HOLOGRAPHIC-BASED LEAKY WAVE ANTENNAS AND PHONON-ENHANCED WAVEGUIDES FOR INFRARED APPLICATIONS

BY

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“Holographic-based leaky wave antennas and phonon-enhanced waveguides for infrared applications,”

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ABSTRACT

Title: Holographic-based leaky wave antennas and phonon-enhanced waveguides for infrared applications

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This work contributes to critical requirements for antenna-coupled detection applications: 1) large capture cross section with design control of angular response and 2) long propagation lengths of complex signals. Antenna-coupled detectors for infrared applications provide appreciable capture cross sections while delivering signals to sub-wavelength-sized detectors. In this work, leaky-wave antennas are considered for the first time in the infrared spectrum. Design characterization of angular radiation patterns as well as polarization is presented. While leaky-wave antennas hold great promise for enhancing capture cross section, thus narrowing the field of view of the detector, there is a parallel need to preserve coherent signals over long propagation lengths. Infrared electromagnetic fields can couple to phonon-polaritons, the collective oscillation of lattice charges, much like plasmonic coupling that is prominent in the visible spectrum. Hybrid plasmonic optical waveguides have been shown to provide an excellent tradeoff in propagation length and modal confinement. Here we integrate hybrid waveguide designs with silicon carbide, a polar material with a phonon resonance in the long-wave infrared, such
that phonon-coupled enhancement, analogous to visible plasmonic effects, are achieved in the infrared.
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<tr>
<td>$E$</td>
<td>Electric field intensity [v/m]</td>
</tr>
<tr>
<td>$\varepsilon_r$</td>
<td>Relative dielectric constant of substrate [-]</td>
</tr>
<tr>
<td>$\lambda_o$</td>
<td>Free space wavelength [M]</td>
</tr>
<tr>
<td>$R_r$</td>
<td>Radiation resistance [Ω]</td>
</tr>
<tr>
<td>$\delta_{\text{eff}}$</td>
<td>Effective loss tangent [-]</td>
</tr>
<tr>
<td>$P_d$</td>
<td>Dielectric loss [W]</td>
</tr>
<tr>
<td>$\tan \delta$</td>
<td>Loss tangent of the dielectric [-]</td>
</tr>
<tr>
<td>$B$</td>
<td>Magnetic flux density [wb/m$^2$]</td>
</tr>
<tr>
<td>$h$</td>
<td>Height of the substrate [M]</td>
</tr>
<tr>
<td>$Q_r$</td>
<td>Represents the quality factor for radiation [-]</td>
</tr>
<tr>
<td>$j$</td>
<td>Induced current [A]</td>
</tr>
<tr>
<td>$j_s$</td>
<td>Surface current [A]</td>
</tr>
<tr>
<td>$D$</td>
<td>Electric flux density [C/m$^2$]</td>
</tr>
<tr>
<td>$H$</td>
<td>Magnetic field intensity [A/m]</td>
</tr>
<tr>
<td>$\beta$</td>
<td>Propagation constant [rad/m]</td>
</tr>
<tr>
<td>$n_d$</td>
<td>Refractive index</td>
</tr>
<tr>
<td>$Z_o$</td>
<td>Characteristic impedance Ohms[Ω]</td>
</tr>
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Abbreviations

- IR  Infrared
- CAD  Computer aided design
- CPU  Central processing unit
- FDTD  Finite difference time domain
- FEM  Finite element method
- MoM  Methods of moments
- RF  Radio frequency
- TEM  Transverse electric-magnetic
- TM\textsubscript{mn}  Transverse magnetic in m by n dimensions
- TE\textsubscript{mn}  Transverse electric in m by n dimensions
- UHF  Ultra high frequency
- IMI  Insulator metal insulator
- MIM  Metal insulator metal
- SPP  Surface plasmon polaritons
- SW  Surface wave
- WG  Waveguide
- RA  Radiating aperture
- LWIR  Long wave infrared
- SPhPs  Surface phonon polaritons
- SMRS  Sinusoidally modulated reactance surface
- **WPP**  Wedge plasmon polariton
- **PBC**  Periodic boundary condition
- **CP**  Circular polarization
- **LP**  Linear polarization
- **LWA**  Leaky wave antenna
- **FS**  Free space
- **TL**  Transmission line
- **CRLH**  Composite right left hand
This page is dedicated to the memory of my mum, Nancy, who was always impressed by her son’s achievement. Mum, you were the best in the world. If all mothers were like you, every day would burst with sunshine. May you rest in eternal peace.
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Chapter 1

1.0 INTRODUCTION

This dissertation is mainly divided into two but related topics. What ties the two topics together is a common point of view of their design, which is based on the travelling wave that propagates along an antenna aperture as a surface wave. The first part is on plasmon/phonon wave guiding at infrared (IR). Hybrid model waveguide configuration is simulated based on our new proposal. The second part is on infrared (IR) metasurfaces that have directional wave absorption and radiation. Here, linear and circular polarized metasurfaces are designed and characterized at infrared range of frequency. This chapter is separated into two sections, which give an introductory overview of the topics pursued in this dissertation.

1.1 Motivation on Plasmon/ Phonon Waveguiding

Although surface plasmon polariton (SPP) waveguides have been extensively studied in the visible and near-IR, there has been much less development in the long-wave IR (LWIR) and terahertz (THz) frequency range. While SPPs are still supported at these frequencies, they are very poorly bound to the metal surfaces (low field confinement). Several studies have considered traditional microstrip and two-wire transmission line structures; however, metal
losses lead to unacceptably short propagation lengths [1]. Another possibility is to heavily dope silicon, controlling the plasma frequency so that SPPs are supported at LWIR [2]. However, this is still accompanied by low propagation lengths.

A new approach is to use Surface Phonon Polaritons (SPhP), which arise due to the coupling of the electromagnetic field to the lattice vibrations of polar dielectrics. This occurs at mid- and long-wave IR frequencies. SPhPs propagate only as surface waves close to the interface between two different media [3,4]. Surface phonon polaritons are excited using IR or terahertz radiation and can be generated within a wide spectral range of 8-200 μm wavelengths and at the surface of a large variety of semiconductors, insulators, and ferroelectric materials [4]. This study fills a knowledge gap of surface waveguiding at the LWIR, which has remained a gray zone since plasmonic waveguiding, which has matured in optical frequencies, is limited at this range. This work is covered in Chapter 3.

1.1.2 Literature review on plasmonics and phonon polaritons

The optical properties of materials change drastically as the size of particle is reduced. In the nanoscale range, the optical properties of metallic particles are dramatically modified by the appearance of surface plasmon polaritons (SPP) [5,6]. Metals, such as silver and gold, contain free electrons that collectively oscillate and propagate as the surface wave in the optical frequency region [7]. SPPs are formed when free electrons simultaneously oscillate with the applied electric fields and couple to form electromagnetic excitations propagating at the interface between a
dielectric and a conductor and are evanescently confined within a sub-wavelength region near the interface in the perpendicular direction [5,8,9,10,11]. Fig. 1.1 shows SPPs formed at an interface of metal and air dielectric. This SPP mode is capable of waveguiding but, due to material loss of the waveguide, the propagation length shrinks as the confinement of the mode increases corresponding to a trade-off between confinement of a waveguide mode and its propagation distance [5, 8,10, 12].

SPPs are TM polarized, surface bound in nature, and cannot be excited by light beams directly because of phase mismatch. This mismatch is due to the fact that a bound mode has a higher phase velocity than unbound mode, i.e., \( \beta > k_o \) where \( \beta \) is the phase velocity of the bound mode and \( k_o \) the free space wave vector.
Ways of overcoming this mismatch include use of grooves on the metal surface or prism coupling where Kretschmann and Otto geometries have been employed [11].

The plasmonics waveguide has been considered in the visible and near infrared spectrum, while metallic waveguides are utilized in the microwave range [8]. Conventional waveguides utilizing light propagation on macroscopic scales are reliant on index contrasted media for signal storage and transmission. However, when device sizes are scaled to theoretical limits, conversion of the photon mode into a surface plasmon mode becomes a viable mechanism for subwavelength scale propagation [5]. This is because in nanostructures, surface effects are more important than volumetric effects due to a high surface area to volume ratio. This suggests that surface polaritons may play an important role in energy transport along films with nanoscale thickness because plasmonic waveguides can provide sub-wavelength confinement by storing optical energy in electron oscillations within the metal-dielectric interface [14,15]. In contrast to dielectric waveguides, plasmonic waveguides can simultaneously carry optical and electrical signals, giving rise to new capabilities [12].

For a single dielectric-metal interface, confinement is achieved since the propagation constant is greater than the wave vector in the dielectric, leading to fast fading on both sides of the interface, as shown in Fig. 1.1. The confined SPP modes of a single metal-dielectric interface can propagate over several microns under optical illumination despite the fact that the metals are characteristically lossy [5,15]. The loss in metals is due to the fact that a substantial part of light energy is
confined to the metal film. Surface plasmon dispersion and propagation are governed by the real and imaginary components respectively of the in-plane wave vector. The imaginary part of the propagation constant represents the attenuation suffered by the travelling SPP waves [5]. The propagation distance is determined by the imaginary part of the propagation constant and has an inverse relation. It is high for low permittivity dielectric with a small imaginary propagation constant [12].

Metal strips, such as in Fig. 1.2, possess long range plasmons. Nanoscale confinement of light is possible with such plasmonic waveguides, but they suffer from limited propagation distance due to the high Ohmic loss/absorption in metal. This dissipation limits the efficiency with which light can be concentrated to nanoscale volumes using SPPs in metal strips [16]. The loss increases as confinement of an SPP mode increases because a larger fraction of the modal field is located inside the metal and ohmic dissipation causes the SPPs to decay with a characteristic propagation length.

![Fig. 1.2 Metal strip in air](image-url)
Decreasing the dimensions of the metal strip in Fig. 1.2 can effectively increase the propagation distance of long range SPP, but at the cost of reduction in the overall mode confinement [10,17,18,19]. Metallic strip waveguides with a width $w$ to thickness $t$ ratio $>>1$ have shown promising results in electromagnetic wave guiding at visible and near infrared frequencies [16]. Embedding a metal strip into a dielectric medium with high permittivity not only leads to high confinement but also leads to high propagation losses [18,19]. Interfacing a high-permittivity dielectric semiconductor with a metal strip can lead to high confinement, but it has high optical loss [14]. A metal strip embedded in a homogenous dielectric environment has a significantly higher propagation distance and lower propagation loss than a regular SPP based on metal-dielectric when similar materials are considered [17].

In order to increase the propagation distance without losing confinement, various design principles have been put forward. Studies on the guidance of plasmonic modes have demonstrated subwavelength plasmonic waveguide capability in cylindrical, coaxial, triangular metal edges and the V-channel guides [17,19,20]. Other configurations that have been studied include the dielectric loaded waveguide shown in Fig. 1.3, the two-wire waveguide of Fig. 1.4, the wedge configuration in Fig. 1.5, and the hybrid model shown in Fig. 1.6.
In general, all these plasmonic waveguides have a trade-off between the mode confinement and the propagation length. The dielectric loaded waveguides of Fig. 1.3 have been presented in [17,21]. Here, we have SPP propagating at the dielectric-metal interface. Although it is easier to build, it has low confinement and a relatively large mode area. The two-wire waveguides in Fig. 1.4 have field concentration in the air gap between the two metals. The coupling strength is greatly influenced by the separation distance of the two conductors [17]. Wedge plasmon polaritons (WPP) arise from the coupling of surface plasma oscillations across the metal wedge shown in Fig. 1.5. The tip geometry strongly influences the property of the mode. Blunt tips and large side wall angles lead to a loss in confinement and mode cutoff while sharp tip and narrow side wall angle enhances confinement [20]. WPP has excellent confinement but is very lossy. The dielectric
loaded guide is a fair compromise between strip and wedge/channel, but it is still not good enough [17,20]. In the hybrid optical waveguide structure shown in Fig. 1.6, confinement of the optical field is strong in the gap region between metals and the dielectric material [22]. High optical loss results when high permittivity dielectric materials are used in the gap [14].

In the previous studies, a hybrid plasmonic waveguide has been shown to yield superior performance to other structures as reported in [9,14,17,20,22]. In the hybrid configuration, the existence of a low-index gap leads to high confinement and moderate losses. By tuning the geometrical properties of a hybrid structure, the propagation distance can be enhanced to the millimeter range while maintaining high confinement [14]. High confinement has been achieved with a system composed of a high index dielectric cylinder separated from a metal surface by a narrow low-index gap [9,12,20], with electromagnetic energy being stored and transmitted in the low permittivity region [12,14]. The resultant hybrid plasmonic TM mode can be approximated as a TEM mode, as has been shown in [23]. Therefore, the transmission line model, applicable in TEM, is also applicable to hybrid plasmonic waveguides. With TEM, both electric and magnetic fields have no component in the direction of propagation.

Confining terahertz (THz) radiation within waveguide structures offers potential advantages in size performance and versatility and integration of THz systems thus avoiding the difficulties associated with terahertz beam shaping and beam steering optics [15]. Studies have also shown that regardless of the
waveguide shape, a fundamental lower bound on the attenuation constant exists for a given mode size; this limit is as a result of the dissipative loss of the waveguide core material. SPP waveguides can provide subwavelength confinement even beyond the diffraction limit [11]. The diffraction limit dictates that light cannot be confined in a space smaller than approximately $\frac{\lambda}{2n}$ where $\lambda$ is the light wavelength and $n$ is the refractive index of the material. This means that all of the waveguides must be built at micrometer scales or larger [12, 24] for visible range application. Use of materials with a negative dielectric permittivity overcomes diffraction limitation because interfaces between polaritonic ($\epsilon < 0$) and dielectric ($\epsilon > 0$) materials support surface polaritons that can be confined to subwavelength dimensions [25]. The field, which is generated by the SPP with maxima occurring at the surface, as well as significant field penetration into the surrounding dielectric [5,9] can be enhanced and strongly confined spatially in the near field (with the distance less than the wavelength) from the metal surface as a non-irradiative evanescent field. The ability to confine and enhance the optical field to the vicinity of the metal surface or nano-metal particle has been applied to design applications such as immunosensor, fluorescence sensor, solar cell, plasmonic laser, and super lens [7].

Although plasmonic waveguides have been extensively studied, previous works have barely mentioned the transition of plasmonic modes between near infrared and microwaves [8]. At lower frequencies (mid-infrared range), the plasmonic effect is too weak and the surface phonon polaritons (SPhPs) become
dominant. SPhPs are collective excitations coupled to an electromagnetic wave and can propagate on a layer of different materials. SPhPs arise due to the coupling of the electromagnetic field to the lattice vibrations of polar dielectrics at mid-IR frequencies [4,26,27,28]. The behavior of SPhPs is analogous to both localized and propagating SPPs. Just like SPP, SPhPs can only propagate as surface waves close to the interface between two different media, and the electric field is strongly coupled to lattice vibrations with amplitude of the fields decaying exponentially with the distance from the interface [4,26,27,28]. Limited study has been done on surface phonon polaritons despite the fact that they can be generated within a wide spectral range of 8-200μm wavelength and at the surface of a large variety of semiconductors, insulators and ferroelectric materials [3,4,28].

SPhPs are excited using infrared (IR) or terahertz radiation, and optical energy can be coupled into SPhP modes at discontinuities in the materials [27,28]. The coupling of the mechanical vibrations of the lattice and electromagnetic radiation leads to the formation of mixed modes, which are called phonon polaritons. In ionic crystals, electromagnetic waves are strongly coupled to polar lattice vibrations near optical phonon resonances [25,29].

From the literature review, it is apparent that most studies have concentrated on plasmonics, but since phonon polariton is analogous to plasmonics, most of the theory from plasmonics can be applied to phonon polaritons without loss of generality despite the fact that the concept of boundary conditions and
coupling leading to their excitation is different. A contribution of this dissertation is to close this gap through the design and analysis of SPhP-coupled IR waveguides.

1.1.3 Surface plasmons theory

Surface plasmons on metals are TM guided waves and are made possible by free electrons oscillating in the propagating field just below their plasma frequency. They propagate on the surface of the metal and enable the routing and manipulation of light at the nanoscale. The optical properties of metals exhibit a negative real part of permittivity at optical frequencies and this supports the SPP mode bound to the surface [30]. SPP waves exist only at interfaces between materials with opposite signs of the real part of their dielectric permittivity, i.e., between a conductor and an insulator, as shown in Fig. 1.7. The inequality $\varepsilon_m < -\varepsilon_d$ is the plasmon condition for a single surface to support a guided wave. Plasmons are described by Maxwell’s equations, and their dispersion relationship is found by matching boundary conditions.

![Fig. 1.7 Metal dielectric interface geometry.](image)
From Fig. 1.7, plane \( Z = 0 \) coincides with the interface. A TM mode propagating along the \( x \) direction satisfies the one dimensional wave equation such that

\[
\frac{\partial^2}{\partial z^2} H_y + (k_o^2 \varepsilon - \beta^2) H_y = 0
\]  

(1.1)

where \( k_o \) is the free space wavenumber, and \( \beta \) is the propagation constant.

The solution of 1.1 is given by

\[
H_y(z) = A_2 e^{i\beta x} e^{-k_2 z} \text{ for } z > 0
\]

(1.2)

and

\[
H_y(z) = A_2 e^{i\beta x} e^{k_1 z} \text{ for } z < 0
\]

(1.3)

where

\[
k_1^2 = \beta^2 - k_o^2 \varepsilon_m \text{ and } k_2^2 = \beta^2 - k_o^2 \varepsilon_d
\]

(1.4)

However, boundary conditions require that

\[
\frac{k_2}{k_1} = - \frac{\varepsilon_d}{\varepsilon_m}
\]

(1.5)

Substituting \( k_1 \) and \( k_2 \) from Equations 1.4 to 1.5 leads to

\[
\beta = k_o \left( \frac{\varepsilon_d \varepsilon_m}{\varepsilon_m + \varepsilon_d} \right)^{1/2}
\]

(1.6)

The propagation constant (\( \beta \)) is a complex number that describes a mode’s effective wavelength and attenuation as it propagates. The dispersion relation in Equation 1.6
is modified for SPP in lossy conductors of complex permittivity ($\varepsilon_m = \varepsilon'_m + \varepsilon''_m$) as

\[
\beta = k_o \left( \frac{\varepsilon_d \varepsilon'_m}{\varepsilon_m + \varepsilon_d} \right)^{1/2} + jk_o \left( \frac{\varepsilon_d \varepsilon'_m}{\varepsilon_m + \varepsilon_d} \right)^{3/2} \frac{\varepsilon''_m}{2(\varepsilon_m)^2}
\] (1.7)

where $\varepsilon'_m$ is the real permittivity and $\varepsilon''_m$ imaginary permittivity of the metal.

