design parameters including periodic spacing, frequency of operation, phase constant and the size of the fast wave region. As previously discussed in Chapter 2, the slopes of the space harmonics are inversely related to the effective dielectric constant of the substrate material.

The effect of dielectric constant on periodic antenna design can be approximated by creating Brillouin diagrams from a 50 ohm transmission line on both silicon and fused silica. Transmission line dimensions and operating parameters are identical to those previously outlined in Section 4.1.1. These Brillouin diagrams are constructed by assuming the periodic perturbations do not alter the propagation constant of the line. Matlab [31] software was used to generate the Brillouin diagram in Figure 4.2 for the $n = \pm 1$ space harmonic, with the code provided in section A.3 of the Appendix.

Figure 4.2 – Brillouin Diagram with Silicon and Fused Silica $n = \pm 1$ Space Harmonic

Upon inspection of Figure 4.2, it is apparent, that the fused silica substrate has a larger fast wave region resulting in greater flexibility when selecting the periodic spacing between elements and the structure's phase constant. Silicon suffers from a more limited fast-wave range making it an inferior substrate material for periodic leaky-wave antennas.

4.1.4 Implementation Practicality

In order to be considered feasible, the preliminary antenna design must be efficient and easy to implement. As previously mentioned, the preliminary architecture involves a microstrip leaky-wave periodic antenna on a silicon substrate material. Microstrip designs can be easily manufactured on a silicon substrate material using standard etching techniques. Beam scanning is achieved by illuminating selective

regions of semiconductor material and then varying optical intensity. The power supplied to the optical source can be easily adjusted to vary light intensity. Selective illumination can also be accomplished by placing a mask over the light source permitting light to strike only designated regions of the semiconductor.

As discussed in Chapter 3, the induced plasma has a thin finite depth. The semiconductor ideally should be illuminated from above to ensure electrical continuity between the plasma and copper trace as depicted in Figure 4.3.

Figure 4.3 – Top View of Antenna with Induced Plasma

This illumination technique would potentially physically obstruct the antenna's radiation pattern, compromising gain. An alternate approach involves illumination from below through small holes created in the ground plane. This method would require an optically transparent substrate material coupled with a semiconductor superstrate.

4.2 Revised Optical Antenna Architecture

Based upon the preceding feasibility assessment a revised antenna architecture has been derived to overcome some of the identified shortcomings in the initial design concept of Figure 4.1. The revised design concept now involves the use of a transparent fused silica substrate for improved radiation performance. Silicon elements are placed periodically along the length of the antenna and illuminated from below through the transparent fused silica and a series of small holes in the ground plane. The silicon elements must have a high resistivity to minimize loss in the dark state. Since it is not possible to sputter high resistivity silicon, the elements can be constructed by dicing a wafer into small pieces. The proposed structure and optical illumination is illustrated in Figure 4.4.

Figure 4.4 – Revised Optical Antenna Architecture

In addition to improved radiation performance, the fused silica will significantly increase the fast-wave region. Illuminating from below will not obstruct the antenna's main radiation lobe, thus improving overall gain.

Although the revised design offers numerous improvements, there is one significant drawback. The holes required in the ground plane for optical illumination will disrupt the current distribution resulting in some backside radiation, which in turn will degrade antenna gain. Despite this shortcoming, the revised design provides a number of benefits and therefore was selected as a basis for further design and implementation.

4.3 Passive Antenna Design

A detailed design of a passive microstrip leaky-wave antenna without optical control is performed to establish basic antenna parameters. The passive structure was designed on a 0.5 mm thick fused silica substrate and consists of a copper trace made up of a 50 ohm microstrip transmission line with copper elements placed periodically along its length, as illustrated in Figure 4.5.

Figure 4.5 – *Top View of Passive Antenna Structure*

The following parameter definitions are applied to the design: (i) L refers to the antenna length, (ii) i refers to the element length, (iii) d refers to the periodic separation between elements and (iv) t refers to the element width.

Leaky-wave antenna design is an iterative process, therefore, prior to commencing the design it is necessary to initially fix a number of the following antenna parameters: (i) input and output impedances of the antenna are set to 50 ohms to ensure an impedance match with all standard test equipment, (ii) periodic elements were chosen to have a 50 ohm width of $t = 1.02$ mm for the following reasons: (a) higher impedance transmission lines are narrow thus more difficult to fabricated and (b) lower impedance lines are thick limiting periodic spacing range and design flexibility, (iii) ten elements are placed along the length of a $L = 100$ mm long antenna to ensure that a high percentage of the input power is radiated and (iv) the element spacing is set at $d = 9.5$ mm for antenna operation in the fast wave region. A number of these parameters are optimized throughout the course of the design as discussed in subsequent sections.

4.3.1 Element Size

As previously outlined in Chapter 3, periodic leaky-wave antennas radiate due to regularly spaced modulating elements along the structure's length. Typically, the elements should be designed such that less than 10 percent of the input power remains at the end of the antenna. The size of the elements affects the overall complex propagation constant of the antenna, where the real and imaginary parts determine the location of the beam peak and the rate of radiation, respectively. The effect of element length on the rate of radiation is examined using scattering parameter analysis.

The S_{11} and S_{21} parameters are used to quantify the amount power reflected back towards the source and the amount arriving at the output terminal, respectively.

Energy propagating along the antenna is lost due to conductor, dielectric and radiation losses. The radiation effects of the elements are isolated from the transmission line losses by normalizing the antenna S_{21} values with those of a 50 ohm transmission line of equal length.

Scattering parameters simulations were performed using Ensemble software [34] for different elements lengths: $i = \lambda_g/2 = 2.6$ mm, $i = \lambda_g/6 = 0.86$ mm, $i = \lambda_g/4 = 1.3$ mm and $i = \lambda_g/8 = 0.65$ mm. A 100 mm long 50 ohm transmission line was simulated for the baseline comparison. All of these structures were designed for an operating frequency of 32 GHz. The S_{11} and normalized S_{21} results are graphically displayed in Figures 4.6 and 4.7, respectively.

Figure 4.6 – S_{11} Simulated Results for $i = 2.6$ mm, $i = 1.3$ mm, $i = 0.83$ mm and $i = 0.65$ mm

Figure 4.7 – Normalized S_{21} Simulated Results for $i = 2.6$ mm, $i = 1.3$ mm, $i = 0.83$ mm and $i = 0.65$ mm

The relationships between element length, S_{11} and normalized S_{21} parameters, at 32 GHz, are summarized below in Table 4.1.

Element Length (mm)	S_{11} (dB)	S_{21} (dB)
$\lambda_g/2 = 2.6$	-23	-2
$\lambda_g/4 = 1.3$	-11	-15
$\lambda_g/6 = 0.83$	-2.5	-53
$\lambda_g/8 = 0.65$	-3	-21

Table 4.1 –*Scattering Parameter Simulated Results at 32 GHz as a Function of Element Length*

The $\lambda_g/2$ elements do not produce radiation but rather cause the antenna to behave like a transmission line. According to transmission line theory impedance is periodic, repeating in multiples of $\lambda_g/2$. Therefore, element lengths that are multiples of $\lambda_g/2$ also, will not produce radiation [23]. The results for the $\lambda_g/6$ and $\lambda_g/8$ elements reveal that a large amount of the input power is being reflected and little power is being transmitted. Only the $\lambda_g/4$ elements have little transmission and are well matched indicating they produce good radiation. The passive structure design was adjusted to reflect these findings and element length was selected as $i = 1.3$ mm long.