Equation 1.6 can be rewritten as

\[
k_x = \frac{\omega}{c} \left( \frac{\varepsilon_d \varepsilon'_m}{\varepsilon_m + \varepsilon_d} \right)^{1/2}
\] (1.8)

where $k_x$ is the propagation constant of the travelling wave in the direction of propagation. This equation is known as the dispersion relation for SPP, or simply plasmonic dispersion. Metals (Ag, Au, Cu, Al) have negative real permittivity just below their plasma resonance frequency ($\omega_p$). Fig. 1.8 shows the variation of dielectric and metal permittivity with frequency, where $\omega_p$ is the plasma frequency (natural free oscillation of the electron sea) and $\omega_{sp}$ is the plasmon frequency.

It can be noted that the dielectric permittivity constant is relatively constant with frequency while the permittivity of metal is strongly dependent on frequency.
Two cases are considered to determine the behaviour of $k_x$. At a point when $\varepsilon_m = -\varepsilon_d$, which happens at $\omega_{sp}$, the solution is that $k_x \to \infty$ (approaches infinity). When low frequencies ($\omega$) are considered such that $\varepsilon_m \to -\infty$, then we have

$$k_x \cong \frac{\omega}{c} \varepsilon_d^{1/2}$$

Both solutions lie below the light line as shown in Fig. 1.9.

![Fig. 1.9 Confined mode below a light line](image)

In terms of propagation, all guided modes occur for solutions lying below the light line, and unguided modes are located above the light line. Therefore, SPP is a guided mode since its dispersion curve is below the light line.

While considering plasmonic modes, coupled mode theory describes the mode hybridization of two uncoupled modes. The modes are characterized in terms of an effective index proportional to the real part of its eigenvalue, quantifying the phase velocity in the direction of propagation $[9,14]$. The effective index is calculated as

$$n_{spp} = \sqrt[2]{\frac{\varepsilon_m \varepsilon_d}{(\varepsilon_m + \varepsilon_d)}}$$
The magnitude of the in-plane surface plasmon wave vector is given by

\[ k_{spp} = n_{spp} k_o \]  

(1.11)

where \( k_o \) is the free space wave number. For TM polarized waves the confined mode exists when

\[ \varepsilon_m + \varepsilon_d < 0 \]  

(1.12)

In addition, the SPP modal index is higher than the refractive index of the dielectric, i.e.,

\[ \text{Re}(n_{spp}) > n_d = \sqrt{\varepsilon_d} \]  

(1.13)

In considering the confinement of a mode in a waveguide, one of the methods that has been used is plotting of normalized intensity [22]. This method gives power per unit area plot, hence shows regions of power concentration. While the method is applicable in optical fiber theory, it is inconsistent when applied to subwavelength confinement because of rapid variations in the shape of the mode [20]. In the dissertation, calculation of the mode area has been adopted. The modal area (\( A_m \)) is defined as the ratio of the total modal energy to the peak energy density along the propagation direction [14,18,20].

\[ A_1 = \frac{\text{Modes total energy density per unit length}}{\text{peak energy density}} \]

\[ A_1 = \frac{1}{\max(W(r))} \int_{-\infty}^{\infty} W(r) dA \]  

(1.14)

where \( W(r) \) is the energy density given by
\[ W(r) = \frac{1}{2} \text{Re}\left\{ \frac{d(\omega \varepsilon(\omega))}{d\omega} \right\} |E(r)|^2 + \frac{1}{2} \mu_0 |H(r)|^2 \]  

(1.15)

E(r) and H(r) are the electric and magnetic fields, \( \varepsilon(\omega) \) is the electric permittivity, and \( \mu_0 \) is the magnetic permeability. The energy density is derived from Poynting’s theorem in a linear, lossy dispersive medium \[5\]. The equation for the modal area \( A_1 \) is not sensitive to a strong concentration of mode field in small distances, which is typical for SPP waveguides \[20\]. Confinement can be gauged by the normalized mode area, which is defined as

\[ \frac{A_1}{A_o} \]  

(1.16)

where \( A_o = \left( \frac{\lambda_0}{2} \right)^2 \) and \( \lambda_0 \) is the free space wavelength.

The propagation distance \( L_m \) is defined as the distance a mode travels before decaying to \( e^{-1} \) of its original power \[14, 19, 20, 31\].

\[ L_m = \frac{1}{(2 \text{Im}\{\beta\})} \]  

(1.17)

Equation 1.17 can be rewritten as \[32\]

\[ L_m = \frac{c}{\omega} \left( \frac{\varepsilon_m' + \varepsilon_d}{\varepsilon_m' \varepsilon_d} \right)^{\frac{3}{2}} \left( \frac{\varepsilon_m'}{\varepsilon_m''} \right)^{\frac{3}{2}} \]  

(1.18)

where \( \varepsilon_m' \) and \( \varepsilon_m'' \) are the real and imaginary parts of the permittivity of metal.

In terms of attenuation, the propagation distance can also be written as \[22\]

\[ L_m = \frac{1}{2\alpha} \]  

(1.19)

where \( \alpha \) is the attenuation constant (Np/m).
The characteristic impedance of the plasmonic waveguide is calculated as

\[ Z_{\text{waveguide}} = \frac{Z_0}{n_{\text{spp}}} = \frac{377}{n_{\text{spp}}} \]  

(1.20)

where \( Z_0 \) is the characteristic impedance of air and \( n_{\text{spp}} \) is the effective refractive index of the plasmonic mode.

The theory on plasmonics is a well-researched area and most of its formulas are applicable to phonons without loss of generality. The dielectric function \( \varepsilon(\omega) \) of a material is a tensor that gives the relationship between the electric field \( E \) and the electric displacement \( D \). In metals, the dielectric response is dominated by the free electron plasma, which can be approximated by a Drude model [33]

\[ \varepsilon(\omega) = \varepsilon_\infty \left[ 1 - \frac{\omega_p^2}{\omega^2 + i\omega\gamma} \right] \]  

(1.21)

where \( i = \sqrt{-1} \), \( \gamma \) is damping constant assumed positive and very small, \( \omega_p^2 = \frac{ne^2}{\varepsilon_o m} \) is the plasma frequency of the material, \( n \) is the electron concentration, \( e \) charge of an electron, \( \varepsilon_o \) is the permittivity of free space, and \( m \) is the mass of an electron.

For the phonon vibrations that dominate in polar dielectrics, such as SiC and SiO\(_2\), the corresponding dielectric function is given by a Lorentz model [27]

\[ \varepsilon(\omega) = \varepsilon_\infty \left[ 1 + \frac{\omega_{\text{LO}}^2 - \omega_{\text{TO}}^2}{\omega_{\text{TO}}^2 - \omega^2 - i\omega\gamma} \right] \]  

(1.22)

where \( \omega_{\text{LO}} \) is the longitudinal optical phonon frequency and \( \omega_{\text{TO}} \) is the transverse optical phonon frequency.
Because of the large magnitude of the real part of permittivity for metals in the LWIR, SPPs are not well bound to the surface and their waveguiding functionality is limited. SPhPs perform better within this spectral range.

1.2 Holography Metasurface Antennas

This section presents a literature review on holography, surface waves and leaky wave antennas (LWAs). LWAs can be classified into uniform, quasi-uniform and periodic with linear or circular polarization. In the dissertation, we have dwelled on planar rectangular type of array and the attention has focused on completely planar setup with a thin substrate loaded periodically with a printed array of patches joined with microstrip lines forming a holographic metasurface antenna. Metasurfaces are 2D surface versions of metamaterials. They are artificially engineered materials to achieve some unique electromagnetic properties not achievable by a single material in nature \([34,35]\). The construction of metamaterials is usually performed by periodically arranging a set of scatterers with subwavelength periodicity in a host medium to obtain some desired electromagnetic properties \([35]\). Due to the 2D configuration of the metasurface, they tend to occupy less space and are less lossy than metamaterials \([34,35]\). Any periodic two-dimensional structure with a small thickness and periodicity compared to a wavelength in the surrounding media can be referred to as a metasurface. Metasurfaces have been reviewed in \([34]\) where many examples of metasurfaces have been presented. What will serve as a metasurface in this dissertation is an
array of micro-patches, small compared to the wavelengths that are periodically arranged on a surface plane. These patches are arranged in two dimensions such that they support leaky-wave coupling. A linearly polarized leaky wave surface is covered in Chapter 4, while a circular polarized metasurface is presented in Chapter 5.

1.2.1 Holographic concept

The earliest metasurfaces were achieved via holography. Optical holography is a technique that allows the light scattered from an object to be recorded and later reconstructed. In this technique, the principle of interference is used to record and reconstruct the object image in real three dimensions. This concept of optical holography has been extended from its more common usage in the optical frequency range to the design of holographic antennas and holographic artificial impedance surfaces operating at microwave frequencies [36,37]. Just like in optical holography, microwave holography involves producing an interference pattern using two waves and then using the interference pattern and one wave to produce the other [38]. This interference pattern, referred to as a hologram, is a recording of the intensity of the interference pattern of two wave fronts on a surface. After a hologram has been fabricated, one of the interfering wave fronts can be reconstructed by illuminating the hologram with the other interfering wave front, provided that precautions have been taken to prevent a direct wave from the feeder interfering with the reconstructed wave [39].
Though theories have been advanced on the holography concept, practical implementations of the holographic antennas had been done only after the introduction of the artificial impedance/magnetic surfaces [40]. Artificial impedance surfaces include pin-bed structures, high impedance surfaces and left-handed meta-materials [36]. The merits of the artificial impedance are that phase and magnitude of the radiation can be defined independently, and also control of polarization of the radiated field by use of tensor artificial impedance surface is possible [36, 40, 41].

The microwave holograms have been built as a collection of scatterers that correspond to the interference pattern [36]. Reported methods of implementing the hologram include the lithographic method, which has utilized etching. In this method, the antenna has been designed as an approximation to a hologram by etching metallic strips spaced periodically on a dielectric substrate. This form of implementation is common for holographic antennas operating in the millimeter wave frequency range [39]. Modulated surface impedance has also been reported as a way of implementing the hologram. A sinusoidally modulated reactance surface refers to a surface whose modal surface impedance is modulated sinusoidally. The modal surface impedance is defined as the ratio of the tangential electric field to the tangential magnetic field of the surface wave guided by the reactance surface [42]. By modulating the wave impedance on a surface, the energy in a surface wave traveling on that surface can be radiated away in a desired direction and with the required intensity at a prescribed rate and angle [36, 40,42]. Currents from the
source antenna are scattered by the modulated surface impedance to produce the desired radiation pattern [36, 37, 43]. By varying the modulation depth along the antenna, tapering of the aperture magnitude is possible and can be used to control the side lobe levels of the radiation pattern [42]. The required impedance distribution for a desired radiation pattern can be realized with artificial impedance structures [44]. For square patched artificial surface on a grounded substrate, the effective surface impedance depends on the size of the patches and the gap between the metal patches. Both can be varied as a function of position to realize the proper surface impedance distribution in order to achieve the desired radiation/scattering characteristics [36,42,44]. High impedance values are obtained with a narrower gap, a higher dielectric constant and a thicker substrate [36]. Other variables, such as the average surface impedance and periodicity of the sinusoidally modulated impedance, control beam direction, while the modulation depth controls the attenuation constant, with larger values of attenuation resulting from higher depths of modulation [42].

With holography, it is possible to create antennas with novel properties such as radiation toward angles that would otherwise be shadowed [41]. The extension of holography to the microwave range has resulted in significant improvement of a conventional antenna’s radiation characteristics, reduction of antenna size, the enhancement of antenna directivity, and the simplification of antenna feed structure [43]. Redirection of a plane wave around a solid object with the holographic artificial impedance approach is also possible. In this case, the impedance patterns
are designed to capture the plane wave energy incident on one side of a solid cylinder into a surface wave that traveled around the cylinder and then re-radiated on the opposite side of the cylinder at a prescribed angle [44]. Holographic antennas have a low profile and are lightweight. Research in the area of holographic antennas has validated their use as high gain antennas, which may serve as possible alternatives to microstrip phased array, reflector, and lens technologies [37]. A holographic-antenna-inspired structure in [45] is used to control the surface wave (SW) excited by a microstrip patch antenna. Here, the hologram is designed to support a periodic leaky-wave, which radiates at broadside and enhances the radiation of the patch while suppressing the horizontal lobe.

The efficiency of the holographic antenna has been formulated by quantifying the effects of various losses in the system. These losses are called the spillover efficiency, taper efficiency, and the termination efficiency [37]. The spillover efficiency accounts for losses incurred when portions of the feed power are radiated away from the surface and thus do not contribute to exciting the antenna [37]. The taper efficiency represents losses sustained when the feed does not illuminate the holographic surface uniformly. Any tapering of the amplitude along the width of the holographic strips results in a decrease in holographic antenna gain [37]. The termination efficiency corresponds to losses associated with the residual power that has not been scattered by the holographic antenna [37].

Planar holographic antennas are electrically large and dielectrically inhomogeneous and this has been a bottleneck for detailed electromagnetic
modeling and numerical optimization [39]. The holographic antenna concept has shown that it is viable for leaky mode applications; holographic-surface antennas [41] can be interpreted as tapered 2D LWAs, where a local modulation of the surface impedance leads to a transformation from a guided surface wave to a radiative leaky wave. The connection between the holographic pattern and the leaky wave mechanism was clarified in [46]. Planar leaky antennas leak power from the travelling waves propagating along the antenna surface. Beam shaping of a leaky wave antenna is through independent control of phase and leakage constant along the antenna [42]. From the literature review, the author found that, despite a mature field of holography and impedance surfaces, applications of holography and the leaky wave concept in the infrared range have not been considered yet to the best of the his knowledge. This is a gap that the current research intends to bridge.

1.2.2 Surface waves

Surface waves are waves that are bound to a surface and propagate parallel to the interface and decay exponentially in the perpendicular direction [47]. They form at the interface between two materials (dielectric, metal, free space, etc.). Surface waves appear in many situations involving antennas as ground waves and were first examined by Sommerfield [48]. Advancement on his research was done by Goubau in 1950 [49], and application of surface waves to waveguiding in planar configurations was done by Attwood in 1951 [50].
Surface wave antennas belong to a class of travelling wave antennas. The performance of a travelling wave antenna is limited in the sense that, for high gains, the surface wave becomes loosely bound to the guiding surface and the antenna performance becomes sensitive to the irregularities in the structure. Modulating the surface impedance has an effect of confining the guided wave if the modulating spacing is small and leads to a tightly bound surface wave [51]. If the spacing of the modulation is larger than some critical value, the modulated surface wave gives rise to one or more leaky waves, which will radiate away from the surface at an angle and, depending on the level of modulation, some stop bands may be introduced [51]. The propagation and radiation of leaky waves are controlled by the magnitude, modulation depth, and period of the surface impedance [36, 51]. Changing the phase or the magnitude of modulation leads to a radiated beam that is tilted at an arbitrary angle with the surface [51]. The problems associated with surface waves include multipath interference and unwanted mutual coupling [52]. These problems can be eliminated by the use of high impedance surfaces. A high impedance surface can be generated on a conducting surface by incorporating some special texture on the surface. Some of these textures include bumpy surfaces, corrugated surfaces, and an abstraction of the corrugated surface as reported in [52]. In his works, Sievenpiper proposed a high impedance metallic electromagnetic structure that conducts dc currents but does not conduct ac currents within a forbidden frequency band gap. The band gap was achieved by the use of vias structures, forming a textured structure that is a photonic crystal. Although the
surface does not support bound surface waves within the band gap, it does support TE polarized leaky waves, which radiate energy into the surrounding space as they propagate [52]. A transverse electric (TE) wave on a capacitive impedance surface produces a horizontally polarized radiation [36] while a transverse magnetic (TM) wave propagating on an inductive impedance surface leads to vertically polarized radiation.

When used as a ground plane, a high-impedance surface has been reported to provide drastic improvements in the radiation characteristics of vertical monopole antennas, patch antenna, and the horizontal wire antenna; its implementation would not have been feasible without maintaining spacing between the antenna and ground plane of a quarter wavelength [52]. It has been postulated from [52] that the use of high impedance surfaces could lead to higher antenna efficiency, longer battery life for portable communication devices, and lower weight.

1.2.3 Leaky wave antennas

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 narrow bandwidth design [53]. Non-resonant antennas or travelling wave antennas do not resonate but rather leak energy along the length, thereby producing more efficient broadband structures [54]. Leaky-wave antennas (LWA), just like surface wave antennas, belong to the
class of traveling wave antennas characterized by waves that propagate along the antenna aperture [54]. Traveling wave antennas launch one or more traveling wave modes in a guiding structure and the waves do not radiate except at the discontinuity such as the termination of the structure [47, 54]. Classification of modes in the structure is based on their phase velocity. Guided modes are slow as their phase velocity is lower than the free space velocity [54].

A surface wave is an example of a slow wave. On the other hand, fast waves have their velocity greater than free space velocity and are exemplified by leaky waves. A leaky-wave antenna is a structure that allows a guided wave to leak in a well-defined manner along its length [47, 54]. Almost all the early leaky wave antennas were based on closed waveguides that were made leaky by introducing a cut along the side of the waveguide to permit the power to leak away along the length of the waveguide [42]. They were followed by already open waveguides operating in the millimeter wave range; some of them are dielectric waveguides, groove guides, microstrip lines, and microstrip patches. Nevertheless, the fundamental modes on the open waveguides are typically bound, meaning that no radiation occurs. Introducing periodicity or modifying the geometry are examples of measures done to achieve radiation [47,54].

Leaky waves radiate as they propagate and can be incorporated on conformal surfaces. These waves are modal solutions that are improper mathematically because these waves increase in the transverse direction, in contrast to bound waves, which are proper meaning that they decrease transversely [47,51].
Radiation phenomenon in LWAs, as well as in other aperture-type antennas, takes place at a fixed angle for a given frequency [47,54]. Fig. 1.10 shows a generic representation of an LWA. Unlike purely guided modes, which are characterized by a real propagation constant, a leaky mode is characterized by a complex propagation constant due to the leakage of power, consisting of both a phase constant $\beta$ and an attenuation constant $\alpha$ [54].

![Generic representation of a leaky-wave antenna](image)

Fig. 1.10 Generic representation of a leaky-wave antenna

While both the feed and the end termination pose design problems in surface wave antennas, Leaky-wave antennas are not adversely affected by the mismatch problem because the leaky aperture appears as a modification of the waveguide structure that feeds it [54,55]. In addition, the leaky mode above a leaky surface appears as a modified waveguide mode where the modification refers to different wave numbers [54]. One of the most important characteristics of a leaky wave antenna is a main beam pointing at an angle $\theta_0$ with respect to broadside. It can be scanned theoretically anywhere from forward to back end fire with little beam shape deterioration.
Due to the nature of the leaky waves being improper modes, leaky waves have a physically singular behavior, and they can describe the field on the antennas in certain ranges \([56,57,58]\). Studies have clarified that in printed antennas, the excitation of leaky waves actually enhances the gain of the antenna \([59]\).

1.2.4 Theory of leaky waves

A leaky propagating mode can be depicted as shown in Fig. 1.11.

A leaky wave field attenuates along the direction of propagation and increases exponentially in the leakage direction. Because of the leakage component, leaky waves have a complex propagation wavenumber given by \([47,54,55]\)

\[
k(\omega) = \beta(\omega) - j\alpha(\omega)
\]  

while the field component of a leaky mode propagation along the antenna can be written as
\[ F(x, y, z) = F(x, z)e^{-jk_y y} = F(x, z)e^{-j\beta_y y}e^{-\alpha_y y} \] (1.24)

where \( \beta_y \) and \( \alpha_y \) stand for the phase constant and leakage constants respectively in the y-direction. The leakage constant, also known as radiation rate, represents the loss of power along the structure as the wave propagates as shown in Fig. 1.11, and is composed of reactive rate, loss rate, and leakage rate [55,60]

\[ \alpha_y = \alpha_y^{Reac} + \alpha_y^{Losses} + \alpha_y^{Rad} \] (1.25)

Since boundary conditions require that tangential components be continuous on both sides along the boundary [54], the longitudinal wave number \( k_y \) inside the waveguide in Fig. 1.11 is equal to the longitudinal wave number \( k_y \) in the free space just above the waveguide, i.e., \( k_y^{FS} = k_y^{WG} \). The wave vector component along the \( \hat{z} \) direction is given by

\[ k_z = \sqrt{k_0^2 - k_y^2} = \sqrt{k_0^2 - (\beta_y - \alpha_y)^2} \] (1.26)

The fields outside the structure can be expressed as

\[ F(z) = F_0e^{-j(\beta_y z + \beta_y y)}e^{-\alpha_y y} \] (1.27)

Far fields can be obtained from the Fourier transform of the near fields in the radiating aperture of an antenna [54,58]. From Equation (1.26), assuming \( \beta_y \gg \alpha_y \), we have

\[ k_z = \sqrt{k_0^2 - \beta_y^2} \] (1.28)
If \( \beta_y > k_o \), then the solution of \( k_z \) from Equation (1.28) is imaginary and the mode is confined within the waveguide. If \( \beta_y < k_o \), then \( k_z \) is real and the mode radiates into free space. The radiation condition is given by

\[
\left[ \frac{\beta(\omega)}{k_0} \right] < 1
\]

The phase constant (\( \beta \)) of the leaky wave determines the angle of radiation and is given by (from broadside)

\[
\theta_0 = \sin^{-1} \left[ \frac{\beta(\omega)}{k_0} \right] = \sin^{-1} \left[ \frac{c \beta(\omega)}{\omega} \right]
\]

The equation is valid for \( \beta \gg \alpha \) or when both are comparable [54,55,61].

Another important parameter for a leaky wave antenna is the beamwidth. The attenuation constant of the leaky wave (\( \alpha \)) has a great influence on the beamwidth. For a large leakage constant, the radiating aperture is short and consequently the beamwidth is large. The LWA beamwidth can be obtained by [55, 62]

\[
\Delta \theta \approx \frac{0.91}{\left( \frac{l}{\lambda_o} \cos \theta_0 \right)}
\]

\[
\approx \frac{\alpha_y / k_o}{0.183 \cos \theta_0}
\]

The nominal design length for a leaky wave antenna radiating aperture is to have at least 90% power radiated for a given value of \( \alpha \) [63]. In this case, we have the following relation:
The radiation efficiency of a leaky-wave antenna \( (e_r) \) is defined as the power leaked into space divided by total power injected into the antenna, i.e., \( e_r = \frac{P}{P_{in}} \), and it is related to the attenuation constant as \([55,60]\)

\[
e_r = 1 - e^{-2\alpha l}
\]  

(1.34)

For a one-dimensional LWA fed at one end, the frequency at which \( \beta = \alpha \) is the cutoff frequency of the leaky mode (LM). At this point, the active and reactive power densities associated with the leaky mode in the structure are equal. For a bidirectional LWA, the \( \beta = \alpha \) condition corresponds to a splitting condition and a broadside radiation is experienced when \( \beta < \alpha \) \([55, 64]\). For uniform and quasi-uniform, the latter condition corresponds to a point of maximum power density radiated at broadside \([61]\). Above the cutoff regime, and when the radiation condition is satisfied, the modal solution of the structure corresponds to a leaky wave whose energy is leaked from the angle \( \theta_0 \) (which in principle can be scanned between 0º and 90º); this regime is also called the space wave region. The minimum value of \( \theta_0 \) is produced at the cutoff frequency, which is normally near broadside. A bounded region corresponds to the condition \( \left[ \frac{\beta(\omega)}{k_0} \right] > 1 \), and the modal solution in this case corresponds to a surface wave. The bounded regime is characterized by low leakage \( \left( \frac{\alpha}{k_0} \approx 0 \right) \) while the cutoff regime (also named reactive region) is characterized by values of \( \alpha \) greater than \( \beta \) \([61]\).
A leaky wave has an improper field due to the fact that the wave increases exponentially in the near transverse air region. The behavior of an improper mode is depicted in Fig. 1.12.

![Leaky wave field distribution](image)

Fig. 1.12 Leaky wave field distribution [60,63].

Considering Fig1.12, the field expression in the free space region is of the form

\[
F(z) = F_0 e^{(-j(\beta_z z - \alpha_z z + \beta_y y))} e^{-\alpha_y y}
\]  

(1.35)

where

\[
(\beta_z - j\alpha_z)^2 + (\beta_y - j\alpha_y)^2 = k_o^2
\]  

(1.36)

\(k_o\) is the free space wave number. Equating the imaginary parts on both sides yields

\[
\alpha_z \beta_z + \alpha_y \beta_y = 0
\]  

(1.37)

Since \(\alpha_z\) is positive, \(\alpha_y\) must be negative since both \(\beta_y\) and \(\beta_z\) are positive for outward propagation. This means that the fields increase with \(y\). Therefore, the
fields of a leaky mode do violate the radiation condition leading to an improper wave.

A leaky wave antenna can be modeled as a travelling wave source where a current travels along a guiding one-dimensional uniform (or quasi-uniform) structure, as shown in Fig. 1.13. For simplicity, this structure is considered to be short in the $x$ direction, so that no surface waves are supported in the substrate along the $x$ axis; hence no energy is leaked into that direction. It is also assumed that the structure is matched such that any energy that reaches the end of the structure is fully absorbed by the load. The scheme of an 1D rectangular radiating aperture, whose dimensions are $L$ by $W$, is shown in Fig. 1.13.

Assuming a leaky wave aperture along the $y$ axis in Fig. 1.13 with $x$-polarized electric field, the near fields can be of the form

$$E_t(x, y) = E(x)e^{-jk_yy} = E(x)M(y)e^{+j\phi(y)}$$  \hspace{1cm} (1.38)
\[ x \in \left[ -\frac{W}{2}, \frac{W}{2} \right], \quad y \in [0, L] \]

\[ E(x) = 1, \quad M(y) = e^{-\alpha y} \quad \text{and} \quad \varphi(y) = -\beta y \quad \text{from} \quad k_y = \beta_y - j\alpha_y. \]

The illumination in the aperture along the \( y \) axis depends on the leaky-wave propagation constant. By use of aperture theory in [65], the fields radiated by the aperture can be written as

\[ E_r = 0 \quad (1.39) \]

\[ E_{\theta} = jE_o W k_0 e^{-j k_o r} \frac{\sin X}{2\pi r} \cos \phi \int_0^L M(y') e^{j \varphi(y')} e^{j y' \xi \sin \phi} \, dy' \quad (1.40) \]

\[ E_{\phi} = -jE_o W k_0 e^{-j k_o r} \frac{\sin X}{2\pi r} \cos \theta \sin \phi \int_0^L M(y') e^{j \varphi(y')} e^{j y' \xi \sin \phi} \, dy' \quad (1.41) \]

where \( E_o \) is the wave amplitude, and \( \theta, \phi, r \) are spherical coordinates, \( \xi = k_0 \sin \theta \) and \( X = \frac{1}{2} k_0 \sin \theta \cos \phi \).

The 3dB pattern beamwidth in radians for a 1D unidirectional leaky wave antenna is given by [55]

\[ \Delta \theta = 2 \left( \frac{\alpha}{k_0} \right) \sec \theta_o \quad (1.42) \]

The formula holds also for a 1D bidirectional leaky wave antenna if \( |\beta| \gg \alpha \) [55]. A bidirectional leaky wave antenna is one that has guided leaking waves travelling equally in both directions from a mid-feed point. The beam radiated by a bidirectional leaky wave will always point at broadside for \( \beta \ll \alpha \) [55].

The fundamental mode of a uniform microstrip is a slow wave and does not radiate; however, it can be turned into a fast wave by periodically loading it [55].
Periodic LWAs have an infinite number of space harmonics excited \([54, 55, 61, 64]\) according to the formula

\[
\beta_n = \beta_0 + \frac{2\pi}{\lambda} n \quad n = 0, \pm 1, \pm 2, \ldots
\]  

(1.43)

Backward radiation is produced by the harmonics \(n < 0\). Single radiation from the fundamental harmonic can be obtained if

\[
\frac{\beta_{-1}}{k_0} = \frac{\beta_0}{k_0} - \frac{\lambda_0}{p} = 1 - \frac{\lambda_0}{p} < -1
\]  

(1.44)

\[
p < \frac{\lambda_0}{2}
\]

While periodic structures use a \(-1\) space harmonic to radiate from back fire to end fire as a function of frequency, the composite right left hand metamaterials (CRLH) allow their fundamental guided mode to perform the frequency scanning from a single leaky wave propagation. Typical CRLH LWAs are microstrip lines, coplanar waveguides coplanar strip lines, and Sievenpiper mushrooms \([55]\).

In this dissertation, an array of rectangular patches interconnected via microstrip lines is considered. In contrast to previously mentioned structures, the structure operates in the infrared range of frequency and acts like a quasi-uniform LWA.

### 1.3 Stokes Parameters

Stokes \([66]\) introduced Stokes parameters in 1852 as a convenient way to describe polarized waves characterized by observable power terms and not by amplitudes (and phases). They are versatile since they can describe un-polarized,
partial, and completely polarized light. For a monochromatic wave in the linear horizontal vertical \{H,V\} basis, the four stokes parameters are given by

\[
S_0 = |E_H|^2 + |E_V|^2
\]

\[
S_1 = |E_H|^2 - |E_V|^2
\]

\[
S_2 = 2|E_H|^2|E_V|^2 \cos \varphi_{HV}
\]

\[
S_3 = 2|E_H|^2|E_V|^2 \sin \varphi_{HV}
\]

where \(E_H, E_V\) refers to E field orientation in the horizontal or vertical and \(\varphi_{HV}\) refers to the tilt angle. For a completely polarized wave, the Stokes parameters are related through the equation

\[
S_0^2 = S_1^2 + S_2^2 + S_3^2
\]

The Stokes parameters are sufficient to characterize the magnitude and the relative phase, and hence the polarization of a wave. The Stokes parameter \(S_0\) is always equal to the total power (density) of the wave, \(S_1\) is equal to the power in the linear horizontal or vertical polarized components, \(S_2\) is equal to the power in the linearly polarized components at tilt angles \(\varphi = 45^0\) or \(135^0\), and \(S_3\) is equal to the power in the left-handed and right-handed circular polarized components. If any of the parameters \(S_0, S_1, S_2\) or \(S_3\) has a non-zero value, it indicates the presence of a polarized component in the plane wave [67]. Stokes parameter formulation possesses two main advantages. To start with, the four parameters are measured intensities, which are very important in optical polarimetry. Secondly, the Stokes parameters can be expressed in terms of a Wolf coherence matrix [67]. This \(2 \times 2\) matrix is defined as
\[ [J] = \langle EE^* \rangle = \begin{bmatrix} \langle E_H E_H^* \rangle & \langle E_H E_V^* \rangle \\ \langle E_V E_H^* \rangle & \langle E_V E_V^* \rangle \end{bmatrix} = \begin{bmatrix} J_{HH} & J_{HV} \\ J_{VH} & J_{VV} \end{bmatrix} = \begin{bmatrix} S_0 + S_1 & S_2 + jS_3 \\ S_2 - jS_3 & S_0 - S_1 \end{bmatrix} \]

(1.50)

where \( \langle \ldots \rangle = \lim_{T \to \infty} \left[ \frac{1}{2T} \int_{-T}^{T} \ldots dt \right] \) is the inner product or ensemble averaging of the wave.

The determinant of the matrix is given by

\[ Det([J]) = S_0^2 - S_1^2 - S_2^2 - S_3^2 \]

(1.51)

\( Det([J]) \) represents the total energy of the wave ie

\[ Det([J]) \geq 0 \text{ or } S_0^2 \geq S_1^2 + S_2^2 + S_3^2 \]

(1.52)

When \( Det([J]) = 0 \), it means that \( J_{HV}J_{VH} = J_{HH}J_{VV} \) and the correlation between \( E_H \) and \( E_V \) is maximum, and the wave is completely polarized in which case the wave possesses three degrees of freedom, amplitude, orientation and ellipticity of the polarization ellipse [67]. When \( \{[J]\} \geq 0 \), partial polarization indicating a certain degree of statistical dependence between \( E_H \) and \( E_V \).

The diagonal elements present intensities while the off diagonal elements give the complex cross correlation between \( E_H \) and \( E_V \). For \( J_{HV} = 0 \), it shows that there is no correlation between \( E_H \) and \( E_V \) existing.

1.4 Layout of the Dissertation Document

This document is a comprehensive report on the research performed at the Applied Computational Electromagnetics lab at Florida Institute of Technology (FIT). Chapter 2 presents modeling techniques applicable to this research. It includes a discussion of the Finite Element Method (FEM), discretization of
Maxwell’s equations, Eigen modes, and the theory of the Floquet port and waveport. Key components of this research are in Chapters 3 through 5. In Chapter 3, hybrid plasmon and phonon waveguides are proposed, modeled, simulated, and discussed. This part of the work was presented at the APS/URSI IEEE 2013 conference (Manene F, B. Lail and Ed Kinzel, Waveguiding of surface phonon polaritons APS/URSI, IEEE pg. 114-115, 2013). In Chapter 4, a linearly polarized leaky wave metasurface at LWIR is designed and charaterized. This part of the work was presented in the ACES2014, Transactions on Proceedings (Manene F, B. Lail, E.D Kinzel, Design of an infrared leaky wave antenna surface based on microstrip patch array, ACES, IEEE 2014). Chapter 5 dwells on design and simulation of circularly polarized leaky wave metasurfaces. This work was presented at the SPIE2014 conference (Manene F, B. Lail, E.D Kinzel, Design of a circular polarized infrared leaky wave surface based on microstrip patch Array, SPIE 2014). Chapter 6 draws conclusions and suggestions for future research.
Chapter 2

NUMERICAL MODELING

Modeling and simulation are intermediate steps between design and fabrication. The fundamental goal of modeling is to try to approximate accurately the physical characteristics of a system by constructing a model that closely approaches a physical system. Numerical modeling has been widely used in various branches of engineering. Despite electromagnetic theories being a mature field, the real environment usually presents complicated structures for which analytical solutions do not exist; hence numerical methods are employed. Development of computational numerical methods has been attributed to the development of computer technology. This is due to the fact that elegant theories have been known for decades but have not been useful to engineers for lack of appropriate numerical algorithms to produce accurate results. Most of these theories have been revisited with the development of powerful high-speed computers that have made their solutions viable. In this work, the computational numerical analysis method employed was the Finite Element Method (FEM). The chapter presents modeling techniques applicable to this research. It includes classification of numerical modeling, an introduction to FEM discretization of Maxwell’s equations, Eigen mode theory, and the theory of the Floquet port and waveport. In addition, the
Chapter presents the waveguides and metasurface setup adopted as well as the boundary conditions taken into consideration.

2.1 Classification of Numerical Modeling

FEM is a full computational electromagnetic method in the frequency domain. The classifications of various computational methods are listed in Fig. 2.1 [68].

![Classification of numerical methods](image)

**Fig. 2.1 Classification of numerical methods**

In FEM for electromagnetics, Maxwell’s equations are put in a differential form and solved in the frequency domain. Maxwell’s equations describe the field and current distributions as continuous variables in a system. To solve a structure in HFSS FEM, the algorithm shown in Fig. 2.2 is normally adopted [69].
Finite element analysis involves four critical steps. These include [70]

- Discretizing the problem region.
- Writing governing equations.
- Assembling all the elements in the solution region.
- Solving the systems of the equations.
Electromagnetic computations must discretize the system and the equations under consideration since computer resources are limited. In the FEM, the simulated domain is partitioned into triangular subdomains and the final result is obtained by consideration of the results from each subdomain. The triangular basis functions make FEM superior to other numerical methods when dealing with complex structures, and an accurate result can be expected with a sufficiently small subdomains \[70\].