4.3.2 Element Separation

The element separation influences a variety of antenna parameters including direction of radiation and scan range. The approximate Brillouin diagram for a 50 ohm transmission line operating at 32 GHz on a fused silica substrate was previously illustrated in Figure 4.2. From this diagram it can be seen that there is a wide range of periodic separation values d that fall within the fast-wave region. The periodic spacing values that fall within the fast-wave region, were determined by calculating

the equivalent k_0d and βd values. These results and the corresponding radiation angles are displayed in Table 4.2 and the Matlab [31] code is provided in section A.4 of the Appendix.

$\beta_1 d$	$k_0 d$	d (mm)	θ (Degrees)
-2.35	2.35	3.50	-90.00
-1.85	2.65	3.95	-44.19
-1.34	2.95	4.40	-27.12
-0.84	3.25	4.85	-15.01
-0.34	3.55	5.30	-5.48
0.16	3.85	5.74	2.43
0.67	4.15	6.19	9.23
1.17	4.45	6.64	15.22
1.67	4.75	7.09	20.60
2.17	5.05	7.53	25.49
2.68	5.35	7.98	30.01
3.18	5.65	8.43	34.23
3.68	5.95	8.88	38.22
4.18	6.25	9.32	42.02
4.69	6.55	9.77	45.68
5.19	6.85	10.22	49.24
5.69	7.15	10.67	52.74
6.19	7.45	11.11	56.24
6.70	7.75	11.56	59.77
7.20	8.05	12.01	63.41
7.70	8.35	12.46	67.26
8.20	8.65	12.90	71.51
8.70	8.95	13.35	76.58
9.21	9.25	13.80	84.55
9.31	9.31	13.90	90.00

Table 4.2 – *Element Separation and Corresponding Scan Range for the Fast-Wave Region*

Table 4.2 indicates that in theory it is possible to scan over the entire horizon. The element spacing values determined to be acceptable for this design were between 4 mm and 5 mm. From the data in Table 4.2 it can be seen that for such spacing a relatively small change in β_n can result in a large angular movement of the beam peak.

Scattering parameters simulations were performed for six separate passive structures with the following element separations: (i) $d = 4$ mm, (ii) $d = 4.2$ mm, (iii) $d = 4.4$ mm, (iv) $d = 4.6$ mm, (v) $d = 4.8$ mm and (vi) $d = 5$ mm. Apertures size was maintained more or less constant by adjusting the number of elements along the fixed length of each antenna as outlined below in Table 4.3.

Number of Elements	Element Separation (mm)
24	4.0
22	4.2
22	4.4
20	4.6
20	4.8
20	5.0

Table 4.3 – *Relationship between Element Separation and Number of Elements*

The S_{11} and S_{21} parameters are illustrated respectively in Figures 4.8 and 4.9 below.

Figure 4.8 – S_{11} Simulated Results for $d = 4$ mm, $d = 4.2$ mm, $d = 4.4$ mm, $d = 4.6$ mm, $d = 4.8$ mm and $d = 5.0$ mm

Figure 4.9 – S_{21} Simulated Results for $d = 4.0$ mm, $d = 4.2$ mm, $d = 4.4$ mm, $d = 4.6$ mm, $d = 4.8$ mm and $d = 5.0$ mm

Upon inspection of Figures 4.8, it can be seen that all of the antennas are well matched around 31.84 GHz. The shift in operating frequency from 32 GHz to 31.84 GHz occurs because the initial design procedure was based upon an approximate guided wavelength and Brillouin Diagram. The relationships between the element separation and the S_{11} and S_{21} parameters at 31.84 GHz are summarized below in Table 4.4.

Element Separation (mm)	S_{11} (dB)	S_{21} (dB)
4.0	-11.6	-33.3
4.2	-11.2	-33.7
4.4	-11.1	-32.8
4.6	-10.3	-28.3
4.8	-9.05	-25.2
5.0	-7.17	-19.2

Table 4.4 - *Summary of Scattering Parameter Simulated Results as a Function of Element Separation at 31.84 GHz*

Inspection of Figure 4.9 reveals that all of the structures radiate at 31.84 GHz since a maximum of only 1.25 % of the input power remains at the end of the antennas.

Radiation pattern simulations were also performed for the six separate passive structures outlined above. The simulated radiation results are illustrated in Figures 4.10.

Figure 4.10 – Simulated Gain at 31.84 GHz of the E_ϕ Component in the $\phi = 0$ Plane for $d = 4.0$ mm, $d = 4.2$ mm, $d = 4.4$ mm, $d = 4.6$ mm and $d = 5.0$ mm

The relationships between element separations, location of beam peak location and gain are summarized in Table 4.5.

Element Separation (mm)	Location of Beam Peak	Gain (dBi)
4.0	-53°	12.0
4.2	-41°	12.7
4.4	-32°	13.2
4.6	-24°	13.6
4.8	-17°	14.0
5.0	-11°	14.4

Table 4.5 - Summary of Radiation Simulated Results as a Function of Element Separation at 31.84 GHz

This range of element separation provides adequate gain for the final optical design.

The relationship between scan performance and element separation is further investigated in the subsequent sections.

4.3.3 Beam Scan Range

The relationship between element spacing and beam peak location is further investigated by varying element length for two separate antennas with elements separated by 4.2 mm and 4.8 mm, each having 22 and 20 elements, respectively. From the previous analysis these separations were shown to provide good radiation characteristics and cover a wide range of beam peak locations. In Section 4.3.1 it was determined that radiation occurs with an element length of approximately $\lambda_g/4$ or $i = 1.3$ mm.

4.3.3.1 Elements Separated by 4.2 mm

Scattering parameter simulations were performed for a 4.2 mm passive antenna with element lengths varying from 1.1 mm to 1.8 mm in 0.1 mm increments. The S_{11} and S_{21} simulated results are illustrated in Figures 4.11 and 4.12, respectively.

Figure 4.11 – S_{11} Simulated Results for Antennas with 4.2 mm Element Separation

Figure 4.12 – S_{21} Simulation Results for Antennas with 4.2 mm Element Separation

The resulting relationships between element length and the S_{11} and S_{21} parameters at 31.84 GHz are summarized in Table 4.6 below.

Element Length (mm)	S_{11} (dB)	S_{21} (dB)
1.1	-5.5	-32.3
1.2	-11.0	-53.2
1.3	-11.2	-33.7
1.4	-13.8	-22.4
1.5	-15.8	-18.1
1.6	-16.5	-14.3
1.7	-18.2	-12.3
1.8	-19.9	-10.9

Table 4.6 - *Summary of Scattering Parameter Simulated Results at 31.84 GHz for Antennas with 4.2 mm Element Separation*

The results reveal that at 31.84 GHz element lengths between 1.2 mm and 1.8 mm produce small reflections and are well matched indicating they produce radiation.

Radiation pattern simulations were also performed for the same configurations with results illustrated in Figure 4.13 below.

Figure 4.13 – Simulated Gain at 31.84 GHz for E_ϕ Component in the $\phi = 0$ Plane for Antennas with 4.2 mm Element Separation

The relationships between gain, beam peak location and 3 dB beam width are summarized in Table 4.7.

Element Length (mm)	Gain (dBi)	Beam Peak Location	3 dB Beam Width
1.1	4.6	-52°	53°
1.2	10.1	-47°	18°
1.3	12.7	-41°	11°
1.4	14.2	-36°	8°
1.5	14.8	-34°	7°
1.6	15.1	-32°	7°
1.7	15.3	-31°	7°
1.8	15.4	-30°	7°

Table 4.7 - Summary of Radiation Simulated Results at 31.84 GHz for Antennas with 4.2 mm Element Separation

The 1.1 mm, 1.2 mm and 1.3 mm structures all suffer from low gain and broad beam widths. From Table 4.6, the 1.1 mm element was shown to provide a poor match,

which may explain the low gain. The 1.2 mm and 1.3 mm element structures however, are well matched but continue to suffer from inferior radiation performance. An investigation of surface current using Ensemble [30] revealed that not all of the elements were radiating indicating that the structure is longer than required. In the case of the 1.1 mm, 1.2 mm and 1.3 mm structures fewer elements are excited and consequently the antennas have smaller apertures with reduced gain despite the fact the individual elements may radiate larger amounts of energy.