### 2.2 Discretization of Maxwell’s equation by FEM

Discretization of Maxwell’s equations is actually discretization of a three-dimensional solution space. The time harmonic electromagnetic fields equation can be written as

\[
\nabla^2 E = \nabla \times M_i + j\omega \mu J_i + \frac{1}{\varepsilon} \nabla q_{ev} + j\omega \mu \sigma E - \omega^2 \mu \varepsilon E
\]  

(2.1)

For a source free system, the equation reduces to

\[
\nabla^2 E = j\omega \mu \sigma E - \omega^2 \mu \varepsilon E = \gamma^2 E
\]

(2.2)

where

\[
\gamma^2 = j\omega \mu \sigma - \omega^2 \mu \varepsilon = j\omega \mu (\sigma + j\omega \varepsilon)
\]

\[
\gamma = \alpha + j\beta = \text{Propagation constant}
\]

\[
\alpha = \text{attenuation constant (Np/m)}
\]

\[
\beta = \text{phase constant (rad/m)}.
\]

For a lossless media, \(\sigma = 0\), Equation 2.2 can be written as the inhomogeneous scalar Helmholtz equation.
\[ \nabla^2 \phi + k^2 \phi = g \]  

(2.3)

where \( \phi \) can denote \( E \) or \( H \) fields depending on the mode being solved, \( g \) is the source function, and \( k \) is the wave number \( (k^2 = \omega^2 \mu \varepsilon) \). The boundary conditions for a three layer plasmonic waveguide shown in Fig. 2.3 is given by

\[
H_x = \begin{cases} 
A e^{-k_2y} & , y > d \\
C e^{k_1y} + D e^{-k_1y} & , -d \leq y \leq d \\
B e^{k_3y} & , y < -d 
\end{cases}
\]  

(2.4)

where

\[
k_i^2 = \beta^2 - k_0^2 \varepsilon_i
\]  

(2.5)

Since the magnetic field is continues along the boundary, we have;

\[
A e^{-k_2y} = C e^{k_1y} + D e^{-k_1y}
\]  

(2.6)
\[ Be^{k_3y} = Ce^{k_1y} + De^{-k_1y} \quad (2.7) \]

Using Ampere’s law, the electric fields can be calculated as

\[ E = \frac{1}{j\omega\varepsilon} \nabla \times H \quad (2.8) \]

The solution of Equation 2.3 can be written as

\[ L\phi = g \quad (2.9) \]

where \( L \) is a linear operator, \( g \) is the source function and \( \phi \) is the field quantity to be determined [68,70]. FEM approximates the solution by a variational method to minimize the error function, hence producing a stable solution. Assuming that a suitable inner product has been defined for the problem such that

\[ I(\phi) = \langle L, \phi \rangle - 2\langle \phi, g \rangle \quad (2.10) \]

the solution to the Helmholtz equation is equivalent to minimizing the function

\[ I(\phi) = \frac{1}{2} \iint [||\nabla \phi||^2 - k^2\phi^2 + 2\phi g] \, ds \quad (2.11) \]

\( \phi \) and \( g \) can be expanded in terms of basis functions over triangular elements as

\[ \phi_e(x, y) = \sum_{i=1}^{N} \alpha_i \phi_{ei} \quad (2.12) \]

\[ g_e(x, y) = \sum_{i=1}^{N} \alpha_i g_{ei} \quad (2.13) \]

where \( \phi_{ei} \) and \( g_{ei} \) are the tangential values of \( \phi \) and \( g \) at nodal point \( i \) of element \( e \) and \( \alpha_i \) is the vector basis function while \( N \) denotes the total number of edges.
This allows the equation to be put in a matrix notation and field solved in a three-dimensional matrix array. As a direct solution of Maxwell’s equations, the FEM method is an accurate frequency domain method in EM modeling. Accuracy can be adjusted to meet certain convergence criteria of output scattering parameters so as to attain near homogenous and continuous field solution. However, this accuracy depends on the spatial resolution used in the simulation. Accuracy is achieved if fine mesh (small values of $\phi_{el}$) is used. However, fine mesh is related to computational resources required since computers can only store and calculate a limited amount of data. When sweeping over a given frequency spectrum, the highest frequency is usually selected as the corresponding mesh frequency, which results in the smallest volume mesh elements and the best accuracy available. For periodic structures, the model is normally simplified to solving one unit cell of the periodic structure.

### 2.3 Eigen Mode

Eigen modes are resonances of a structure. The Eigen mode solver finds the resonant frequencies (Eigen values) of the structure and the fields (Eigen modes) at those resonant frequencies. The basic form of an Eigen problem is given by [70]

$$([k] - \omega^2[m])[\phi] = 0$$

(2.14)

For a non-trivial solution

$$[k] - \omega^2[m] = 0, \quad [\phi] \neq 0$$

(2.15)
where $\lambda = \omega^2$

The determinant is zero only at a set of discrete Eigen values $\lambda_i$. There is an Eigen vector $[\emptyset_i]$ that satisfies Equations 2.10 and 2.11, and corresponds to each Eigen value. Therefore

$$([k] - \lambda_i[m])[\emptyset_i] = 0 \quad i = 1,2,3,... \quad (2.17)$$

Each Eigen value and Eigen vector defines a free vibration mode of the structure.

### 2.4 Modeling Hybrid Plasmonic Waveguides

The Eigen solver approach in HFSS was used. Its main advantage is its capability of solving SPP modes of transmission lines with complex geometry. Plasmonic devices have the capability of confining light in a deep subwavelength scale and overcoming the diffraction limit that pegs the waveguide to dimensions to $\lambda/2$. In the model, the waveguides are assumed to be infinitely long. The waveguides are made infinitely long by applying master slave (periodic) boundary conditions in the direction normal to propagation. The model setup with the main critical dimensions is shown in Fig. 2.3.
Fig. 2.4 Unit cell and corresponding dimensional parameters of a hybrid waveguide

The height above the model is made at least $\frac{\lambda}{2}$ to avoid interference on bound surface wave modes, and not too high so as to eliminate introduction of dense cavity modes \[36\]. This restricted computational domain still supports a wide variety of cavity modes in addition to the SPP modes of interest supported by the waveguide \[10\]. Hence, reviewing of the modes having their field localized in the low permittivity material is done to ascertain SPP in a hybrid model.

The critical dimensions are in the ranges of $d = 0.15 - 2.25\mu m$, $s = 0.25 - 2.5\mu m$ and $g = 100nm$. The length of the waveguide in the model is $l = 10nm$. It is desirable to minimize $l$ since it reduces the size of the computational domain and thereby saves on the computer resources. Master slave boundaries were applied transverse to the direction of propagation so as to simulate an infinitely long waveguide. The model had a symmetrical boundary to save on computer
resources and to reduce the size of the computational domain as much as possible, thereby decreasing the mode density. Symmetry planes enable modeling of only part of a structure. Perfect $E$ or Perfect $H$ symmetry planes are applicable on a planar surface and must be exposed to the outer surface. They reduce model complexity by eliminating part of the solution volume. For instance, Perfect $E$ forces the electric field to be perpendicular to the surface, as shown in Fig. 2.5, and it reduces model complexity by eliminating conductor loss [69].

![Perfect E surface](image1)

Fig. 2.5 E-field Perpendicular to the surface

Perfect $E$ represents metal surfaces, ground planes, and ideal cavity walls.

Perfect $H$ forces the electric field tangent to the surface as shown in Fig. 2.6. It does not exist in the real world, but represents a useful boundary constraint for electromagnetic modeling. It represents openings in metal surfaces.

![Perfect H surface](image2)

Fig. 2.6 E-field parallel to the surface

In the simulation, Perfect $H$ was used in the symmetry plane since the TM mode was the dominant mode. It has to be noted that mechanical symmetry does not equal electrical symmetry when applying the symmetrical planes. The guidelines to decide the type of symmetry to use are stated as follows [69]. If the
symmetry is such that the E-field is normal to the symmetry plane, use a perfect E symmetry plane, and if the symmetry is such that the E-field is tangential to the symmetry plane, use a perfect H symmetry plane. The remaining boundaries are of perfect electric walls placed far enough from the wire to leave the SPP modes unperturbed.

One of the critical parameters to be determined is the phase delay, which can be calculated as

\[ P = n_{\text{eff}} \frac{2\pi}{\lambda_0} l \]  \hspace{1cm} (2.18)

where \( P \) is the propagation delay and \( l \) is the distance between the two planes of periodicity. \( n_{\text{eff}} \) is at first approximated by

\[ n_{\text{eff}} = \sqrt{\frac{\varepsilon_m \varepsilon_d}{(\varepsilon_m + \varepsilon_d)}} \]  \hspace{1cm} (2.19)

This aids in the first approximation of phase delay, \( P \). Through numerical iteration, a value for \( P \) is determined at a point where the Eigen resonance of the structure closes to the target frequency. The phase delay is a critical parameter in the master-slave boundaries and is inputted at this point, as shown in Fig. 2.7
Phase delay is critical since it ensures that the electromagnetic field on one periodic boundary surface of the computational domain matches the field on the opposite face of the periodic boundary, such that [10]

\[ f(z + l) = e^{ip}f(z) \] (2.20)

where \( Z \) is the space coordinate along the direction of propagation and \( l \) is the distance between the two planes of periodicity.

In order to achieve suitable accuracy, the volume mesh elements must be small enough at the given simulation wavelength so that a near-homogeneous and continuous field solution is obtained. With HFSS, the accuracy can be adjusted to adhere to certain convergence criteria of the output scattering parameters. Adaptive meshing is applicable in HFSS and can be adjusted to critical regions. Fig. 2.8 shows a meshed structure. Very fine meshing is noted in metallic regions and other critical areas such as material boundaries. There is always a trade-off between the size of the mesh and the desired level of accuracy.
The Eigen solver in HFSS provides complex Eigen frequencies, enabling the calculation of both the propagation and attenuation constants in post-processing operations. Wave velocity in the waveguide is given by

\[ V_p = \frac{c}{n_{eff}} \]  \hspace{1cm} (2.21)

The propagation length \( L_m \) is calculated from

\[ L_m = \frac{1}{2\text{Im}(\beta)} = \frac{1}{2k} \]  \hspace{1cm} (2.22)

where \( K = \frac{2\pi}{\lambda}, \lambda = \frac{V_p}{f} \) and \( f \) is the imaginary frequency solution.

The attenuation constant is then calculated as

\[ \alpha = \frac{1}{2L_m} \]  \hspace{1cm} (2.23)

Another way to extract attenuation was using

\[ \alpha = \frac{\beta}{2Q} \]  \hspace{1cm} (2.24)
where is the Q factor from the Eigen mode solution and $\beta$ is given by

$$\beta = n_{\text{eff}} \frac{2\pi}{\lambda_0}$$  \hspace{1cm} (2.25)

Further on, attenuation can be calculated by integrating the volume loss density over the whole computational domain and dividing it with transmitted power per unit length.

$$\alpha = \frac{P_{\text{loss}}}{2P_T}$$  \hspace{1cm} (2.26)

where $P_{\text{loss}}$ is the total power loss in the computational domain and $P_T$ is the power transmitted along the waveguide per unit length. This procedure was considered the easiest and most consistent and therefore was adopted.

### 2.5 Floquet’s Theorem

Floquet modal analysis is a mathematical tool that can be used for electromagnetic analysis of periodic structures. It can be used in situations where the magnitudes of the fields are periodic, but the phases of the fields have a constant shift between periods. A periodic vector function can be written as

$$p(x, y, z + nd) = p(x, y, z)$$  \hspace{1cm} (2.27)

where $p$ is a periodic vector function of $z$ with period $d$.

Since $p(x, y, z)$ is periodic, it can be expressed in a Fourier series as

$$P(x, y, z) = \sum_{n=-\infty}^{\infty} p_n(x, y)e^{i(2\pi/n)zd}$$  \hspace{1cm} (2.28)
A time harmonic electromagnetic field $\varphi(x, y, z)$ of a normal mode guided along an axially periodic structure can be written by defining Floquet’s theorem in Bloch waveform as [54]

$$\varphi(x, y, z) = e^{ik_{zo}d}P(x, y, z)$$

(2.29)

$P(x, y, z)$ describes the local “microscopic” field within a period while $e^{ik_{zo}d}$ represents the “macroscopic” field, which represents a uniform travelling wave with an axial propagation constant $k_{zo}$. Substituting Equation 2.23 into 2.24, we have

$$\varphi(x, y, z) = \sum_{n=-\infty}^{\infty} p_n(x, y)e^{ik_{zn}z}$$

(2.30)

with

$$k_{zn} = k_{zo} + \frac{2\pi}{d} n \quad n = 0, \pm 1, \pm 2, \ldots$$

(2.31)

and

$$k_{zn} = \beta_n + i\alpha = \beta_o + \frac{2\pi}{d} n + i\alpha$$

(2.32)

$\beta_o$ and $\alpha$ represent the corresponding phase and attenuation constants. The dominant $n = 0$ harmonic often adequately describes the field. The Fourier components are traveling waves with the same angular frequency but different propagation constants. The directions of attenuation and propagation are normal to each other for an inhomogeneous plane wave in free space above a lossless
interface. With periodic loading, attenuation and phase of the guided wave can be controlled to achieve a desired aperture illumination and far-field patterns [54].

In this research, a driven model with a Floquet port is used to design, evaluate, and analyze an array of interconnected patch elements. The simulation models an infinite periodic structure by surrounding a unit cell element with master-slave boundaries. Floquet analysis is used to model a plane wave incident at an angle on a planar periodic array of interconnected microstrip patches. Each element of the structure receives the identical magnitude of the incident field and the scan angles are used to calculate the spectral characteristics of the periodic leaky metasurface as the frequency is scanned. When excited by a plane wave, each element’s response has identical magnitude, but the response of the whole periodic structure is influenced by the frequency of the incident field. An element design is acceptable when the scattering matrix meets certain criteria over both a range of incidence angles and a specified frequency bandwidth [69].

Floquet analysis allows the response of the structure to be decomposed into a set of spatially-harmonic modes. Since it is a planar array, the Floquet modes are two-dimensional. Only a finite number of modes representing propagating plane waves influence the far-field radiation pattern, while the remaining modes (of infinite number) represent fields that are evanescent along \( z \) (reactive power). Floquet modes are often decomposed into transverse-electric (TE) and transverse-magnetic (TM) modes. The computational simulations determine the magnitude and phase of these modes with respect to the magnitude and phase of the incident
wave. Leaky modes are generally TM and the frequency scanning properties are based on TM scattering parameters. This dissertation illustrates linear polarization (LP) as well as circular polarization (CP). In CP, the plane of the polarization circle is perpendicular to the direction of propagation.

Because the patches in the LWA array are periodically located and infinite in the $x$ and $y$ directions, the whole structure can be studied by investigating only one unit cell and by applying periodic boundary conditions (PBC). In the unit cell approach, the size of the computational domain is small as only a single leaky patch element is included in the simulation; consequently, the computational burden is small for FEM schemes and a dense mesh is affordable.

2.6 Waveport

Modeling of the unit cell in HFSS using waveport analysis provides quick predictions for the fundamental propagation constant. Waveports represent places in the geometry through which excitation signals enter and leave the structure. Usage of a waveport is recommended only for surfaces exposed to the background. The port solver in HFSS assumes that the waveport defined is connected to a semi-infinite long waveguide that has the same cross-section and material properties as the port. Each waveport is excited individually and each mode incident on a port contains one watt of time-averaged power. Waveports calculate characteristic impedance, a complex propagation constant, and generalized S-Parameters that provide an easier way to get frequency-dependent characteristic impedance ($Z_o$).
The port is perfectly matched at every frequency. In setting up of a wave-port, the following microstrip port-sizing guidelines shown in Fig. 2.9 were used. The height of the port is affected by the permittivity of the substrate. The higher the permittivity, the less the fields propagate in the air and the shorter the port can be made. The width of the port affects the port impedance and propagating modes. The wider the port, the greater chance that a higher frequency waveguide mode can propagate.

Waveport sizing for a microstrip is strongly dependent on the width of the microstrip and thickness of the substrate. Assuming a microstrip width $w$ and substrate dielectric height $d$, the port height $H$ should lie between $6d$ and $10d$. The dimensions tend towards the upper limit as the dielectric constant drops and more fields exist in air rather than in the substrate. Port width ($T$) should be $10w$, for microstrip profiles with $w \geq h$, $5w$, or on the order of $3d$ to $4d$, for microstrip profiles with $w < d$ [69].
$H = 6d \rightarrow 10d, \ T = 10w$ \ for \ $w \geq d$ \ or \ $5w(3d \rightarrow 4d)$ \ for \ $w < d$

2.7 Modeling of a Metasurface
Master and slave are linked and paired boundaries that are used to model a unit cell of a periodic structure to create an infinite periodic geometry. The master and slave surfaces must be of identical shapes and sizes, and this forces identical fields on master-slave surfaces but with a phase shift. The Floquet port is always linked to master-slave boundaries; hence, it excites and terminates waves propagating down the unit cell of an infinite array. It supports multiple modes and de-embedding and allows for scanning through a progressive phase shift. Just like a waveport, a Floquet port is perfectly matched at every frequency and scan angle and decomposes the fields into Floquet modes. Floquet modes are orthogonal and are plane waves propagating in different directions. These modes are TE or TM in nature. TE Floquet modes represent $\phi$ polarized plane waves while TM Floquet modes represent $\theta$ polarized plane waves. The Floquet field orientation and propagation direction are dependent on frequency, lattice dimensions, mode number, and scan angle. The proper number of modes should be defined by use of a mode calculator for accuracy. A Floquet port enables computation of generalized S-Parameters, which help in the calculation of the frequency-dependent propagation constant. The main advantages of Floquet Port are that the Floquet modes do not have the scan limitations of either radiation boundaries and provide both magnitude and phase information for transmit and reflection data.
2.8 Summary

An ideal leaky structure that has infinite periodicity does not exist in the real world. Any leaky structure used in the real world has boundaries. Periodic boundary condition (PBC) is the tool that enables the computation of an infinite array through a limited computational domain. A Floquet port is a driven modal solution in HFSS and is used to provide a wide band excitation at different angles inside the unit cell domain. Only the modes that are supported by the patch array as propagating modes propagate. Because the patches in the LWA array are periodically located and infinite in the $x$ and $y$ directions, the whole structure can be studied by investigating only one unit cell and by applying periodic boundary conditions (PBC).
Chapter 3

PLASMON AND PHONON WAVEGUIDING

This chapter deals with the analysis and simulation comparison of surface Plasmon and phonon polariton-coupled waveguides. Surface polariton (SP) is a collective excitation that happens when electromagnetic (EM) surface waves propagate along the surface of materials and enable confinement and control of EM energy at subwavelength scales. Unlike dielectric waveguides, which confine volume electromagnetic waves to an optically dense core, these surface electromagnetic waves are localized at interfaces between dielectric materials and metals or ionic solids that support charge density resonance [71]. In this chapter, we propose a hybrid topology consisting of a high index material, a low index dielectric and a metal for plasmon/phonon waveguiding. The main macroscopic difference of the noble metals from dielectric objects at optical and near infrared (NIR) frequencies is a negative real part of the electrical permittivity ($\varepsilon$) for the metals. Relying on this fact, metal–dielectric layers may support unique modes called surface plasmon polaritons (SPPs), and for the polar dielectrics, the analogous to SPPs are called surface phonon polaritons (SPhP), which may be confined in one or two dimensions and are guided by waveguides in the sub-wavelength regime [11]. Surface plasmon extends from plasma frequency...
$(Re \, \varepsilon(\omega) \approx 0)$ down to zero frequency $(Re \, \varepsilon(\omega) \approx -\infty)$, while surface phonon polariton resonance occurs only over a finite frequency range, which is roughly between the transverse optical and longitudinal phonon frequencies [29]. The SPPs and SPhPs are evanescent electromagnetic waves bound to the interface and are strongly coupled to the coherent oscillations of the free charges at the metal surface and lattice of the polar dielectric respectively [18].

Although plasmonics can be supported by interaction of high frequency EM fields with noble metal objects, the Ohmic losses are relatively high for highly confined SPPs and they cannot be guided for long distance. This warrants using hybrid plasmonic/dielectric components. [18, 19, 72]. The use of hybrid structures to achieve plasmon polariton has been investigated in recent works in [14]. In the present work, I have considered materials that can support plasmon and phonon waveguiding in the LWIR spectrum range. The chapter starts with some background on the plasmonic waveguide model and then discusses waveguide configuration and analysis for a hybrid plasmon polariton based waveguide. The second part of this chapter analyzes a model that can support phonon polariton coupled waveguide and its justification and then proceeds to analyze the effects of dimensions on the hybrid model. Propagation distance, confinement, and modal area are simulated.
3.1 Waveguide Configurations and Analyses

The waveguides considered in this study are infinitely long. This condition is met by applying periodic boundaries at the ends of the waveguides perpendicular to the direction of propagation. In order to reduce the mode’s density, the periodicity is made to be very short, about 10 nm. While this size of domain can still support a wide variety of other Eigen frequencies apart from the desired SPP/SPhP mode supported by the waveguide, the modes with fields localized on the interface region are the most important ones. The hybrid geometry considered here had a high permittivity semiconductor nanowire on top of a low permittivity dielectric that acted as a spacer between the wire and the metal ground surface as depicted in Fig. 3.1. The concept of hybrid model is that, by placing a dielectric medium of a high refractive index near a metal surface with a sub-wavelength low permittivity dielectric material acting as a spacer between the high index material and the metal, the field is relaxed to the low permittivity dielectric material and the propagation length of the guided mode is extended [14]. In other words, there is fusion of index-contrast and plasmonic-guiding mechanisms. A rectangular as well as a cylindrical configuration for the high index wire have been considered in this research.

Silicon (Si), zinc sulfide (ZnS), and gold (Au) were the materials used for the wire, spacer, and the metal surface respectively, while another model had silicon carbide (SiC), benzocyclobutene (BCB) and gold. The waveguides were operated in the long wave infrared at around $\lambda_0=10.6$ μm ($f = 28.3$ THz). The
various waveguide designs have been modeled with material characteristics at LWIR range. The Eigenmode approach presented in [10] was used to model the different waveguides in ANSYS HFSS, with material properties obtained from spectroscopic ellipsometry (IR-VASE, J.A. Woollam) [73]. Eigenmodes are the resonances of the structure and the solver finds the resonant frequency of the design as well as the fields at the resonant frequency. SPP/SPhP are Eigen modes of an interface between a dielectric and a metal due to the fact that they are solutions of Maxwell’s equations that can be formulated in the absence of an incident field [31]. Because of the symmetry of the structures, the analyzed models were reduced to half of the full structure with a perfect magnetic conductor (PMC) wall on the symmetrical plane. This was exploited to minimize the computational expense and help convergence on the TM mode [10]. The designs were optimized for propagation length and attenuation. For the rectangular wires, the solution with the Eigen solver was not trivial because of the $90^\circ$ corners that led to strong field singularities that required an extremely fine mesh. This problem was overcome by slightly rounding the corners of the rectangular wires by a radius of curvature. This slight modification does not alter the results significantly [10].

The Eigen solver in HFSS provides a complex Eigen frequency solution, enabling the calculation of both the propagation and attenuation constants. The accuracy and versatility of the method used has been tested in [10], where results in [30] based on semi-analytical methods of lines had been reproduced. In order to have a meaningful comparison with previous plasmonic waveguides, the modal
area, the propagation length, and attenuation are defined by Equations (1.14), (1.17), and (1.19), which are found in Chapter 1. Here we vary the wire thickness, \( d \), and the dielectric gap thickness, \( s \), between the dielectric wire and the metallic ground plane so as to control the propagation length, mode field distribution, modal area, and attenuation of a single hybrid at around \( \lambda_0 = 10.6 \mu m \).

### 3.1.1 Hybrid model of gold (Au), silicon (Si) and zinc sulphide (ZnS)

This model consisted of a metallic part, gold (Au -4802.5) that acted as a ground; silicon (Si 11.69) wire, which is a high index dielectric; and a zinc sulphide (ZnS 4.74), a low index dielectric that acted as a spacer at the working wavelength of 10.6\( \mu m \). The configuration is as shown in Fig. 3.1.

![Fig. 3.1 Rectangular and cylindrical SPP hybrid models](image)

The model considered has SPP capability and can be referred to as a hybrid plasmon polariton (HPP) waveguide. Plasmons were first described by Ritchie in 1957 [33,74] and correspond to an interaction between matter and the
electromagnetic field of the light. The analysis of plasmons entails solving Maxwell’s equations with appropriate boundary conditions while the optical absorption spectrum allows one to detect the excitation of SPPs [16]. A typical challenge for plasmonic waveguides lies in the contradiction between their mode field confinement and propagation loss along the axis of propagation as studies have shown that the higher the confinement, the higher the loss; the converse is also true [11,29]. The parameters that are taken into account for the optimization of the waveguide in our model include the height of the low-index spacer gap and the width/diameter of the high index dielectric.

Fig. 3.2 (a, b) shows the dependence of normalized modal area, propagation length, and attenuation on wire thickness, \( d \), and spacer thickness, \( s \).
Fig. 3.2 (a) Attenuation, propagation distance modal area vs. wire thickness 0.25μm spacer

![Graph showing attenuation and propagation distance modal area vs. wire thickness.]

Fig. 3.2 (b) Attenuation, propagation distance modal area vs. spacer thickness 0.42μm wire

From Fig. 3.2 (a), it is noted that the modal area varies inversely to wire thickness. For a very thin wire, the modal area becomes high, implying that confinement is low. It can be inferred that for a small dimension $d$, the field penetrates and spreads above the high-index wire; hence the field confinement is low and the propagation length is correspondingly increased. The field plot of Fig. 3.3 (a) corroborates this as it shows that field energy spreads over the medium leading to a large normalized modal area. A small $d$ results in an SPP-like mode with very weak localization parallel to the metal surfaces. Surface polaritons have long propagation lengths particularly on thin films [33]. For a large enough $d$, it is
noted that the field starts to concentrate in the high index material and leads to reduced field confinement and decreased propagation length. It can be concluded that propagation distance has inverse variation to wire thickness for models with thin spacers as depicted in Fig. 3.2 (a). To trade off the modal area, propagation length, and attenuation, an optimum combination of width of the wire, \(d\), and spacer gap, \(s\), is required. It can be seen that a wire thickness of 0.5\(\mu\)m and below leads to high propagation distance [Fig. 3.2 (a)]. Further decrease in the wire thickness leads to an increase of modal area and may lead to loss of confinement for very small wire thickness. A wire thickness of 0.42\(\mu\)m was taken to be an acceptable tradeoff for propagation length and attenuation, and the spacer thickness was varied to optimize the propagation length and modal area. It is noted that an increase in spacer thickness, \(s\), results in an increase in propagation length at a cost of rapid increase in modal area [Fig. 3.2 (b)]. Fig. 3.3 (b) shows field distribution at moderate wire thickness and spacer. Here, mode coupling results in a hybrid mode that features both wire and SPP characteristics with its electromagnetic energy distributed over both the dielectric wire and the adjacent metal–dielectric interface. As the gap distance increases, the hybrid mode tends to confine the light energy both in the gaps and the dielectric nanowire as depicted in Fig. 3.3 (c). For a large wire and spacer thickness, the hybrid waveguide supports a dielectric waveguide mode with electromagnetic energy confined to the high permittivity dielectric core as shown in Fig. 3.3 (d).
Fig. 3.3 Electromagnetic field distributions for a, \([d,s]=[0.15, 0.25] \, \mu m\) (b),\([d,s]=[1.125, 0.25] \, \mu m\) (c), \([d,s]=[1.5, 2.5] \, \mu m\) (d),\([d,s]=[1.875, 2.5] \, \mu m\)

In the dimensions considered, the range of the modal area lies within the strong confinement region according to Oulton criteria [14], where typical strong mode confinement lies between \(\lambda^2/400\) to \(\lambda^2/40\).

Considering a cylindrical wire, we get results as shown in Fig. 3.4.
Here mode coupling results in concentration of field between the wire and the spacer. The propagation distance is $351.987 \mu m$, with high confinement in the range of $5.55 \times 10^{-4}$. The high confinement and propagation distance can be attributed to hybrid SPP coupling.

In the hybrid structure, two modes exist, a plasmonic mode between the metal and dielectric and a dielectric mode through the dielectric wire [14]. Considering the dielectric mode through the high index wire, it has been postulated that sub-wavelength confinement along all dielectric waveguides is due to light enhancement and confinement caused by the large discontinuity of the electric field at high index contrast interfaces [72]. The guiding mechanism is based on total internal reflection (TIR) in a high index material (core) surrounded by a low index
material cladding [75]. It has also been shown in [72] that the optical field can also be enhanced and confined in a low-index material even when light is guided by TIR. Attenuation in a high permittivity dielectric waveguide is too high for long distance propagation. One of the ways of mitigating for high attenuation is by use of a fiber with a subwavelength core. This means that a small part of the wave propagates in the core of high loss while the greater part lies in the free space, but this leads to loss in confinement [29].

The hybrid model takes into consideration the synergy of dielectric waveguiding and plasmonics to extend the sub-wavelength confinement capabilities with high propagation lengths. This has been put forward by coupled mode theory, which has been advanced in [14] to explain the working of a hybrid model. The low dielectric spacer enables a capacitor like energy storage and provides a means to store electromagnetic energy leading to sub-wavelength optical guiding with low mode loss. Another plausible theory is that the dielectric discontinuities at the silicon and zinc sulphide interface produce a polarization charge that interacts with the plasma oscillations of the metal-zinc sulphide interface, hence causing the gap region to have an effective optical capacitance [14,76]. In addition, due to the presence of the dielectric, the field is spread over the low loss dielectric, reducing the amount of field in contact with the high loss metal surface. Therefore the losses are reduced with respect to the metal dielectric counterpart. In this way, surface plasmon polaritons can travel over long distances.
3.1.2 Hybrid Model of gold (Au), silicon carbide (SiC) and Benzocyclobutene (BCB)

The hybrid phonon polariton LWIR waveguide geometry is shown in Fig. 3.5. The waveguide geometry considered is based on the hybrid plasmonic waveguide in [14]. The hybrid waveguide consists of a dielectric nanowire separated from a planar metallic reflector by a dielectric spacer of lower electrical permittivity than that of the wire.

![Fig. 3.5 Hybrid phonon-polariton waveguide geometry](image)

By tuning the wire diameter and the dielectric spacer thickness, the hybrid waveguide achieves subwavelength confinement and long propagation lengths. Due to the symmetry plane, only half of the structure is shown. For LWIR applications, the wire is made of silicon carbide (SiC) so phonon polariton propagation can take place. Surface phonon polaritons (SPhPs) have been studied less extensively than metal-based surface plasmon-polaritons despite being supported by a number of
materials [4]. They are electromagnetic waves that propagate along the interfaces of polar dielectrics and exhibit a large local-field enhancement near the interfaces at infrared frequencies [29].

Silicon carbide (SiC), benzocyclobutene (BCB), and gold (Au) were the materials for the wire, spacer, and the metal respectively. SiC was used because of its very low loss at approximately 12 µm, where it’s real permittivity $\varepsilon' < 0$ [26]. The goal is to support a mode that will propagate over a length of many wavelengths while maintaining a well-confined field distribution. When SiC is used as a wire in the hybrid model, the choice of substrate limits the spectral range useful for the waveguide because the hybrid condition requires that the absolute permittivity of the wire must be greater than the spacer substrate. Simulations were performed at 11.174µm, 11.78µm, and 12.002µm respectively so as to capture the effect of rapidly changing dielectric constant near the SiC phonon resonance.

The field distributions are shown in Fig. 3.6 for a 0.25µm and 2.5µm spacer. The field scale (colormap) was maintained constant, thus facilitating comparisons. Here we varied the model dimensions as well as the operating frequency. From Fig. 3.6 (a), top row, there is a weak confined mode within the spectral range and dimensions considered. There is a weak confinement in the case of the 0.15µm thick wire as the mode appears large and diffuse in comparison to the modes of the thicker wire structures.
This mode can be identified to be similar to a long range surface plasmon (LRSP), found on metal dielectric slab waveguides where the dimensions of structure play a great role in exciting the mode. In a metal strip over a dielectric substrate, the mode evolves into a plane wave supported by the background as the metal film vanishes. An interesting transition is noted in Fig. 3.6 (a), bottom row. Here, the field concentration shifts location as the wavelength is increased. At first, the field localization is on top of the wire, but with increase in the wavelength, the concentration of the field moves to the space beneath the wire. Though the mode is highly confined at this point, it has a relatively high propagation length.

In Fig. 3.6 (b), top row, we have a long range SPhP. This is a loosely bound mode with a vanishing imaginary part of the wave vector just like a TEM mode wave. Fig. 3.6 (b), bottom row, shows that the field localization shifts from
wire mode to hybrid mode with increasing wavelength. Next, modal area and propagation parameters were extracted. Propagation lengths and normalized modal areas are shown in Fig. 3.7 and Fig. 3.8.

Fig. 3.7 Propagation length as a function of wire thickness, d, for spacer thickness s=0.25 µm (left) and s=2.5 µm (right)

Fig. 3.8 Normalized modal area as a function of wire thickness, d, for spacer thickness s=0.25 µm (left) and s=2.5 µm (right)
Plots in Fig. 3.7, show that the propagation distance is high for very small wires and that it drops rapidly as the wire thickness is increased for a spacer thickness of 0.25μm. The origin of the increased propagation lengths can be postulated to be as a result of less confinement of the coupled modes with thin wires, thus producing lowered damping due to absorption in the wire. A thin wire results in a weak field concentration on top of the wire as shown in the top rows of Fig. 3.6, leading to higher propagation lengths but at the cost of larger modal area.

The rapid decrease in propagation length with increasing wire thickness for 0.25μm spacer is more pronounced as the phonon resonance point is approached. The analysis shows that attenuation and confinement decrease with decreasing wire thickness for a thin spacer. For a spacer thickness of 2.5μm, the general trend from the plots in Fig. 3.8 shows that an increase in wire thickness leads to higher propagation lengths and a lower modal area. At 11.78 μm, enhanced performance is noted and was attributed to the near phonon resonance performance. Simulating at the phonon resonance frequency of 12.4μm showed very low propagation lengths. This was due to the large imaginary permittivity of SiC at this range of frequency leading to high attenuation.

Setting up a cylindrical SiC wire, we obtain the field plots of Fig. 3.9. The propagation constant, attenuation, and the modal area are extracted and tabulated as shown in Table I. It is noted that design has superior performance to the previous design.
### Table I. Propagation Results for Hybrid SPhP and SPP Waveguides

<table>
<thead>
<tr>
<th></th>
<th>Phonon coupled</th>
<th>Plasmon coupled</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wavelength (( \mu m ))</td>
<td>11.174</td>
<td>10.6</td>
</tr>
<tr>
<td>Attenuation (Np/mm)</td>
<td>0.009</td>
<td>2.86</td>
</tr>
<tr>
<td>Propagation length (( \mu m ))</td>
<td>5.56 ( \times 10^4 )</td>
<td>174.82</td>
</tr>
<tr>
<td>Normalized Modal area</td>
<td>2.97 ( \times 10^{-3} )</td>
<td>5.55 ( \times 10^{-4} )</td>
</tr>
</tbody>
</table>

The enhanced performance can be attributed to phonon polariton coupling. Silicon carbide is a crystalline polar dielectric material that can support surface phonon polaritons [77]. It should be noted that SPPs may exist at any interface between a dielectric and a material with a negative real part of its dielectric constant. This is a necessary but not a sufficient condition to guarantee a true SPP.
mode, which can propagate over a substantial distance or which can interact resonantly with an incident electromagnetic wave. The imaginary part of the dielectric constant determines loss in the material and, by extension, the propagation distance. Therefore, the imaginary part of the dielectric constant of the plasmonic material must be relatively small [78]. At LWIR, the imaginary part of the dielectric constant of metals is high and as such, SPP does not perform well. At this range of frequencies, SPhP outperforms SPPs.

The structure has favorable confinement and propagation characteristics due to the involvement of the low loss dielectric waveguide (BCB permittivity of 2.28 and loss tangent of 0.04) and interaction with SiC and phonon polariton of SiC with air. High field concentration is noted on the wire and spacer interface as well as the surrounding near air region. This high concentration of field is attributed to surface phonon coupling. Surface phonon-polariton originates from the resonant coupling between the electromagnetic field and optical phonons in polar dielectrics and is an infrared counterpart of surface plasmon-polariton that usually exists on metal surfaces in the visible and ultraviolet range [29]. These surface waves are modes of the system that can be resonantly excited and they are characterized by large energy densities at the interface, which decays rapidly with distance from the surface. The frequency of this mode occurs in spectral regions where the sample’s dielectric function is negative, which is between the transverse-optic (TO) and the longitudinal-optic (LO) phonon frequencies (TO < SPhP < LO).
The thickness of the dielectric spacer is critical for confinement and optimal confinement occurs for small thickness. For large spacer thickness, the metal plane and dielectric waveguide decouple, leaving essentially lossless dielectric waveguide \[79\].

In order to gain further insight into phonon polariton capability of SiC, we plotted the characteristics of the SiC material within the mid-infrared range. SiC is a polar dielectric that supports phonon coupling in the mid infrared. A closer look at the dispersion characteristics of SiC in Fig. 3.10 shows that it has a phonon resonance at 12.4 \(\mu m\) with negative dielectric permittivity in the mid-IR spectral range between 10.3 and 12.6 \(\mu m\).