The array of elements investigated causes the beam to scan over a 22 degrees range in the direction of broadside. This trend indicates that structures with longer element lengths have smaller phase constants β_n which on a Brillouin diagram would translate into a more gradual slope for the $n = \pm 1$ space harmonic.

4.3.3.2 Element Separated by 4.8 mm

A similar scattering parameter analysis was carried out for an antenna with elements separated by 4.8 mm. The S_{11} and S_{21} simulated results are illustrated below in Figures 4.14 and 4.15, respectively and summarized in Table 4.8.

Figure 4.14 – S_{11} Simulated Results for Antennas with 4.8 mm Element Separation

Figure 4.15 – S_{21} Simulation Results for Antennas with 4.8 mm Element Separation

Element Length (mm)	S_{11} (dB)	S_{21} (dB)
1.1	-6.4	-58.2
1.2	-8.5	-34.5
1.3	-9.0	-25.2
1.4	-11.9	-18.7
1.5	-12.9	-14.8
1.6	-14.0	-12.7
1.7	-14.8	-11.0
1.8	-18.6	-9.8

Table 4.8 - *Summary of Scattering Parameter Simulated Results at 31.84 GHz for Antennas with 4.8 mm Element Separation*

As in the previous case the results indicate that at 31.84 GHz element lengths between 1.2 mm and 1.8 mm produce minimal reflections and are well matched indicating the structures radiate.

Radiation pattern simulations were performed for a 4.8 mm configuration and the results are illustrated in Figure 4.16 and summarized in Table 4.9.

Figure 4.16 – Simulated Gain at 31.84 GHz of the E_ϕ component in the $\phi = 0$ Plane for Antennas with 4.8 mm Element Separation

Element Length (mm)	Gain (dBi)	Beam Peak Location	3 dB Beam Width
1.1	9.3	-23°	10°
1.2	12.8	-19°	9°
1.3	14.0	-17°	6°
1.4	14.9	-15°	6°
1.5	15.3	-14°	6°
1.6	15.5	-13°	6°
1.7	15.5	-13°	6°
1.8	15.7	-12°	6°

Table 4.9 - Summary of Radiation Simulated Results at 31.84 GHz for Antennas with 4.8 mm Element Separation

The results are similar to those obtained for the 4.2 mm configuration except the range of beam scan is reduced to 11 degrees in the direction of broadside. The beam peak of the 4.8 mm structures is also located closer to broadside. In summary the 4.2

mm structure is preferable since it was found to provide a broader angular range for beam scanning.

4.4 Passive Antenna Design Summary

Based upon the preceding analysis the design parameters of the final passive antenna design are summarized as follows: (i) the operating frequency is 31.84 GHz, (ii) the substrate is 0.5 mm thick fused silica with a dielectric constant $\epsilon_r = 4.0$, (iii) the trace and ground planes are copper, (iv) the antenna length is $L = 100$ mm, (v) the input and output impedances of the antenna are both 50 ohms, (vi) the element thickness is $t = 1.02$ mm, (vii) the element separation is $d = 4.2$ mm and (viii) the element length is $i = 1.2$ mm.

The 4.2 mm element separation provides the greatest range for beam scanning, however it has slightly lower gain. The above antenna begins to radiate a discernable beam peak when the elements are 1.2 mm long. The optical design requires that the passive elements be further loaded with pieces of silicon. It is important that the passive elements minutely perturb the guiding structure so that when the silicon is added each entire element contributes to the radiation and the aperture size is not significantly reduced.

The final antenna design will be built at the microfabrication facility at the Communication Research Centre. This facility however, is unable to sputter copper or process structures larger than 3 inches. These fabrication limitations require that several alterations be made to the above antenna design. Since the laboratory is

capable of only sputtering gold both the conductor and the ground plane layers will be fabricated using gold rather than copper. The length of the antenna was also shortened from 100 mm to 50.8 mm (2 inches), reducing the number of elements from 22 to 10. S_{11} and S_{21} simulations for both the gold and original copper structures are similar and are illustrated in Figure 4.17 and Table 4.10 below.

Figure 4.17 – *Scattering Parameter Simulated Results for Gold and Copper Passive Antennas*

Antenna	S_{11} (dB)	S_{21} (dB)
Gold	-9.4	-24.7
Copper	-11.0	-53.2

Table 4.10 - *Comparison of Scattering Parameter Simulated Results at 31.84 GHz for Gold and Copper Passive Antennas*

Radiation pattern simulations were also compared for both the 10 element gold and the original 22 element copper antennas and the results are illustrate below in Figure 4.12 and Table 4.11.

Figure 4.18 – Simulated Gain at 31.84 GHz for E_ϕ Component in the $\phi = 0$ Plane for Gold and Copper Passive Antennas

Antenna	Co-pol Gain (dBi)	Co-pol Beam Peak Location	3 dB Beam Width of Co-pol	Cross-pol Gain (dBi)	Cross-Polarization Level (dBi)
Gold	9.8	-45°	20°	-2.0	11.91
Copper	10.1	-47°	18°	-4.0	14.23

Table 4.11 - Comparison of Radiation Simulated Results at 31.84 GHz for Gold and Copper Passive Antennas

Despite the reduction in the number of elements, both antennas exhibit similar radiation performance. This is based upon the fact that the original 22 element copper design was oversized and not all of the elements were in fact radiating. The simulations show that the gold antenna has slightly lower gain and a broader beam width. The beam peak of both structures is in relatively the same location with gold causing only a 2° shift away from broadside. The beam shift indicates that the phase

constant of the gold structure is slightly larger resulting from a more gradual slope of the $n = -1$ space harmonic.

4.5 Optical Design

Section 4.4 addressed the design of a passive leaky-wave antenna. This section deals with the design of optical control elements and their integration within the passive structure to provide beam scanning.

A photo-induced plasma layer on the silicon elements of a periodic leaky-wave antenna, as shown in Figure 4.19, below will cause the beam peak to scan by changing the phase constant of the structure. The dimensions of the silicon elements are shown to be critical in determining the overall radiation performance. Several factors limit the dimensions of the silicon elements including the periodic spacing of the gold elements, previously selected to be 4.2 mm. Also, there is a need to limit the size of the holes in the ground plane required for optical illumination to minimize backside radiation losses. The thickness of the silicon element was set at 0.5 mm based upon available wafers, and the width w was fixed at 2 mm. This width allows a minimum separation s of 2.2 mm between individual silicon elements. The effect of varying the length of the silicon element y is investigated in the following section.

Figure 4.19 – Optical Antenna Geometry

4.5.1 Silicon Element Length

The length of the silicon element was varied and the effects on radiation performance analyzed. Scattering parameter simulations were performed using HFSS software [34] for structures with lengths ranging from $y = 4$ mm to $y = 7$ mm in 1 mm increments. HFSS software was used because it can simulate complex structures such as a plasma layer while Ensemble cannot [34]. The holes in the ground plane were set at the same dimensions as the corresponding silicon elements. The resulting S_{11} and S_{21} parameters are illustrated in Figures 4.20 and 4.21, respectively and the 31.84 GHz results are summarized in Table 4.12.

Figure 4.20 – S_{11} Simulated Results for a Dark Antenna with $y = 4$ mm, $y = 5$ mm, $y = 6$ mm and $y = 7$ mm

Figure 4.21 – S_{21} Simulated Results for a Dark Antenna with $y = 4$ mm, $y = 5$ mm, $y = 6$ mm and $y = 7$ mm

Silicon Length (mm)	S_{11} (dB)	S_{21} (dB)
4	-21.4	-19.7
5	-9.8	-46.4
6	-9.4	-26.2
7	-14.3	-18.2

Table 4.12 - Summary of Scattering Parameter Simulated Results at 31.84 GHz as a Function of Silicon Length

This data reveals that all of the antenna structures are radiating and are well matched.

Investigation of radiation pattern simulations showed optimum radiation at 31 GHz.

The frequency shift from 31.84 to 31 GHz is due to the effect of the holes in the ground plane. These results are summarized in Figure 4.22 and Table 4.13 below.

Figure 4.22 – Simulated Gain at 31 GHz of E_ϕ at $\phi = 0$ Plane for a Dark Antenna with $y = 4$ mm, $y = 5$ mm, $y = 6$ mm and $y = 7$ mm

Silicon Length (mm)	Gain (dBi)	Location of Front Lobe (Beam Peak)
4	7.7	-34°
5	2.8	-30°
6	0.9	-26°
7	7.7	-35°

Table 4.13 - *Summary of Simulated Radiation Results at 31 GHz as a Function of Silicon Length*

Comparison of the antenna gain of these structures with those of Figure 4.13 and 4.16 reveal a considerable loss in gain. This is most likely due to high backside radiation through the ground plane. All of the antenna structures suffer from relatively high backside radiation through the ground plane holes. The 6 mm structure was eliminated from further consideration due to its low gain. Since the performance of the 4 mm and 7 mm structures are almost identical, only the 7 mm and the 5 mm configurations are carried forward for further analysis under optical illumination.

4.5.2 Photoconductivity Characterization

Carrier density levels required for beam scanning have been obtained from available literature. Research in this area indicates that an optical intensity of 20.4 W/cm^2 will optically induce a carrier density of 10^{17} cm^{-3} . At this level of excitation the silicon behaves similar to a conductor [36].

The complex permittivity and plasma penetration depth for a silicon wafer are calculated using Equations 3.28, 3.29 and 3.30. These parameters are required for simulation purposes and were evaluated using Matlab [31] at 31.0 GHz for the following excess carrier concentrations: (i) $n = 10^{17} \text{ cm}^{-3}$, (ii) $n = 10^{16} \text{ cm}^{-3}$ and (iii) n

$= 10^{15} \text{ cm}^{-3}$. The above calculations assume the effective electron mass is 0.33 times the electronic charge and the electron mobility is $1500 \text{ cm}^2/\text{V}$ [37] and [30]. The plasma depth calculations assume, a temperature of 300 K, a surface recombination rate is 1 m/s [30], electron and hole mobility of $1500 \text{ cm}^2/\text{V}$ and $600 \text{ cm}^2/\text{V}$ respectively, radiation absorption coefficient of $4.0 \times 10^6 \text{ m}^{-1}$ at 900 nm [21] and excess carrier lifetime of 1 μs [30]. The corresponding Matlab [31] code is found in section A.5 of the Appendix.

The plasma depth was found to be 47.34 μm and the complex permittivity results are displayed in Table 4.14 below.

Carrier Density (cm^{-3})	Complex Permittivity
10^{17}	$-64.40 - j1388$
10^{16}	$4.17 - j138.80$
10^{15}	$11.03 - j13.88$

Table 4.14 – *Complex Permittivity as a Function of Carrier Density at 31 GHz*

From the imaginary component of the complex permittivity in Table 4.14 it can be seen that the assumed excess carrier concentration levels cover a broad range from a quasi dielectric to a quasi metal.

4.5.3 Optical Scan Range Simulations

The 5 mm and 7 mm antenna configurations discussed in Section 4.5.1 are simulated over the range of complex permittivity values calculated in Section 4.5.2. The purpose of the simulations is to identify the configuration that provides the best combination of beam scanning and radiation efficiency.

Scattering parameter simulations were performed and the results for both structures are illustrated below in Figures 4.23, 4.24, 4.25 and 4.26 and are summarized in Table 4.15.

Figure 4.23 – S_{11} Simulated Results for the 5 mm Structure with No Illumination, $z = 10^{15} \text{ cm}^{-3}$, $z = 10^{16} \text{ cm}^{-3}$ and $z = 10^{17} \text{ cm}^{-3}$

Figure 4.24 – S_{21} Simulated Results for the 5 mm Structure with No Illumination, $z = 10^{15} \text{ cm}^{-3}$, $z = 10^{16} \text{ cm}^{-3}$ and $z = 10^{17} \text{ cm}^{-3}$

Figure 4.25 – S_{11} Simulated Results for the 7 mm Structure with No Illumination, $z = 10^{15} \text{ cm}^{-3}$, $z = 10^{16} \text{ cm}^{-3}$ and $z = 10^{17} \text{ cm}^{-3}$

Figure 4.26 – S_{21} Simulated Results the for 7 mm Structure with No Illumination, $z = 10^{15} \text{ cm}^{-3}$, $z = 10^{16} \text{ cm}^{-3}$ and $z = 10^{17} \text{ cm}^{-3}$

Carrier Density (cm^{-3})	5 mm Structure		7 mm Structure	
	S_{11} (dB)	S_{21} (dB)	S_{11} (dB)	S_{21} (dB)
No Illumination	-16.3	-54.5	-8.8	-25.9
10^{15}	-14.8	-35.6	-11.5	-30.5
10^{16}	-13.4	-31.1	-17.0	-28.1
10^{17}	-14.9	-20.6	-11.2	-35.0

Table 4.15 - Summary of Scattering Parameter Simulated Results at 31 GHz for the 5mm and 7 mm Antenna Structures as a Function of Carrier Density

The results indicate that at 31 GHz all of the antenna structures are well matched and radiating. The 5 mm structure, however appears to have a slightly better match.

Radiation pattern simulations were also carried out for both of the antenna structures and the results are shown in Figures 4.27 and 4.28 and Tables 4.16 and 4.17 below.

Figure 4.27 – Simulated Gain at 31 GHz for the E_ϕ Component in the $\phi = 0$ Plane for the 5 mm Structure with No Illumination, $z = 10^{15} \text{ cm}^{-3}$, $z = 10^{16} \text{ cm}^{-3}$ and $z = 10^{17} \text{ cm}^{-3}$

Figure 4.28 - Simulated Gain at 31 GHz for the E_ϕ Component in the $\phi = 0$ Plane for the 7 mm Structure with No Illumination, $z = 10^{15} \text{ cm}^{-3}$, $z = 10^{16} \text{ cm}^{-3}$ and $z = 10^{17} \text{ cm}^{-3}$

Carrier Density (cm ⁻³)	Gain (dBi)	Location of Front Lobe (Beam Peak)	3 dB Beamwidth	Back Lobe Level (dBi)	Front to Back Ratio
No Illumination	2.8	-31°	16°	0.7	2.1
10 ¹⁵	-0.3	-29°	18°	-1.2	0.9
10 ¹⁶	-0.2	-28°	18°	-6.2	6.0
10 ¹⁷	5.7	-33°	16°	-11.7	17.4

Table 4.16 – *Summary of Simulated Radiation Results at 31 GHz for the 5 mm Structure as a Function of Carrier Density*

Carrier Density (cm ⁻³)	Gain (dBi)	Location of Front Lobe (Beam Peak)	3 dB Beamwidth	Back Lobe Level (dBi)	Front to Back Ratio
No Illumination	7.7	-34°	18°	-2.2	9.9
10 ¹⁵	5.5	-31°	20°	-2.5	8.0
10 ¹⁶	2.3	-29°	19°	-3.6	5.9
10 ¹⁷	-2.7	-31°	19°	-4.9	2.2

Table 4.17 – *Summary of Simulated Radiation Results at 31 GHz for the 7 mm Structure as a Function of Carrier Density*

Comparison of antenna patterns in Figures 4.27 and 4.28 with those previously shown in Figure 4.13 show that the combination of silicon elements and holes in the ground plane degrades the performance of the antenna due to backside radiation.