![Variation of permittivity of SiC with wavelength](image)

Fig. 3.10 Variation of permittivity of SiC with wavelength
Silicon carbide particles exhibit both electric and magnetic optical resonances. The origin of these resonances lies in the optical phonons supported by these materials. The relative permittivity of SiC exhibits a sharp resonance near 12.5\(\mu m\) due to excitation of transverse optical phonons [80]. At the high frequency side of this resonance, the dielectric function is negative and the optical response is similar to metals. However, unlike metals that exhibit a free electron resonance \((\varepsilon \rightarrow -\infty)\) at dc frequencies \((\omega = 0)\), the SiC phonon resonance occurs at a finite frequency. This allows for a low frequency regime where the permittivity is large and positive with moderate damping [80]. A large absorption should occur where the imaginary part of the dielectric function is large. If the imaginary part of the wave vector is small (less than a tenth of the real part), it indicates that the damping is low and there surface modes exist.

### 3.2 Summary

Achieving high confinement and propagation distance requires a multifaceted approach that takes into consideration the material properties, dimensions, and topology. The structure of a dielectric cylinder above a metal surface has an advantage of not yielding singular fields, which are in general a mechanism of loss at the corners of a rectangular-shaped waveguides [10,22]. Surface phonon interaction enhances propagation distance and, depending on the materials and configuration used, long propagation length as well as high confinement is feasible. The hybrid structure supports highly confined guided
modes. The performance of the rectangular hybrids of Si, ZNS, and Au as well as SiC, BCB, and AU perform better than that of doped zinc oxides in [32], whose propagation lengths are predicted to be 70, 115, and 370 μm for the Zn0.995Al0.005O, Zn0.982Ga0.018O, and Zn0.974Ga0.026O films respectively. Further, it is noted that cylindrical wires perform better than rectangular ones. This is because rectangle-shaped waveguides tend to yield singular fields at the sharp corners of waveguides, which results in a large propagation loss [81]. Phonon waveguides with sharp features may provide strong local field enhancement, but this could also lead to higher propagation losses. Since the thickness of the wire and spacer strongly influence the overlap of the fields and hence the coupling between the interface SPhP modes, adjusting the cross-sectional geometries can help in tuning the field localization and rate of attenuation of the mode.

The performance of cylindrical SiC, BCB, and gold combination has a better performance in long propagation than doped silicon and silicides, whose propagation lengths are predicted in [79]. It is postulated that, at least in principle both structures, rectangular and cylindrical wire hybrids can be fabricated by current nano fabrication technology. However, compared with the rectangular shaped waveguides, the fabrication process for the cylindrical waveguide is not easy since placing a cylindrical nano wire precisely to implement the cylindrical hybrid waveguide may be quite challenging with current nano fabrication technology. Though the performance of a rectangular wire based hybrid is inferior compared to a cylindrical wire based one, it is physically more feasible to set up
rectangular shaped waveguides because the manufacturing challenges are lower in rectangular shaped waveguide though it sacrifices the waveguide performance. Straight waveguides are a fundamental building Bloch for any passive element and as such, it is important to understand their behavior.

In conclusion, it is noted that a hybrid phonon waveguide incorporating metal, low index and high index dielectric materials, exhibits longer propagation distances than metal dielectric combinations. Surface phonon polariton (SPhP) reveals the coupling between EM field and lattice vibration around the surface of ionic crystals. Although it has attracted less attention than surface plasmon polariton (SPP), there are still many useful prospects in surface enhanced infrared (IR) absorption and transmission, high-density IR data storage, coherent thermal emission, subwavelength scale phononic photonics and negative index metamaterials. However, due to the material’s natural characteristics, SPhP is normally excited in a narrow and fixed frequency range in infrared.
Chapter 4

FINITE ELEMENT METHOD ANALYSIS OF A LINEAR POLARIZED METASURFACE

This part of the dissertation reports on the design and model of a metasurface composed of an array of microstrip patches using the FEM technique. The objective of this research was to design and characterize a leaky wave metasurface. Understanding how the propagation and leakage constants of a leaky mode in a traveling wave antenna are affected by modifying the geometry of the structure is vital for improving the far-field pattern and bandwidth of a leaky wave surface. A parametric study of various dimension parameters was done to ascertain their effect on the overall metasurface behavior. It was noted that periodicity as well as the length of the patch played a critical role in the overall metasurface characteristics. Detailed properties of the leaky wave antenna are presented and an FSS surface is designed and simulated. The FEM method, which is a computational electromagnetic (CEM) technique was used for simulation. The propagation and attenuation constants were extracted from the scattering parameters resulting from the FEM simulation. A possible application of this metasurface is in Infrared detectors.
4.1 Analysis and Design of a Linear Polarized Leaky Wave Metasurface

A leaky-wave antenna (LWA) is basically a guiding structure that possesses a mechanism that allows leakage of power along the structure [42,47,54,55,82]. The leaky-wave antenna belongs to the travelling wave type of antennas. The main difference between a travelling wave antenna and a leaky wave antenna is that travelling waves structures have a slow mode and therefore radiate at the ends of the structure [54], while Leaky wave antennas radiate continuously along its length. The basics on leaky-wave antennas have been reviewed in book chapters such as [47,54,63]. A recent summary on the latest advances can be found in [55]. One of the main advantages of a LWA is that it has inherent frequency scanning capabilities; hence beam scanning is possible without the need for additional phase shifters [55]. With the advent of planar leaky-wave antennas, there has been broad research on improving their radiation characteristics as well as discovering the theory behind them. Their low profile, high directivity, and simple feeding as well as the property of frequency scanning makes them popular in the microwave band [63, 83, 84]. In particular, the compatibility with printed circuit board technology, their low profile, easiness of fabrication, and integration with other planar components are the strongest features of these antennas [85].

4.2 Model Analysis and Design

In this section, analysis and characterization of a periodic leaky wave surface is done with the aim of designing a metasurface leaking at an arbitrary
angle. Prior to commencing the design, it was necessary to take into consideration a number of structural factors. The substrate material is a key component. A low cost planar substrate that meets the design specification would be more desirable. In addition, the dimensions of the transmission lines have to be taken into consideration so as to accommodate fabrication limits. Very narrow transmission lines have high impedance and are difficult to fabricate, while thick ones are easier to fabricate and have low impedance; however, they limit the periodic spacing range and design flexibility.

The adopted microstrip LWA configuration is shown in Fig. 4.1 (a) and its main geometrical parameters in Fig. 4.1 (b). From the diagrams, it can be noted that the periodic planar structure is a microstrip line of negligible thickness etched on the surface of a grounded dielectric layer, which is periodically modulated. The patch is made of gold with dispersive material properties at LWIR. The background dielectric slab is made of zinc sulphide (ZNS) material. The dielectric slab has a relative permittivity $\varepsilon_r$ which is frequency dependent, a relative permeability $\mu_r=1$, and a thickness, $h$. Along the $y$ direction, the array is infinite with a lattice constant (period) $p$. Along this direction, the surface is modeled as a modulated impedance stripline, with high impedance in the connecting microstrip lines and low impedance in the patches. The periodic structure is infinite in the $x$ and $y$ direction, as shown in Fig. 4.1 (c). This is captured in the model via use of periodic boundaries. This class of printed structures allows for leakage, both in the dielectric substrate and into free space [86].
FEM HFSS was the numerical tool used to model the metasurface. Fig. 4.1 (a) shows the physical layout of the unit cell in Ansoft’s High Frequency Structure Simulator (HFSS) software package. In modeling the metasurface, two types of driven modal feeds were considered, the waveport and the Floquet port. Dispersion diagrams based on the S-parameters from the driven mode simulation for the proposed unit cell are investigated in detail using the waveport feed. Waveport analysis is applied on the modulated patch to extract the scattering parameters for a single patch. The transmission coefficient scattering parameter \( S_{21} \) is particularly important because the phase constant and leakage constant information can be extracted from it [55]. By applying Floquet port and periodic boundary conditions, it was possible to characterize the absorption/emission properties of the metasurface.

Numerical simulations were performed to determine the dispersion diagrams for both the transverse electric (TE) and transverse magnetic (TM)
polarization. The patch dimensions were selected so as to avoid patch plasmon as well as phonon resonance by use of the equation [87]

$$\lambda = (\pm \sin \theta_i + 1) \frac{p}{m}$$  \hspace{1cm} (4.1)

where $p$ is periodicity, $\theta_i$ is the angle of incidence, and $m$ is the order. The patch length ($b$) determines the operating frequency of the antenna, which satisfies the resonance condition of patch [65]. Limited literature on the structure is available [88]; hence, an attempt was made to characterize the behavior of its leaky nature.

Preliminary studies on the structure indicated that the microstrip array operated as a quasi-uniform structure; therefore, the propagation constant, which determines the leaking angle, is independent of the number of elements/patches. Therefore, an approximate method based on a single unit cell in isolation was employed to get the dimensions of the patch needed to leak at an arbitrary angle. The fundamental mode for a quasi-leaky wave antenna is a fast wave, and the radiation occurs via the fundamental mode and not via the space harmonics [55]. Leaky modes are characterized by a complex propagation constant along the longitudinal direction ($y$ axis in this case) of the patch as [54, 55, 86, 89]

$$k_y = \beta_y - j\alpha_y$$  \hspace{1cm} (4.2)

where $\beta_y$ stands for the propagation or phase constant, and $\alpha_y$ is the attenuation rate due to the radiation or leakage induced by the leaking wave. To design and analyze a LWA, the dispersion curves of the constituent leaky-mode are critical. Particularly $\beta_y$ determines the pointing or radiating angle of the LWA in the
elevation plane. The main beamwidth is determined by the length of the antenna and the aperture illumination [55, 88].

Patch antennas are broadside radiators by nature. The patches act as resonant cavities while the metal strips bridge the gaps between the patches and enhance the capacitance between the patches [53]. This type of antenna can also be modeled as a cavity LWA due to its analogy with optical resonant cavities. The electromagnetic waves that arise from the feeding point are bounced back and forth between the patch and the ground planes, becoming leaky-modes of the guiding structure when the patch is modulated. The top patch acts as a Partially Reflective Surface (PRS); with its transparency, it controls the amount of energy that is leaked from the cavity to free space [90].

4.2.1 Choice of substrate

The metasurface was designed on an electrically thin substrate of low permittivity in order to reduce surface wave losses. Surface waves reduce antenna efficiency and gain, limit bandwidth, increase end-fire radiation, increase cross-polarization levels, and limit the applicable frequency range of Microstrip antennas [91]. A zinc sulphide (ZNS) substrate was given priority since it has low relative permittivity in the infrared range of frequency. Typically, bandwidth increases rapidly as the substrate dielectric constant nears that of free space. The drawback of a lower dielectric constant is that it leads to a very low leakage constant (α) across most of the leaky band, which results in little energy radiating per unit length [55]. ZNS is an inorganic compound naturally found as the mineral sphalerite. The
atomic geometry of zinc and sulphide are tetrahedral and this leads to some of its characteristics such as high density and transparent nature of the material; hence, it can be used as a window for visible optics and infrared optics [92, 93].

4.2.2 Approximate design equations; Patch and line dimensions

Leaky wave antenna design is an iterative process. Design equations are developed using standard antenna theory. The analysis and expressions are only approximate but nevertheless provide good starting values for the design parameters, which are refined through full-wave electromagnetic simulation. It was assumed that the leakage across the antenna structure was small. Standard patch antenna theory was used to determine the patch dimensions so that an individual patch was in the fast wave region along the patch. This was necessary since preliminary research had shown that a rectangular array had quasi-uniform characteristics. The initial dimensions of the patch were approximated to be given by [53,65]

\[
b = 0.49 \frac{\lambda_o}{\sqrt{\varepsilon_{eff}}}
\]

(4.3)

\[
a = \frac{\lambda_o}{2} \sqrt{\frac{2}{\varepsilon_{eff} + 1}}
\]

(4.4)

where \(b\) is the length of the patch and \(a\) is the width of the patch. Effective permittivity \((\varepsilon_{eff})\) was extracted by Eigen iteration of a section of the patch 10nm
width. A microstrip line was extended for interconnecting the patches in the array as shown in Fig. 4.1 (b). With the initial patch dimensions in place, a model was set up and dimensions of the patch swept so as to meet the required phase constant to enable our patch leak at $30^\circ$ at 28.3 THz.

4.2.3 Circuit equivalent of unit cell in isolation

A circuit representation of the unit cell was used to aid in the design. It should be understood that the equivalent-circuit model proposed here is not a complete substitute of the full-wave method. Instead, it is a very convenient complement that helps to drastically reduce the computational effort. In fact, due to the current disposal of full-wave electromagnetic simulators, the numerical benefit given by the equivalent circuit may not be so crucial. However, the key feature of the equivalent-circuit approach is that it provides a simple and accurate comprehension of the problem, which makes possible many important predictions on the behavior and role of the different elements of the structure under study. This predictive nature can be fundamental for many analysis and design applications. The equivalent circuit clearly showed how the elements changed with each iterative modification of the unit cell geometry. This equivalent circuit offered insight when making adjustments to the patch dimensions to achieve the desired propagation constant. The width, length, and height of the patch were iteratively adjusted using the insight gained from the equivalent circuit to attain the required propagation constant. The equivalent circuit of this unit cell in isolation is shown in Fig. 4.2.
Fig. 4.2 Unit cell equivalent circuit [94,95]

where

\[ G = \frac{1}{90} \left( \frac{w}{\lambda_0} \right)^2 \text{ for } \frac{w}{\lambda_0} \ll 1 \] (4.5)

\[ G = \frac{1}{120} \frac{w}{\lambda_0} \text{ for } w \gg 1 \] (4.6)

\( w \) is the width of the patch.

Equivalent capacitance is given by [94]

\[ C = \frac{\Delta l}{vZ_0} \] (4.7)

\( v \) is phase velocity, \( Z_0 \) is the characteristic impedance, and \( \Delta l \) is the normalized line extension given in [53,65,94] as

\[ \frac{\Delta l}{h} = 0.412 \frac{\varepsilon_{eff} + 0.300w/h + 0.262}{\varepsilon_{eff} - 0.258w/h + 0.813} \] (4.8)

Here \( h \) is thickness and \( \varepsilon_{eff} \) is the effective permittivity of the substrate.

The leaky surface was simulated with varying geometrical parameters and the resulting effects on the propagation constants were analyzed to meet the design
specifications. The geometrical parameters considered were the periodicity $p$, the length of the patch, $b$, and the height of the substrate $h$.

4.3 Simulation

Simulation/modeling of the unit cell in HFSS using waveport analysis provides quick predictions of the fundamental propagation constant; however, it does not take into account the mutual coupling effects. The diagram in Fig. 4.3 shows the meshing due to the FEM solution. On the metallic patch, the mesh intensity is higher than on the dielectric material. This is due to adaptive analysis whereby mesh is automatically refined in critical areas where the $E$ field is concentrated. In HFSS, the geometric model is automatically divided into a large number of tetrahedra, where a single tetrahedron is a four-sided pyramid. This collection of tetrahedra is referred to as the finite element mesh. There is a tradeoff among the size of the mesh, the desired level of accuracy, and the amount of computing resources.
By performing solve port only with waveport analysis, it was noted that the considered structure can support $TM_{00}$, $TE_{10}$, $TE_{01}$, and $TE_{11}$ natural modes when four modes were considered, as shown in Fig. 4.4.
Since the antenna is operating as a quasi-periodic structure, it was prudent to operate on the dominant mode. Single mode operation is assured by specifying the number of modes in the feed port to be one. Therefore, in the analysis of the proposed LWA, it is assumed that the LWA operates in the dominant TM$_{00}$ leaky-mode with propagation directed along the $y$ axis in Fig. 4.1. The number of modes supported was specified to two so as to obtain the characteristics of TM and TE.

In this study, we determine the influence of the geometrical parameters on the complex propagation constant ($\alpha$ and $\beta$) in a leaky-wave antenna based on microstrip array structure and extract some fundamental parameters. The
fundamental parameters in the analysis and characterization of the leaky wave antenna as well as a metasurface include the following: the return loss ($S_{11}$), the transmission coefficient ($S_{21}$), phase constant ($\beta$), leakage constant ($\alpha$), dispersion diagram, and Bloch impedance.

### 4.4 Parametric Study of the Influence of Geometrical Parameters on Fundamental Parameters

#### 4.4.1 Periodicity of the unit-cell variation

The periodic dimension of the unit cell was considered and its effect on the characteristics of the leaky surface noted. In this parametrical study, the dimensions that remain constant have the following values: $b$ (patch length) = 2.6 $\mu$m, $a$ (patch width) = 2.157 $\mu$m, $h$ (substrate thickness) = 0.65 $\mu$m, and $p$ (periodicity) varies from 3.5 $\mu$m to 4.5$\mu$m. A frequency sweep was also run, from $20 - 40$ $THz$ in steps of 0.5 $THz$.

To start with, a plot of return loss $S_{11}$ with frequency with varying periodicity is plotted.

#### 4.4.1(a) Return loss $S_{11}$ plot

$S_{11}$ is used to quantify the amount of power reflected back towards the source. The $S_{11}$ plot can be used to predict the impedance bandwidth with a typical antenna; however, for a travelling wave case antenna, the propagation constant is
the one primarily used for bandwidth predication [63,89]. A threshold of $-10dB$ for $S_{11}$ (also known as return loss) parameter is imperative for appreciable radiation to be noted in an antenna. The $S_{11}$ parameters here are in reference to coupling energy along the patch in the propagation direction. Fig. 4.5 presents return loss with variation in periodicity and frequency. It can be noted that the return loss is a function of frequency with the mid-frequency range ($24 – 32 THz$) having the lowest return loss. There exists a narrow range of return loss at $38 – 40 THz$. From Fig. 4.5, $S_{11}$ improves slightly with a decrease in periodicity. At $3.5\mu m$, the antenna presents a $-10dB$ relative bandwidth of $31.5 – 24.5/27.5 = 25.5\%$. There is a high coupling of power within the LWIR range to the patch along the direction of propagation with the highest return loss occurring between $36 – 38 THz$. 
4.4.1(b) $S_{21}$ plot

Transmission parameter $S_{21}$ specifies relative surface wave power between the two ports. The amount of accepted energy at the input is consumed not only by radiation but also by losses. Leaky regions have low $S_{21}$ dB, while bounded regions are high. It can be explained that the bounded surface waveguides the power from port one to two. Leaky waves reduce this power since it escapes from the surface.
Transmission parameter $S_{21}$, which is at least below -1 dB within the LWIR, promises some leakage radiation performance. Here, the wave is guided with some leakage/attenuation from one port to the other. The $S_{21}$ parameter has decreased due to the radiation losses present in the structure. Between $36 - 38 \, THz$, the $S_{21}$ drops sharply, meaning higher leakage is being experienced and this further leads to wide beamwidth of leaked wave. However from the $S_{11}$ plot, this range (36-38 THz) also received the highest return loss. It can be observed that regions with low return loss have a high transmission coefficient and the converse is also true.
4.4.1(c) *Phase constant and leakage constant*

A major step in the design of a leaky wave antenna is the determination of the phase constant and the attenuation constant. In the dissertation, use of commercial software ANSY HFSS is applied as a numerical computation tool to aid in extracting these two critical parameters. With the knowledge of the complex wave number $k_y$ or $\beta_y$ and $\alpha_y$, a proper design can be done effectively. The beam direction, radiation efficiency, beam-steering with frequency, and the handling of the side lobe level can be controlled efficiently [63]. To extract the phase constant and leakage constant, a waveport was applied as the excitation for a single element and the following equations used to extract the attenuation constant and phase constant respectively [55]:

\[
\alpha = \frac{-\ln(|S_{21}|)}{l} \quad (4.9)
\]
\[
\beta = \frac{-Arg(s_{21})}{l} \quad (4.10)
\]

where $S_{21}$ is the transmission coefficient scattering parameter and $l$ is the length of the aperture. Energy propagation along the antenna is lost due to the conductor, dielectric and radiation losses [54]. $S_{21}$ denotes the amount of energy arriving at the output terminal. The attenuation constant $\alpha$ and the phase constant $\beta$ components of the propagation constant give an accurate means of comparing the bandwidth, leakage rate, main beam direction, and approximate far-field pattern of different traveling wave antennas [63, 89].