Both structures exhibit a 5 degree scan range. This scan range is considerably less than what was expected based upon the passive design simulations. In the passive design an element separation d of 4.2 mm was selected since a small change in β_n should produce a large beam scan as previously demonstrated in Table 4.2. Also, the data in Table 4.7 indicates that small changes in element length should perturb the phase constant β_n sufficiently to produce noticeable beam scanning of approximately 22 degrees.

In formulating the original design of the structure it was originally assumed that the rectangular silicon elements would significantly perturb the phase constant β_n and therefore yield a large beam scan range. This original assumption does not appear to be consistent with the above simulations. This must be due to a shift in the operating point between the passive and optical antenna structures to a point that is in close proximity to a stop band region. In this region a change in the loading effects of the periodic elements can result in only a small change in the phase constant β_n thus limiting the range of beam scanning.

Comparison of the data in Tables 4.16 and 4.17 shows that the 5 mm structure exhibited a narrower beam peak and a better front-to-back ratio at the maximum and minimum values of the scan range. Based on the above considerations the 5 mm structure was chosen for implementation. Although, the scan range is not optimum it is still measurable for the purposes of establishing proof-of-concept for optical control. It is also interesting to note that in both antenna configurations the scan range is not monotonically related to excess carrier density.

For completeness, the cross polarization performance of the 5 mm structure was also simulated at the extreme ends of the scan range and the results are shown in Figure 4.29 and Table 4.18 below.

Figure 4.29 – *Summary of Co-Polarization and Cross Polarization Simulated Results at 31 GHz for the 5 mm Antenna Structure*

Carrier Density (cm ⁻³)	Co-pol Gain (dBi)	Co-pol Beam Peak Location	Cross-pol Gain (dBi)	Cross-Polarization Level (dBi)
10 ¹⁷	5.7	-33°	-15.9	21.6
10 ¹⁶	-0.2	-28°	-8.3	8.1

Table 4.18 – *Summary of Simulated Co-polarization and Cross Polarization Results at 31 GHz*

4.6 Design Summary

A microstrip leaky-wave antenna with optical beam scanning capabilities was designed. A practical antenna architecture was derived from an initial assessment of critical design parameters including, the effect of substrate material on insertion loss, radiation characteristics and the fast-wave region as well as implementation practicality. A detailed design procedure of a passive antenna, consisting of

theoretical analysis and simulations was performed. This design procedure established the effects of basic antenna geometry on radiation performance and scan range. Fabrication constraints were considered and the passive antenna design was revised. Optical control elements were integrated into the passive design and simulations carried out to optimize the antenna's radiation and scan performance. The final microstrip leaky-wave antenna architecture consisted of ten $2 \times 5 \text{ mm}^2$ silicon element separated by 4.2 mm and periodically placed along the length of a 10 cm transmission line. Simulations indicate that optical illumination of the silicon elements with excess carrier concentration densities ranging up to 10^{17} cm^{-3} produces 5 degrees of beam scanning.

CHAPTER 5

OPTICALLY CONTROLLED LEAKY-WAVE ANTENNA MEASUREMENTS AND PROPOSED IMPROVEMENTS

This chapter presents measured results for the antenna design previously outlined in Chapter 4. The test-set up for scattering parameter and radiation pattern measurements is briefly described. Also, the method to measure the light intensity of the optical source is outlined. Measurements and simulated results are presented and compared and improvements to the leaky-wave antenna are proposed.

5.1 Antenna Fabrication and Experimental Setup

The metalized antenna pattern was created using lithography. Pieces of silicon were cut and manually placed along the antenna structure. These elements were visually aligned with the ground plane holes to ensure electrical contact then adhered using a bead of silicone along the perimeter. The antenna was attached to the experimental setup using a rectangular aluminum jig into which a window was cut to allow light to illuminate the rear of the antenna. Top and bottom views of the antenna with coaxial connectors mounted to the aluminum jig are shown in Figures 5.1 and 5.2, respectively.

Figure 5.1 – *Top View of Antenna and Coaxial Connectors Mounted to Aluminum Jig Illustrating Microstrip Line and Silicon Elements*

Figure 5.3 – *Bottom View of Antenna and Coaxial Connectors Mounted to Aluminum Jig Illustrating Ground Plane Holes*

The experimental set-up used for the measurements is depicted in Figure 5.3. The setup consists of an antenna mount, a light source, a modified heat sink, a reflector and a fan. The high light energy levels require that the antenna be thermally shielded to prevent heat damage.

Figure 5.3 – Experimental Setup

5.2 Optical Characterization

The available light source was a 300 W halogen bulb with the spectral content ranging from approximately 600 nm to 2000 nm and fed with a 115 AC Volt variable power supply. The spectral response of the silicon material has been shown to generate a maximum response at a wavelength of approximately 900 nm [24]. The intensity of the bulb over the entire spectrum was measured using an optical power sensor. This characterization was performed with maximum power supplied to the light source and the resulting intensity level was found to be 1.16 W/cm^2 . This intensity level was markedly lower than the 20.4 W/cm^2 required to generate the desired excess carrier concentrations. Due to the unavailability of a higher intensity light source the range of the resulting measurements were severely limited.

5.3 Measured Results

Scattering parameter measurements were performed on the antenna using a network analyzer without illumination. The light intensity was then increased to 1.16 W/cm^2 and a slight change in the measured results was observed. The S_{11} and S_{21} parameters for both the dark and illuminated states are displayed in Figures 5.5 and 5.6, respectively.

Figure 5.5 – S_{11} Measured Results in Dark and Illuminated States

Figure 5.6 – S_{21} Measured Results in Dark and Illuminated States

The data indicates that in both states a good match is achieved at frequencies ranging between 27 GHz and 33 GHz. The best input match occurs at 29.5GHz, however at this frequency the power transmitted through the output terminal is at a maximum.

These results show that illumination has little effect on the antenna's scattering parameters, indicating the photo-induced excess carrier concentration at this illumination level is very small.

Radiation pattern measurements were also performed in an anechoic chamber with a pyramidal horn used as an energy source. One port of the antenna was connected to a receiver and the other terminated with a 50 ohm load to prevent reflections. A large number of 180 degree pattern measurements were taken at various frequencies

ranging between 27 GHz and 33 GHz in the dark and illuminated states. The best results in terms of antenna gain occurred at a frequency of 31.6 GHz. The corresponding measured radiation pattern is illustrated below in Figure 5.7 and in Table 5.1.

Figure 5.7 –*Measured Radiation Patterns at 31.6 GHz in the $\phi = 0$ Plane in Dark and Illuminated States*

Electric Field Component	Gain (dBi)	Location of Front Lobe (Beam Peak)	3 dBi Beamwidth
Co-pol: Dark	-1.4	-24.6°	26.9°
Cross-pol: Dark	1.1	33.4°	25.9°
Co-pol: Illuminated	-0.5	-23.5°	26.0°
Cross-pol: Illuminated	2.3	37.3°	17.3°

Table 5.1 – *Summary of Measured Radiation Patterns in the Dark and Illuminated States*

From the above data it can be seen that at these relatively low light intensity levels illumination has little if any effect on antenna performance. The measured beam scan

of 1.06 degrees in the direction of broadside is inconclusive since it falls within experimental error of the set-up due to the broad and irregular pattern.

Comparison of the measured results with the previous simulations shows a 75 degree shift in the maximum cross-polarized field component as illustrated in Figure 5.8 and Table 5.2.

Figure 5.8 – *Measured and Simulated Radiation Patterns in the $\phi = 0$ Plane in the Dark State*

Antenna	Co-Pol Gain (dBi)	Co-pol Beam Peak Location	Cross-pol Gain (dBi)	Cross Polarization Levels (dBi)
Simulated	2.8	-31°	-5.2	8.0
Measured	-1.48	-24°	-9.0	7.52

Table 5.2 – *Summary of Measured and Simulated Patterns in the Dark State*

The radiation performance of the measured antenna is degraded due to the effects of the large coaxial connector and reflection from the back of the antenna and the experimental setup.

5.4 Proposed Improvements

The beam scanning performance of the prototype antenna is limited since only a low power optical source was available for testing. Replacing this source with a high power laser would make it possible to achieve the large excess carrier concentration densities required for beam scanning. This high power scheme may have significant thermal issues and require additional shielding. Another alternative would be to replace the single source with several smaller sources. These sources could include fibers or light emitting diodes and would provide high intensity local excitation. The disadvantage of this illumination technique is that it would require complex assembly and may not be low cost. For example, forty 5 mW LED diodes would be required to achieve the optical intensities on the order of 20 W/cm^2 required for optical beam scanning.

Also, tuning the spectrum of the optical source to that of the semiconductor's band gap would induce a larger change in phase constant thus providing large scan range. This could be achieved by filtering the light source.

The radiation performance of the antenna is poor because the holes in the ground plane radiate a large amount of energy. Removing the holes would minimize the

backside radiation and improve the structure's gain. A radiation pattern simulation was performed for an antenna with $2 \times 5 \text{ mm}^2$ pieces of silicon without holes in the ground plane. These results are compared with simulations of a similar structure with holes in the ground plane and are depicted in Figure 5.9 and summarized Table 5.3.

Figure 5.9 – *Simulated Gain at 31 GHz of E_ϕ at $\phi = 0$ Plane for Dark Antennas With and Without Holes in the Ground Plane*

Antenna	Gain (dBi)	Back Lobe Level (dBi)	Front to Back Ratio
Holes	2.8	0.7	2.1
No Holes	12.7	-6.7	19.4

Table 5.3 – *Summary of Simulated Gain at 31 GHz for Dark Antennas With and Without Holes in the Ground Plane*

These simulations indicate that the holes in the ground plane are reducing the antenna's gain and front to back ratio by 9.9 dBi and a factor 17.3, respectively.

Replacing the antenna's ground plane with an optically transparent metal such as indium tin oxide would significantly improve radiation performance. There is however, a trade off between conductivity and transparency associated indium tin oxide.

Integrating the silicon elements and ground plane holes into the passive design shifted the antenna's operating point to close proximity of a stop band region. This shift reduced the antenna's original 22 degree scan range to only 5 degrees. Redesigning the geometry of the structure to move the operating point away from the stop band would improve the antenna's scan range.

5.5 Conclusions

The optical source had insufficient intensity to produce a measurable effect on the antenna's beam peak location. In addition, the antenna's measured radiation performance is degraded due to large coaxial connector and reflection from the back of the antenna and the experimental setup. The measured results are inconclusive, however, theory and simulations indicate optical beam scanning integrated into a leaky-wave microstrip antenna is feasible. A proof-of-concept demonstration would require implementing several of following design improvements: (i) a higher optical intensity source, (ii) tuning the spectrum of the optical source to that of the semiconductor's band gap, (iii) eliminating the holes in the ground plane and (iv) selecting an operating point away from any stop band regions.

CHAPTER 6

SUMMARY AND FUTURE WORK

6.1 Summary

The preliminary feasibility of using optical illumination to control beam scanning in a microstrip leaky-wave antenna has been investigated. This design approach in principle allows the use of economic batch microstrip fabrication techniques while simultaneously providing the possibility for good side lobe performance, high efficiency and low profile characteristics associated with leaky-wave architectures. Optical control enables ultra fast high power shifting without the use of complex expensive biasing networks potentially resulting in a desirable antenna architecture suitable for next-generation wireless systems.

A leaky-wave antenna architecture was proposed and a passive design approach was used to initially characterize the effects of antenna geometry on radiation and beam scanning performance. The effect of element separation on beam peak location and scan range was investigated. An attempt was made to select a separation that potentially could provide a large scan range based upon small perturbations in the phase constant β_n . This was determined to be on the order of 4.2 mm. The effect of element length on beam scanning performance was also assessed to determine regions of high sensitivity to variations in β_n . This was determined to be on the order of 1.2 mm. In all of the above parameter selections an effort was made to provide good radiation performance and high gain.

The passive design was enhanced with optical control by integrating silicon wafers onto the periodic metal elements. Holes were added to the ground plane to enable optical illumination. These holes were shown to produce significant backside radiation, which degraded antenna performance. Complex permittivity values were calculated for a range of excess carrier concentration densities previously shown to produce a significant effect on the electrical properties of silicon. These permittivity values were then used to simulate the effects of optical illumination on the antenna structure. These simulations demonstrated that the above alterations to the passive design resulted in a significant reduction in scan range and degradation in radiation performance. This was attributed to a shift in operating point to region close to a stop band with poor radiation performance and the onset of backside radiation.

Despite the sub-optimum predicted performance the existing design was determined to be sufficient for proof of concept and was implemented. The lengthy time period required for fabrication did not allow for re-design efforts. Measured results were compared to the simulations. Due to the limited intensity of the available optical source, significant phase constant perturbations were not achievable. As a result of the low light intensities and the degraded radiation patterns the experimental results were inconclusive in determining the feasibility of optical beam scanning control.

6.2 Future Work

In theory, it should be possible to demonstrate the feasibility of incorporating optical control for periodic microstrip leaky-wave antennas. Areas of future work to verify proof-of-concept include, (i) a more detailed investigation of antenna geometry to determine an optimum operating point that is well within the structure's pass band region, (ii) implementing a high intensity optical source to increase excess carrier concentration density levels, (iii) tuning the optical spectrum of the light source to that of the semiconductor band gap to induce a larger perturbation of the phase constant and (iv) avoiding backside radiation by implementing an optically transparent ground plane to reduce pattern degradation and increase gain making it easier to assess beam scanning performance.

Despite inconclusive measured results, theory and simulations indicate optical beam scanning can be integrated into a microstrip periodic leaky-wave antenna.

Implementing these recommendations will lead to improved performance establishing the feasibility of the above antenna design.