The phase constant $\beta$ determines the direction of the leaking wave in a leaky wave structure. For a unidirectional leaky wave antenna, $\beta > 0$, hence, the beam maximum is not at broadside. For uniform or quasi-uniform leaky wave antennas, the optimum condition for maximum power density radiated at broadside is at the point where $\beta = \alpha$ [55,58,61]. The beam radiated by a bidirectional leaky wave will always point at broadside for $\beta < \alpha$, and the point $\beta = \alpha$ is a beam splitting point [55]. In a 1-D periodic structure, the condition $\beta = \alpha$ gives rise to maximum radiation at broadside in the lossless case and to the splitting of a single peak of the radiation pattern into two peaks in both lossy and lossless cases [61].

Since leaky-wave antennas are typically long electrically by having large antenna apertures, low attenuation is preferable. If $\alpha$ is high, power leaks away rapidly and the effective aperture is short, beating the purpose of the leaky phenomena or having an array. This leads to a wide radiation pattern and low directivity [89]. For convenience, the normalized phase and attenuation constants are normally included on the same graph, as shown in Fig. 4.7. Phase constant is scanned as the frequency is changed. It can be noted that the leakage constant is almost constant in the mid-frequency range (25-30THz).
Fig. 4.7 Normalized phase and leakage constant plot variation with frequency and periodicity for TM mode

The leaky-wave mode is characterized by a propagation constant $\beta_y$ lower than the free space propagation constant ($k_o$). This mode then travels along the guiding structure faster than the speed of the light, and at the same time, it leaks energy with a leakage constant $\alpha_y$. From Fig. 4.7, it can be noted that the normalized phase constant $\frac{\beta}{k_o}$ is $> 1$ at the low edge of frequencies where the periodicity of the patch is much less than a wavelength ($p < \lambda_o$). By increasing the frequency, the normalized phase constant shifts to the leaky wave zone. It can be inferred that since the wavelength is relatively long at low frequencies, the
propagating mode just perceives the metallic plate; nevertheless, the wave is purely bound through its entire pathway because its propagation constant is bigger than the constant of free space. When the wavelength reduces with increasing frequency, the propagating mode begins to interact with the configuration of the LWA and converts into a leaky wave mode. The modes in the considered range did not have a cutoff frequency, a characteristic of $TM_0$.

The leakage constant remains relatively constant with periodicity within the range considered, but with some sharp variation at the 37 THz point. Large values of leakage constant are associated with reactive effects and with degradation of the scan performances [96]. It is important to highlight that the “same” leaky wave (same $\beta_y$ and same $\alpha_y$) is never found at two different frequencies. However, several leaky-mode solutions with the same leakage rate $\alpha_y$ but different phase constant $\beta_y$ can exist [64].

From Fig. 4.7, it is clear that for the 24-32THz range, the leaky modes present a lower almost constant leakage constant $\alpha_y$. Predictably, the dispersion relation shows high attenuation at low frequencies due to increased series reactance of capacitance of the patch and transmission lines.

When the TE mode was taken into consideration, it can be noted that though the phase constant is in the fast wave region, the leakage constant is very low. Appreciable leakage constant is noted between 38-40THz. as shown in Fig. 4.8. Essentially, this shows the non-existence of the TE leaky wave mode in the model.
Fig. 4.8 TE mode normalized phase and leakage constant variation with periodicity and frequency

4.4.1(d) Dispersion diagrams

Dispersion in principle is due to the velocity of the different frequency components being different. It can also be perceived as the result of other parameters changing with frequency. A dispersion diagram is a plot of phase constant with frequency. Information that can be gathered from a dispersion diagram includes frequencies at broadside radiation, fast (radiating) and slow (non-radiating frequencies) and directions of radiating/leaking beams (forward or backward waves). A wave above the light line is radiating while below the light line is bound. Another way to present the bound and leaky waves is to use a
Brillouin diagram. Detailed literature on Brillouin diagram can be found in [54, 55, 64].

To derive the dispersion diagram from the scattering parameters, the theory of the transmission line model is applied. The ABCD matrix of a lossless long transmission line of length \( l \) with characteristic impedance \( Z_o \) and phase constant \( \beta \) can be written as [97]

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
\cos \beta l & j Z_o \sin \beta l \\
j Y_o \sin \beta l & \cos \beta l
\end{bmatrix}
\] (4.11)

However, the transmission matrix can also be written in terms of scattering parameters as [97]

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
\frac{(1 + s_{11})(1 - s_{22}) + s_{12}s_{21}}{2s_{21}} Z_B & \frac{(1 + s_{11})(1 + s_{22}) - s_{12}s_{21}}{2s_{21}} \\
\frac{1}{Z_B} \frac{(1 - s_{11})(1 - s_{22}) - s_{12}s_{21}}{2s_{21}} & \frac{(1 - s_{11})(1 + s_{22}) + s_{12}s_{21}}{2s_{21}}
\end{bmatrix}
\] (4.12)

Due to the symmetry of the model, equating \( A = D \) is valid. Therefore,

\[
\cos \beta l = \frac{(1 - s_{11})(1 + s_{22}) + s_{12}s_{21}}{2s_{21}}
\] (4.13)

Hence,

\[
\beta l = \cos^{-1} \left( \frac{1 - S_{11}S_{22} + S_{12}S_{21}}{2S_{21}} \right)
\] (4.14)

Plotting \( \beta l \) with frequency, we get the dispersion diagram shown in Fig. 4.9.
From the dispersion, radiating and bound wave regions can be noted. The light line plotted in Fig. 4.9 gives rise to two distinct regions: the radiating region (fast-wave) above it and the guiding region (slow-wave) below the line. Regions below the light line have bound waves while those above the light line are radiating/leaking. Points of broadside radiation can also be seen precisely. These points correspond to regions where the phase constant is zero or approaching zero. As the periodicity is changed, the broadside radiation point shifts with frequency at broadside, decreasing as the periodicity is increased. The physics underlying this radiation behavior is related to the resonant nature of the patch [53].
4.4.1(e) Effect of periodicity on Bloch impedance

For matching, it is important to know the Bloch impedance of a periodic structure. The Bloch Impedance $Z_B$, often referred to as characteristic impedance, is the ratio between the voltage and current at the terminal of a periodic unit cell, and it is computed at the input interface of a unit cell. The value of the Bloch impedance varies within the unit cell, but it is periodically identical at each periodic cross sectional plane and can be retrieved from the field solution by simulation [55,61,85]. Due to the model symmetry along the transverse axis, the Bloch impedance can be derived by equating

$$s_{12} = s_{21}$$

Hence,

$$\frac{j Z_o \sin \beta l}{j Y_o \sin \beta l} = \frac{Z_B}{Z_B} \frac{(1 + s_{11})(1 + s_{22}) - s_{12}s_{21}}{2s_{21}}$$

$$\frac{(1 - s_{11})(1 - s_{22}) - s_{12}s_{21}}{2s_{21}}$$

Therefore,

$$Z_B = Z_o \left[ \frac{1 - 2s_{11} + s_{11}s_{22} - s_{21}s_{12}}{1 + 2s_{11} + s_{11}s_{22} - s_{21}s_{12}} \right]^{1/2}$$

The Bloch impedance value gives some useful information for the final impedance matching. Plot of the Bloch impedance is given in Fig. 4.10 and Fig. 4.11.
Fig. 4.10 Real impedance plot variation with frequency and periodicity

Fig. 4.11 Imaginary impedance plot variation with frequency and periodicity
Due to the presence of radiation losses, the Bloch impedance obtained is complex and presents a capacitive behavior ($X < 0$) for backward leaking frequencies and an inductive behavior ($X > 0$) for forward leaking frequencies [61]. The results for a real Bloch impedance for the unit cell simulation are shown in Fig. 4.10, where real impedance denotes $R = re(Z(k_y, k_o))$. The points of broadside radiation, as depicted by the dispersion curve in Fig. 4.9, are points of lowest real impedance, as shown in the Fig. 4.10. Just before the broadside radiation point, it is noted that the real impedance drops sharply. Also, the real impedance drops with an increase in periodicity. Real wave impedance signifies wave propagation whereas imaginary impedance is an indication of evanescent waves [54,61].

The imaginary impedance ($X = im(Z(k_y, k_o))$) peaks at the transition regions. The first jump is at the transition where the backward leaking mode converts to the broadside leaking mode. The second one is the transition from a broadside leaking to forward leaking wave. It can be noted that the leaky waves in this structure are improper waves that radiate in the forward and backward direction. This is due to the fact that proper solutions correspond to inductive impedance $X > 0$, whereas the improper ones to the capacitive case $X < 0$ [96]. Note that at broadside radiation frequency, the Bloch impedance changes considerably. The imaginary part of $Z_B$ increases significantly and, at the same time, the real part drops drastically. At broadside, the Bloch impedance is complex with a dominant imaginary part with very low values of the real part [96].
It is worth noting that a well-designed leaky wave antenna leaks power traveling along the structure, and an insignificant amount of power is left by the time the wave reaches the end of the structure. Therefore, most adequately fed leaky-wave antennas do not present any mismatch problem. This is because a leaky mode is normally a component of the feeding mode coming from the feed. Even if reflected, this vestigial power cannot cause any serious mismatch problem \[55\].

4.4.1(f) Beamwidth

The 3dB pattern beamwidth in radians for a 1D leaky wave antenna is given by \[55\]

\[
\Delta \theta = 2 \left( \frac{\alpha}{k_o} \right) sec \theta_o 
\]

4.17

The formula assumes an infinite aperture. From the formula, it can be inferred that large values of leakage constant (\(\alpha\)) provide wider main beams because of the fact that all the power has leaked out after a short length. This implies that the resulting radiation aperture is small and the resulting fan beam is wide. In contrast, small values of \(\alpha\) provide narrow beams, since the antenna has to be sufficiently long to leak all the power out of it. This results in a large radiation aperture and consequently, a very directive fan beam \[82\]. In the case of a 1D leaky wave antenna, the beam is narrow in the scan plane and wide in the transverse plane, whereas in 2D leaky wave antennas, the beam is narrow in both planes. Therefore,
the radiation pattern is known as a pencil beam [55]. The plot of beamwidth variation with frequency is given in Fig. 4.12.

The beamwidth narrows and is more stable over a long frequency range as the periodicity is made larger, as shown in Fig. 4.12. A useful property of leaky wave antennas is that their beam can be frequency-scanned with little beam shape deterioration over relatively large sweep angles. Higher periodicity can be noted to hold this property in Fig. 4.12.

As a general rule, the phase constant of the leaky wave controls the beam angle, while the leakage/attenuation constant controls the beamwidth [55,63]. The leakage constant is high in the higher frequency regions and this implies that the
obtained radiation beamwidth will consist of a broad beam, since there exists high attenuation within this range.

4.4.1(g) Effect of periodicity on the leakage angle

The leaking wave along an aperture produces a beam with its maximum at an angle $\theta$ from broadside governed by the following equation [54, 55, 63]:

$$\theta_0 = \sin^{-1} \left( \frac{\beta(\omega)}{k_0} \right) = \sin^{-1} \left( \frac{c\beta(\omega)}{\omega} \right)$$

(4.18)

Fig. 4.13 shows the dependence of the leakage angle to periodicity. Based on these results, it is clearly seen that the leakage angle is strongly affected by periodicity (dimension of the unit cell).

Fig. 4.13 Variation of angle of leakage with periodicity $p$
The variation of the leakage angle with periodicity shows that the rate of frequency scanning decreases with an increase in periodicity. As the periodicity is increased, the angle of leakage decreases as well as the scanning sensitivity.

4.4.2 Patch length variation

The longitudinal patch dimensions determine the series resonance [53,65]. Since the patch width has already been fixed, it remains to tune the series resonance by modifying the length of the unit cell. With patch length variation, the distance the current has to flow varies, affecting the value of serial inductance. The gap between the patches makes up the serial capacitance. This capacitance is modified by the connecting strips dimensions [95].
It can be seen from Fig. 4.14 that as the length of the patch \( b \) is increased, the normalized phase constant \( \beta_y \) reduces. The point of broadside radiation remains, but the range of broadside radiation broadens as the length is increased. Leakage rate \( \alpha_y \) decreases with increase in length in the lower end of frequencies (20-26 THz), while \( \alpha_y \) increases with an increase in patch length in the mid-frequency range (27-32THz). From Fig. 4.14, patch length \( b \) strongly affects the radiation rate of the leaky mode due to the fact that \( b \) controls the transparency of the metasurface [98, 99, 100].

The variation of \( S_{11} \) with patch length is considered in this section. The \( S_{11} \) parameters here are in reference to coupling energy along the patch in the
propagation direction. As the patch length \( b \) is varied, \( S_{11} \) changes, and it can be seen from Fig. 4.15 that \( S_{11} \) is strongly dependent on \( b \). As the patch length increases, the resonant frequency of the patch decreases.

![S11 variation with patch length](image)

**Fig. 4.15** Plot of return loss with patch length variation

On the other hand, the transmission coefficient \( S_{21} \) drops as the patch length is increased, as shown in Fig. 4.16. It can be inferred that an increase in the patch length creates more aperture surface where leakage is experienced. This has also a direct impact on the beamwidth.
Fig. 4.16 Plot of transmission coefficient with patch length variation

Fig. 4.17 shows the variation of angle of leakage with frequency as the patch length $b$ is varied. The plots have a similar trend and are closely packed. The gradient of the plots increases as the patch length decreases, making an angle of leakage $\theta_o$ increase for lower values of $b$. It can be seen that the length of the patch $b$ affects the pointing angle dispersion curves.
From the patch length variation analysis, it can be concluded that the patch length variation affects the angle of leakage, making it sharper for smaller patch lengths. Thus, this feature can be applied to the control of the radiation rate of the LWA for a given design frequency. Consequently, the possibility of controlling the radiation rate by varying the patch length is verified. Yet, when $b$ is varied, a second order effect occurs: the pointing angle $\theta_o$ is also altered to some extent, as shown in Fig. 4.17. This deviation is due to the dependence of $\theta_o$ on the phase of $\beta$, which also varies with patch length as illustrated in Fig. 4.14.

Beamwidth is a critical parameter for a leaky wave surface. The beamwidth is noted to be strongly affected by the patch length $b$, as depicted in Fig. 4.18.
low value of $b$ leads to a narrow beamwidth within a short bandwidth. A high value of $b$ leads to a wide beamwidth, which spans over a longer bandwidth range.

![Fig. 4.18 Plot of beamwidth with patch length variation](image)

When the leaky surface has a finite aperture, the beamwidth is determined primarily by the fixed aperture length and the value of leakage constant influences it slightly [82]; in addition, the leakage constant primarily affects the radiation efficiency.

The plots in Fig. 4.19 and 4.20 show that the Bloch impedance is strongly dependent on the patch length dimensions. An increase in patch length leads to an increase in real and imaginary impedances. This can be attributed to an increase of a metallic surface, meaning more resistance and inductance, as well as the capacitance with the substrate.
Fig. 4.19 Plot of real impedance with patch length variation
It can be noted that at broadside, the input impedance match becomes poor.

In the next part, the effect of patch variation is determined for the phase constant. An increase of the patch length $b$ has the effect of broadening the broadside radiation point, as shown in Fig. 4.21. The dispersion characteristics remain relatively constant as $b$ is varied. As the patch length is increased, the bandwidth above the light line increases.
4.4.3 Thickness of the substrate variation

The influence of the slab thickness in the response of the leaky wave antenna array is expected to be similar to the permittivity effect, due to the fact that it also controls the amount of energy coupled into surface wave modes excited in the substrates [100, 101]. Figs. 4.22 and 4.23 show the variation of $S_{11}$ and $S_{21}$ with substrate thickness variation respectively. $S_{11}$ improves with substrate thickness $h$, as shown in Fig. 4.22. It is postulated that a bigger substrate thickness enables coupling of more power.

Fig. 4.21 Dispersion curve with patch length variation
Fig. 4.22 Plot of return loss with substrate thickness variation

Variation of return loss with substrate thickness h

Variation of S21 with substrate thickness (h)

120
Fig. 4.23 Plot of transmission coefficient with substrate thickness variation $S_{21}$ becomes higher as substrate thickness is increased, as shown in Fig. 4.23. A thicker substrate provides a guided surface wave rather than a leaking wave.

Plotting the normalized phase and leakage constant, it is noted that while the phase constant slightly increases with an increase in substrate thickness, the leakage constant slightly decreases, as shown in Fig. 4.24.

Revisiting a grounded dielectric slab, it is noted that from the expression, $\beta_y^2 + \beta_z^2 = \beta_o^2$, when the substrate is made thick ($h$ large), the propagation constant
along the $z$ axis ($\beta_z$) reduces; hence $\beta_y$ increases as $\beta_o$ is constant. Thus there is a decrement of propagation constant of the vertical component of the electric fields $E_z$, inside the substrate between the top conductor and the bottom plate. From Fig. 4.24, it is also noticed that the leakage constant has an inverse relation to the thickness of the substrate and by extension to the angle of leakage. Higher radiation rates are naturally associated with lower radiation angles because the leaky wave reaches the top radiating surface more times per unit length [89]. However, like permittivity, the height of the substrate is usually dictated by the material availability.