APPENDIX

A.1 Silicon Loss Calculations

```
%-----
% Dielectric and Conductor Losses
% This program calculates the losses for a 50 ohm transmission line
on
% 0.5 mm thick silicon substrate material.
%-----

% Dielectric Constant
Er = 11.7;
% Resistivity ohm*m
res = 40;
% Conductivity in S/m
si_cond = 1/res;
% Thickness of sample in m
h = 0.5*10^-3;
% Impedance of transmission line in ohms
Zo = 50;
% Frequency of operation in Hz
f = 32*10^9;
% Angular frequency
w_rad = 2*pi*f;
% permittivity H/m
Eo = (1/(36*pi))*10^-9;
% loss tangent
l_tan = si_cond/(w_rad*Eo*Er);
% Speed of light m/s
c = 3*10^8;
% Unguided Wavelength
WL = c/f;

% *****Copper *****
% Permeability in H/m
uo = 4*pi*10^-7;
% Conductivity of Copper in S/m
cu_cond = 5.813*10^7;
% Skin depth in m
skin_depth = sqrt(2/(w_rad*uo*cu_cond));
% Select the thickness of copper to be 1 um (thicker than skin
depth)
t = 1*10^-6;

% Calculate thickness of 50 ohm transmission line
A = (Zo/60)*sqrt((Er+1)/2)+((Er-1)/(Er+1))*(0.23+(0.11/Er));

if A >1.52
    w_h = 8*exp(A)/(exp(2*A)-2);
else
    B = 60*pi^2/(Zo*sqrt(Er));
```

```

    w_h = (2/pi)*(B-1-log(2*B-1)+((Er-1)/(2*Er))*(log(B-1)+0.39-
(0.61/Er)));
end

'Width of 50-ohm transmission line in m'
w = w_h*h

% Calculate effective dielectric constant at 32 GHz
Eeff = ((Er+1)/2) + ((Er-1)/2)*(1+12*h/w)^(-0.5);

% Calculate and Dielectric Losses in dB/m

% Guided Wavelength in m

'The Dielectric Losses in dB/m are:'
alpha_d = (27.3*Er/sqrt(Eeff))*(Eeff-1)/(Er-1)*l_tan/WL

% Calculate delta_w in m

if w/h <= 1/(2*pi)
    delta_w = t/pi * (log(4*pi*w/t)+1);
else
    delta_w = t/pi * (log(2*h/t)+1);
end

%Calculate w prime in m

w_prime = w+delta_w;

% Calculate the surface resistance of gold in ohms
Rs = sqrt(w_rad*uo/(2*cu_cond));

% Calculate conductor losses in dB/m

'The conductor losses in dB/m are:'

if w/h <= 1/(2*pi)
    alpha_c = Rs*8.68/(Zo*h*2*pi) * (1 - (w_prime/4*h)^2) * (1 +
h/w_prime + (h/(pi*w_prime)) * ( log(4*pi*w/t) + t/w))
end

if (w/h > 1/(2*pi)) && (w/h <= 2)
    alpha_c = Rs*8.68/(Zo*h*2*pi) * (1 - (w_prime/4*h)^2) * (1 +
h/w_prime + (h/(pi*w_prime)) * ( log(2*h/t) - t/h))
end

if w/h >= 2
    alpha_c = (Rs*8.68/(Zo*h))/((w_prime/h +
(2/pi)*log(2*pi*exp(1)*w_prime/(2*h) +0.94)))^2 * ( (w_prime/h) +
(w_prime/(pi*h))/( (w_prime/(2*h)) +0.94)) * (1+ h/w_prime +
h/(pi*w_prime) * (log(2*h/t) - t/h))
end

```

A.2 Fused Silica Loss Calculations

```
%-----
% Dielectric and Conductor Losses
% This program calculates the losses for a 50 ohm transmission line
on
% a 0.5 mm thick fused silica substrate
%-----

% Dielectric Constant
Er = 4;
% Resistivity ohm*m
res = 40;
% Conductivity in S/m
si_cond = 1/res;
% Thickness of sample in m
h = 0.5*10^-3;
% Impedance of transmission line in ohms
Zo = 50;
% Frequency of operation in Hz
f = 32*10^9;
% Angular frequency
w_rad = 2*pi*f;
% permittivity H/m
Eo = (1/(36*pi))*10^-9;
% loss tangent
l_tan = 0.003
% Speed of light m/s
c = 3*10^8;
% Unguided Wavelength
WL = c/f;

% *****Copper *****
% Permeability in H/m
uo = 4*pi*10^-7;
% Conductivity of Copper in S/m
cu_cond = 5.813*10^7;
% Skin depth in m
skin_depth = sqrt(2/(w_rad*uo*cu_cond));
% Select the thickness of copper to be 1 um (thicker than skin
depth)
t = 1*10^-6;

% Calculate thickness of 50 ohm transmission line
A = (Zo/60)*sqrt((Er+1)/2)+((Er-1)/(Er+1))*(0.23+(0.11/Er));

if A >1.52
    w_h = 8*exp(A)/(exp(2*A)-2);
else
    B = 60*pi^2/(Zo*sqrt(Er));
    w_h = (2/pi)*(B-1-log(2*B-1)+((Er-1)/(2*Er))*(log(B-1)+0.39-
(0.61/Er)));
end

'Width of 50-ohm transmission line in m'
```

```

w = w_h*h

% Calculate effective dielectric constant at 32 GHz

Eeff = ((Er+1)/2) + ((Er-1)/2)*(1+12*h/w)^(-0.5);

% Calculate and Dielectric Losses in dB/m

% Guided Wavelength in m

'The Dielectric Losses in dB/m are:'
alpha_d = (27.3*Er/sqrt(Eeff))*(Eeff-1)/(Er-1)*l_tan/WL

% Calculate delta_w in m

if w/h <= 1/(2*pi)
    delta_w = t/pi * (log(4*pi*w/t)+1);
else
    delta_w = t/pi * (log(2*h/t)+1);
end

%Calculate w prime in m

w_prime = w+delta_w;

% Calculate the surface resistance of gold in ohms

Rs = sqrt(w_rad*uo/(2*cu_cond));

% Calculate conductor losses in dB/m

'The conductor losses in dB/m are:'

if w/h <= 1/(2*pi)
    alpha_c = Rs*8.68/(Zo*h*2*pi) * (1 - (w_prime/4*h)^2) * (1 +
h/w_prime + (h/(pi*w_prime)) * ( log(4*pi*w/t) + t/w))
end

if (w/h > 1/(2*pi)) && (w/h <= 2)
    alpha_c = Rs*8.68/(Zo*h*2*pi) * (1 - (w_prime/4*h)^2) * (1 +
h/w_prime + (h/(pi*w_prime)) * ( log(2*h/t) - t/h))
end

if w/h >= 2
    alpha_c = (Rs*8.68/(Zo*h))/((w_prime/h +
(2/pi)*log(2*pi*exp(1)*w_prime/(2*h) +0.94)))^2 * ( (w_prime/h) +
(w_prime/(pi*h))/( (w_prime/(2*h)) +0.94)) * (1+ h/w_prime +
h/(pi*w_prime) * (log(2*h/t) - t/h))
end

```

A.3 Brillouin Diagram

```
%-----
% This program plots a k/b curve for a 50 ohm transmission line on a
% fused silica and silicon 0.5 mm thick substrate material.
%-----

% Calculate Effective Dielectric Constant
% Dielectric Constant
Er = 11.7;
% Resistivity ohm*m
res = 40;
% Conductivity in S/m
si_cond = 1/res;
% Thickness of sample in m
h = 0.5*10^-3;
% Impedance of transmission line in ohms
Zo = 50;
% Frequency of operation in Hz
f = 32*10^9;
% Angular frequency
w_rad = 2*pi*f;
% permittivity H/m
Eo = (1/(36*pi))*10^-9;
% loss tangent
l_tan = si_cond/(w_rad*Eo*Er);
% Speed of light m/s
c = 3*10^8;
% Unguided Wavelength
WL = c/f;

Er = 11.7;

% Calculate thickness of 50 ohm transmission line

A = (Zo/60)*(Er+1/2)^0.5+(Er-1/(Er+1))*(0.23+(0.11/Er));

if A >1.52
    w_h = 8*exp(A)/(exp(2*A)-2);
else
    B = 60*pi^2/(Zo*sqrt(Er));
    w_h = (2/pi)*(B-1-log(2*B-1)+(Er-1/(2*Er))*(log(B-1)+0.39-
0.61/Er));
end

w = w_h*h;

% Calculate effective dielectric constant at 32 GHz

Eeff = ((Er+1)/2) + ((Er-1)/2)*((1+12*h/w)^(-0.5));

%Setup the slope for the fast wave regions x - +/-y
```

```

x=-12*pi:2*pi:12*pi;
y=abs(x);

%Setup the slope for the spatial harmonics
% ie for N = -1 ; y = +/- (Eeff)^-0.5 * x -2*pi

% Calculate the slope of the spatial harmonics
L=abs(1/sqrt(Eeff));

% Spatial Harmonic N=0
p=-24*pi:2*pi:24*pi;
z=abs(p*L);

% Spatial Harmonic N = -1
p1=p-2*pi;
z1=z;

% Spatial Harmonic N = 1
p2=p+2*pi;
z2=z;

% Spatial Harmonic N = -2
p3=p1-2*pi;
z3=z;

% Spatial Harmonic N = 2
p4=p2+2*pi;
z4=z;

Er_2 = 4;

% Calculate thickness of 50 ohm transmission line

A = (Zo/60)*(Er_2+1/2)^0.5+(Er_2-1/(Er_2+1))*(0.23+(0.11/Er_2));

if A >1.52
    w_h = 8*exp(A)/(exp(2*A)-2);
else
    B = 60*pi^2/(Zo*sqrt(Er_2));
    w_h = (2/pi)*(B-1-log(2*B-1)+(Er_2-1/(2*Er_2))*(log(B-1)+0.39-
0.61/Er_2));
end

w = w_h*h;

% Calculate effective dielectric constant at 32 GHz

Eeff_2 = ((Er_2+1)/2) + ((Er_2-1)/2)*((1+12*h/w)^(-0.5));

%Setup the slope for the spatial harmonics
% ie for N = -1 ; y = +/- (Eeff)^-0.5 * x -2*pi

% Calculate the slope of the spatial harmonics

```

```

L_2=abs(1/sqrt(Eeff_2));

% Spatial Harmonic N=0
p_2=-24*pi:2*pi:24*pi;
z_2=abs(p_2*L_2);

% Spatial Harmonic N = -1
p1_2=p_2-2*pi;
z1_2=z_2;

% Spatial Harmonic N = 1
p2_2=p_2+2*pi;
z2_2=z_2;

% Spatial Harmonic N = -2
p3_2=p1_2-2*pi;
z3_2=z_2;

% Spatial Harmonic N = 2
p4_2=p2_2+2*pi;
z4_2=z_2;

% Send Data of Fast wave region and the spatial harmonics normalized
to pi

save x.out x -ASCII
save y.out y -ASCII
save p1.out p1 -ASCII
save z.out z -ASCII
save p1_2.out p1_2 -ASCII
save z_2.out z_2 -ASCII

plot((x/pi),(y/pi),(p1/pi),(z/pi),(p1_2/pi),(z_2/pi))

```

A.4 Fast-Wave Region

```
%-----
% This program graphically determines the fast wave range kod for a
50
% ohm transmission line on a 0.5 mm thick fused silica substrate
% material. The intercept between the n =-1 space harmonic and the
fast % wave region are calculated.
% -----

% Dielectric constant of fused silica
Er_2 = 4;

% Transmission line impedance
Zo =50;

%thickness of substrate material in m
h = 0.5 *10^-3;

%Frequency of operation in Hz
f = 32*10^9;

%Speed of light in m/s
c = 3*10^8;

% Wavelength in freespace
WL = c/f;

% Calculate thickness of 50 ohm transmission line

A = (Zo/60)*(Er_2+1/2)^0.5+(Er_2-1/(Er_2+1))*(0.23+(0.11/Er_2));

if A >1.52
    w_h = 8*exp(A)/(exp(2*A)-2);
else
    B = 60*pi^2/(Zo*sqrt(Er_2));
    w_h = (2/pi)*(B-1-log(2*B-1)+(Er_2-1/(2*Er_2))*(log(B-1)+0.39-
0.61/Er_2));
end

w = w_h*h;

% Calculate effective dielectric constant at 32 GHz

Eeff_2 = ((Er_2+1)/2) + ((Er_2-1)/2)*((1+12*h/w)^(-0.5));

%Setup the slope for the spatial harmonics
% ie for N = -1 ; y = +/- (Eeff)^-0.5 * x -2*pi

% Calculate the slope of the spatial harmonics
L_2=abs(1/sqrt(Eeff_2));

% Spatial Harmonic N=0
p_2=-24*pi:2*pi:24*pi;
z_2=abs(p_2*L_2);
```



```

% Spatial Harmonic N = -1
p1_2=p_2-2*pi;
z1_2=z_2;

% Find x and y intercept between N =-1 space harmonic, the slow fast
wave
% region and the x=0 axis
% Note : These values have NOT been normalized to pi

% First for the slow wave region and space harmonic has a positive
slope
A1 = 1;
B1 = -1;
C1 = 0;

A2 = abs(1/sqrt(Eeff_2));
B2 = -1;
C2 = A2*2*pi;

y_slow_pos = ((-A1*C2) + (A2*C1)) / ((A1*B2) - (A2*B1));

% The negative slope

A1 = 1;
B1 = 1;
C1 = 0;

y_slow_neg = ((-A1*C2) + (A2*C1)) / ((A1*B2) - (A2*B1));

%The value along the y axis

y_axis = A2*2*pi;

%Calculate the spacing range d in m for f = 32 GHz

ko = 2*pi/WL;
d1 = y_slow_pos/ko
d2 = y_slow_neg/ko
d3 = y_axis/ko

%Half wavelength
WL2 = WL*(1/2);

% Calculate the angle of radiation for all remaining values of d
b = 0;

for mz = y_slow_neg:0.3:y_slow_pos
    b = b+1;
    kd(b)= mz
    Bd(b)= (mz - y_axis)/(A2)
end
kd_final = d1*ko
Bd_final = (kd_final - y_axis)/(A2)
save kd_FW.out kd -ASCII
save Bd.out Bd -ASCII

```

A.5 Complex Permittivity and Plasma Depth

```
%-----
% This program calculates the complex permittivity, loss tangent
% and plasma depth of a piece of illuminated silicon.
%-----

%Calculate Real Component of Permittivity
El = 11.7;
E0 = 8.85*10^-12; % Free Space Permittivity in F/m
m0 = 9.11*10^-31; % Electronic charge in kg
mn = 0.33*m0; % In kg
w = 2*pi*(31*10^9); % In rads
q = 1.6*10^-19; %Charge in coulombs
ue = 0.15; % Mobility in m2/V
Te = (ue*mn)/q;
f=31*10^9; % Frequency in GHz

n = 10^21; %Carrier Conc. in m-3
r = 0;

w= 2*pi*f;
Ereal = El - ((n*(q^2)/(E0*mn))/((w^2)+(1/Te)^2))
EImag = ((1/Te)/w)*((n*(q^2)/(E0*mn))/((w^2)+(1/Te)^2))
LT = EImag/Ereal

n = 10^22; %Carrier Conc. in m-3
r = 0;

Ereal = El - ((n*(q^2)/(E0*mn))/((w^2)+(1/Te)^2))
EImag = ((1/Te)/w)*((n*(q^2)/(E0*mn))/((w^2)+(1/Te)^2))
LT = EImag/Ereal

n = 10^23; %Carrier Conc. in m-3
r = 0;

Ereal= El - ((n*(q^2)/(E0*mn))/((w^2)+(1/Te)^2))
EImag= ((1/Te)/w)*((n*(q^2)/(E0*mn))/((w^2)+(1/Te)^2))
LT = EImag/Ereal

sigma = 400*10^4; % Radidation absorption coefficient m-1 at 900 nm
un = 0.15; % Electron mobility in m2/v
up = 0.06; % Hole mobility in m2/v
T = 1*10^-6; % Excess carrier lifetime in s
K = 300; % Temperature in Kelvin
kb = 1.38*10^-23; % Boltzmann's constant in J/K
q = 1.6*10^-19; % Electronic charge in Coulombs
vs = 1; % Surface Recombination velocity in m/s
```

```

% Calculate the diffusion length in m

La = (((2*un*up*T*K*kb)/(q))*((un+up)^(-1)))^(1/2);
% Plasma Depth

de =
(1/sigma)*((La*(1+sigma*La)+(vs*T))/(La+vs*T))*((1/(sigma*La))*((sig
ma*(La^2)+vs*T)/(La+vs*T)))^(sigma*La/(1-sigma*La))

deum = de*10^6

```

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