Effect on the propagation constant through mutual coupling with neighboring elements is low if the separation distance is at least $0.25\lambda_g$ and virtually no interaction if the spacing is over $0.4\lambda_g$ where $\lambda_g = \frac{\lambda_o}{\sqrt{\varepsilon_r}}$ [53].

Considering the angle of leakage with substrate thickness variation, Fig. 4.25 shows that changing the substrate thickness has some impact on the angle of leaking. It is well known [89] that, as the pointing angle of a leaky-wave is increased, the associated leakage rate decreases.
Fig. 4.25 Plot of angle of leakage versus substrate thickness variation

Considering the beamwidth variation with substrate thickness, it is noted that the beamwidth is minimally affected by the substrate thickness variation, as depicted in Fig. 4.26. The real and imaginary Bloch impedances are plotted in Figs. 4.27 and 4.28 respectively. Real and imaginary Bloch impedance are affected by a change in substrate thickness $h$. Both increase as the substrate is made thicker, as shown in Figs. 4.27 and 4.28 respectively.
Fig. 4.26 Plot of beamwidth with substrate thickness variation

Variation of Real impedance with substrate thickness (h)
Thin substrate of low permittivity reduces surface wave losses [101].

Dispersion curves with substrate thickness variation are shown in Fig. 4.29.
The dispersion curve is slightly influenced by the substrate thickness, as shown in Fig. 4.29. The point of broadside radiation is maintained as well as its range.

Effects of the substrate thickness on the insertion loss, dispersion performance, and other characteristics of the fast wave region have been analyzed. The substrate strongly affects the frequency dispersion, thus providing the control over the pointing angle at a fixed frequency. In order to shape the radiation pattern of a leaky wave surface, both the phase constant ($\beta$) and the leakage rate ($\alpha$) should be independently varied, as was explained in [89]. Changing the periodicity, length of patch, and substrate thickness allow flexible variation of the leakage rate and phase constant of the leaky mode, which propagates in the proposed leaky wave
surface. The independent control of these two parameters is of key importance for the synthesis and the flexible adjustment of the radiation pattern of a practical leaky FSS.

4.5 Metasurface with a Beam Leaking at 30 deg at 10.6μm

Based on the analysis of the designed metasurface in the previous section, it can be seen that all the geometrical parameters considered are involved in the determination of the leaking and radiating characteristics of the metasurface, highlighting that even the substrate thickness affects it considerably. It is a fact that some parameters have a higher impact on it than others. A straightforward (and logical) conclusion of this study is that modifying an isolated geometrical parameter does not provide isolated control on the phase constant or the attenuation constant. Therefore, several geometrical parameters must be simultaneously modified in order to control the phase constant, leakage, and beam direction. It can be also noted that at broadside radiation points, the attenuation constant remains relatively constant. The real impedance drops to low values while the imaginary impedance peaks negatively. Consolidating the parametric studies done, I designed a metasurface that leaks at 30° at 10.6μm. The designed metasurface has a unit cell of the following dimensions. The ZNS substrate has a thickness of 0.65μm while the patch and the ground planes are 75nm and 100nm thick respectively. The dimensions of the unit cell \( p \times q \) are \( 3.75 \times 4.3625 \mu m \).
Fig. 4.30 shows a plot of normalized phase and attenuation constant. The leakage constant is relatively level over the target range, while the phase constant varies with frequency.

As can be seen from Fig. 4.30, the fundamental spatial harmonic of the designed metasurface is fast for frequencies above approximately 25 THz. The attenuation constant is relatively constant, increasing slightly with increasing frequency before rising sharply at 40 THz. The points where $\alpha = \beta$ are critical in LWA analysis. For a bidirectional LWA, the $\beta = \alpha$ condition corresponds to a splitting condition and a broadside radiation is experienced when $\beta < \alpha$ [55, 64]. For uniform and quasi-uniform conditions, the latter corresponds to a point of maximum power.
density radiated at broadside [61]. Fig. 4.31 shows the emissivity plots within the LWIR.

Fig. 4.31 Field plots TM and TE modes

The dominant mode is the TM mode and minimal leaky behavior is noted on the TE mode. The leaking beam scans with frequency as shown in Fig. 4.32.

Fig. 4.32 Frequency scanning
From Fig. 4.32 (a), beam scanning with frequency is noted with a scanning capability of the antenna, ranging from broadside to near backfire. The surface shows a frequency scanning behavior related to its leaky wave nature. Forward leakage is noted for wavelengths of 7.5μm - 8.9μm, while backward leakage exists within wavelengths of 9.0μm - 12μm. Fig. 4.32 (b) presents the scanning capabilities of the antenna, as a function of the operating frequency. Fig. 4.32 (b) presents the scanning capabilities of the metasurface, as a function of the operating frequency. In Fig. 4.32 (b), 10.5μm is leaking at 28 degrees with half beamwidth of about 15 degrees. Plotting for a 10.6μm beam, we have Figs. 4.33 (a), (b), and (c) respectively. Fig 4.33 (a) denotes a beam leaking backwards, while a ϕ =90 degree plot of Fig 4.33 (b) shows the angle of leakage to be 30 degrees backwards. Fig 4.33 (c) is a ϕ=0 degrees plot.
The beam width of a leaky-wave antenna is controlled by its leakage constant $\alpha$, while phase constant $\beta$ controls the scan angle. Both are dependent on the geometry of the LW surface. The gain on each of the beams is almost unity, while the beamwidths are varying. This is because the normalized attenuation varies with frequency.

Since leaky wave modes belong to the improper discrete spectrum of a structure, they do not respect the radiation condition because of growing away from the structure. For this reason, these modes can describe the field on the antenna's aperture or on a surface close to the antenna and can determine the nature of the far field [58, 102, 103]. The achievable increments of gain depend on the leaky-wave pole that controls the near field distribution over the antenna aperture. In particular, for configurations with a leaky-wave pole with a small attenuation constant, it is possible to achieve extremely high gain performance but at the expense of a really small bandwidth and a very critical design. It is then important to find a tradeoff between enhancement in gain and bandwidth.
The dispersion curve in Fig. 4.34 consists of a straight line, called the light line, and dispersion plots of the spectral components. The light line denotes $\beta = \pm k$, where $k$ is the free space wave number $k$. Spectral components of the frequency range considered below the light line are bound modes. From the dispersion curve, a backward propagation wave (25-32.5 THz) is noted, while a forward leaking wave exists from 33.5-40 THz.

Fig. 4.34 Dispersion curve plot

Below 25 THz, there exist bound modes that exist as surface waves. Broadside radiation, meaning radiation perpendicular to the antenna, is noted at 34THz. At the lower frequencies belonging to the backward radiating frequency domain, the radiation beam points towards backfire, as depicted in Fig. 4.32 (a). Antennas with high directivity reduce transmit power requirements due to their narrow beams. In addition, there is a reduction of interference from other undesired
sources [88]. The beamwidth is strongly dependent on the attenuation constant $\alpha$, with low $\alpha$ leading to higher directivity and lower beamwidth [54]. To increase the frequency scanning capability of printed antennas, the permittivity of the substrate should be low [101].
Chapter 5

FINITE ELEMENT METHOD ANALYSIS OF A CIRCULAR POLARIZED METASURFACE

This chapter deals with the analysis and simulation of a circular polarized (CP) metasurface made by loading a grounded substrate with patches. In this chapter, we propose a metasurface based on the use of planar axial asymmetrical topology periodic patches to achieve circular polarization. The use of axial asymmetry topology to achieve circular polarization has been investigated by studies reported in [104, 105, 106, 107]. The common foundation of these works is the use of axial asymmetry on a planar structure to control orthogonal $E$ and $H$ fields that are particularly critical in generation of CP. In the present work, the philosophy is extended to a metasurface in the LWIR spectrum range. Stokes parameters are used to characterize the polarization at this optical range. The goal is to achieve a CP leaky wave phenomenon from a surface made up of an array of interconnected single fed microstrip patches, periodic in both longitudinal and transverse axes. This chapter starts with some background on circular polarization and its justification and then proceeds to analyzing the effects of axial asymmetry as well as dispersion properties, and both are considered for the design process. The geometrical details of the final antenna structure and the simulation results are then shown.
5.1 Circular Polarized Leaky Wave Surface

Most of the conventional LWAs are linearly polarized (LP). While it is easier to generate LP radiation, configurations of a structure with preferential CP radiation is more difficult because transverse coherence between two components of the electromagnetic field is required [108]. The key point for the generation of CP is that there should exist a phase difference of $90^0$ between the fields in the plane of polarization. Another key aspect is to control the ratio of these fields by tuning the dimensions so as to effectively take advantage of the effect of radiation from the edges [104]. The CP antenna is attractive in the deteriorated environment of the radio wave communications and is useful for sensor, radar, and mobile telecommunications [109]. It eliminates the need for polarization alignment of the transmitting and receiving antennas because CP does not require alignment of the electric field vector between the transmitter and receiver locations; hence it eliminates polarization mismatch losses [110]. A circular polarized system can enhance a wireless system capacity since research has shown that CP waves are more fade resistant than linearly polarized waves [111,112,113]. Circularly polarized microstrip antennas can be single or dual-fed type depending on the number of feeding points required in generating the circular polarized waves [113]. A LWA that can generate CP has been presented in [114]. This structure requires two LWAs and a phase shifter for generating two orthogonal polarizations that are $\pm 90^0$ out of phase. An array solution to CP generation based on dual feed network would have the disadvantages of a cumbersome feeding network accompanied by a
probable low efficiency. A single-feed circularly polarized LWA having a composite right/left handed transmission line (CRLH-TL) structure is presented in [104, 106]. Here, generation of circular polarization (CP) from the CRLH-TL is obtained by selecting a suitable position on the CRLH-TL between both edges of the ground. Other reported methods of achieving CP have been documented in [115, 116] whereby crossed slots with axial offset and dielectric image lines have been shown.

A simple and effective way to achieve CP is the use of axial asymmetrical patches. In this chapter, we have considered a single fed patch to come up with a circularly polarized metasurface. Single fed circularly polarized patches are very attractive because they can be arrayed and fed like any linearly polarized patch. However, a single-fed design generally provides narrow axial-ratio bandwidth [114]. On the other hand, CP polarized antennas provide wider bandwidth without increasing the antenna size and thickness along with greater manufacturing tolerances [117]. Naturally occurring infrared waves (thermal waves) are unpolarized [108,118] and, since circular polarization does not occur naturally, a CP thermal emission signature would be easily identified and can be filtered out from the background spectrum and be applicable for tagging operations [108, 109].

The polarization sense of a CP wave changes from RHCP to LHCP when reflected, and an antenna which can receive both senses would be significant. A circular polarized antenna can receive either a RHCP or LHCP wave and hence would suit such a situation when the direct signal is obstructed and the antenna needs to
receive the reflected signal. These are some of the merits of circularly polarized (CP) antennas that make them preferred over linearly polarized antennas.

5.2 Unit Cell Structure

The configuration of a unit cell of the proposed leaky surface is shown in Fig. 5.1 (a), together with its main geometrical parameters. As can be seen, the unit cell consists of a gold patch on top of a grounded low permittivity substrate zinc sulphide (ZNS) and a feeding microstrip line. The propagation direction is parallel to the y-axis. The patch portion has $x \times y$ dimensions of $2.157 \times 2.6 \mu m$ respectively. The dimensions of the unit cell $p \times q$ are $3.85 \times 4.3625 \mu m$, while the offset $d$ from the y axis was varied. The ZNS substrate has a thickness of $0.65 \mu m$ while the patch and the ground planes are $75 \text{nm}$ and $100 \text{nm}$ thick respectively. The material thickness was electrically thin. A thick and low dielectric constant substrate yields large bandwidth, while spurious radiation increases with substrate thickness [110].
The unit cell is periodic in the $x$ and $y$ dimensions, forming a metasurface. The patch is single fed, unlike in common arrangements where a dual feed with a $90^\circ$ phase shifter is required for CP generation. This makes the setup a very attractive solution compared to dual fed patches and other arrays, which need more complicated networks. The aim of the design was to have a CP beam scanning within the LWIR range (20-40THz). The design is such that the beam can be steered with frequency. The model setup was a driven modal in ANSYS HFSS, with material properties obtained from spectroscopic ellipsometry (IR-VASE, J.A. Woollam) [73]. In modeling, the metasurface was assumed to have an infinite array of asymmetrical patches on a grounded plane with a ZNS dielectric substrate. This was achieved by employing master slave boundaries to create an array of infinite patches.

The equivalent circuit of the unit cell model based on lattice network transmission line is given in Fig. 5.2. The patch is modeled by a two-port lattice circuit model [95,107] and transverse symmetry is assumed.
This topology was adopted because it decouples the series impedance $Z_{se}$ and shunt admittance $Y_{sh}$ under odd and even output terminal conditions respectively [119], hence having an edge over $T$ or $\pi$ networks. The series impedance is given by [95]

$$Z_{se} = R + j2L(\omega - \omega_0) \quad (5.1)$$

while the shunt admittance is given by

$$Y_{sh} = G + j2C(\omega - \omega_0) \quad (5.2)$$

where

$$R = 2Z_h^2G_{rad} \quad (5.3)$$

Fig. 5.2 Equivalent circuit of a unit cell patch [95]
Here $Z_h$ is the microstrip line impedance, $Z_l$ patch impedance, $G_{rad}$ and $G_{loss}$ conductance radiation in the patch and the microstrip line respectively. Resistance $R$ and conductance $G$ model the overall power dissipation including loss and radiation. It has been noted in [107] that the total powers in the series and shunt elements are always equal at broadside. The conductance is given by [120]

$$G = 0.00162 \left( \frac{w}{\lambda_0} \right)^{1.757}$$  \hspace{1cm} (5.7)

For $0.0033 \leq \frac{w}{\lambda_0} \leq 0.254$

where $G$ is the total radiation conductance of each patch and $w$ the width of the patch.

The formulas presented above provide design guidance and insight into the influence of the different structural parameters in the development of the specific leaky-wave antenna of interest. When designing the array, the width of the patch is varied to achieve the desired amplitude taper [120].
5.3 Parametric Studies on the Asymmetry of the Unit-cell Variation

The antenna analysis and parametric studies were performed using Ansoft HFSS based on the finite element method (FEM). The parameters that remarkably affect the antenna's behavior are the periodicity and the length of the patch as from the previous chapter. Effects of longitudinal asymmetry on the unit cell to obtain CP characteristics are studied and detailed here. The dimensions of the unit cell are considered and their effect on the characteristics of the leaky wave surface noted. In this parametrical study, the dimensions that remain constant have the following values: \( b \) (patch length) = 2.6 μm, \( a \) (patch width) = 2.157 μm, \( h \) (substrate thickness) = 0.65 μm; \( p \) (periodicity) is 3.85 μm and the asymmetry on the \( x \) axis(\( d \)) varies. A frequency sweep is also run, from 20 − 40 THz in steps of 0.5 THz.

5.3.1 Scattering parameters

To start with, a plot of return loss \( S_{11} \) with frequency with varying asymmetry \( d \) is plotted. Fig. 5.3 presents return loss with variation in asymmetry and frequency.
It can be noted that the return loss is a function of frequency with the mid-frequency range having the lowest return loss in the considered spectral range. When the asymmetry is considered, the return loss increases with an increase in the longitudinal asymmetry with $d = 0.72 \mu m$ having the highest return loss. It can be inferred that as the asymmetry is increased, coupling of power along the patch degrades. The $S_{11}$ parameters here are in reference to coupling of energy along the transmission line to the patch in the propagation direction. At large longitudinal asymmetries, the power coupling to the patch degrades appreciably. In addition, the impedance bandwidth decreases with an increase of longitudinal asymmetry. From Fig. 5.3, 0.72 $\mu m$ and 0.64 $\mu m$ asymmetries have the lowest impedance bandwidth.
while the asymmetries in the ranges of $0.1 - 0.4 \mu m$ have large impedance bandwidths.

Fig 5.4 denotes the amount of energy arriving at the output terminal. Energy propagation along the antenna is lost due to the conductor, dielectric and radiation losses. The plot of $S_{21}$ shows that at low asymmetry, the transmission coefficient is high and becomes low with increasing asymmetry. This shows that a lot more power is coupled to the leaky mode with an increase in asymmetry as the conductor, and dielectric losses are anticipated to be constant.

The observed $S_{11}$ and $S_{21}$ for $d = 0.33 \mu m$ and $d = 0.489 \mu m$ are quite low indicating a good matching and a good radiation performance. At $d = 0.33 \mu m$, the antenna presents a $-10dB$ relative bandwidth of $31.5-25/28= 23.2\%$. 
5.3.2 Dispersion diagram

In the design phase, it is important to have the dispersion diagram for the modes. The dominant design parameters are the length of the patch \((b)\), periodicity \((p)\), and the offset from the axis \((d)\). The other dimensions play a minor role. The dispersion characteristic of the FSS is shown in Fig. 5.5.
Fig. 5.5 Dispersion variation with asymmetry

The dispersion diagram is obtained considering the transmission theory in the periodic structure. Broadside radiation is around $35\ THz$ and shifts slightly as $d$ is varied with backward leaking between $25 - 35\ THz$. Below $25\ THz$, we have a bound leaky wave (surface wave). Any region below the light line is in the bounded region. The dispersion curve $\beta_p$ should be lower than that of free space for a leaky wave to occur. The dispersion curves clearly show that there exists some frequency scanning in the backward leaking wave. As $d$ is increased, a discontinuity appears on the dispersion curves. This discontinuity can be attributed to self resonance of
the patch due to the asymmetry [104]. The leaky wave surface operates in the fundamental mode and not in higher space harmonics.

5.3.3 **Leaking angle**

Fig. 5.6 shows variation in angle of leakage with frequency and asymmetry. The span angle is noted from broadside to 70° off broadside for the backward leaking wave. Frequency scanning is normally a signature characteristic of a leaky mode.

![Graph of Leaking angle variation with d and frequency](image)

**Fig. 5.6 Angle of leakage variation with asymmetry**

5.3.4 **Normalized phase constant and leakage constant.**

For convenience, the normalized phase and attenuation constants are normally included on the same graph. The attenuation constant $\alpha_y$ and the phase
constant $\beta_y$ components of the propagation constant give an accurate means of comparing the bandwidth, leakage rate, main beam direction, and approximate far-field pattern of different traveling wave antennas [54].

![Variation of Phase and Leakage constants with $d$](image)

**Fig. 5.7** Plot of normalized phase and leakage constant variation with asymmetry

The dotted line on the graph separates the bounded and leaky region. The region above the line is bounded and the region below the line is in the fast wave region hence unbounded. In some instances from the plot it can be noticed that the bounded region has a high leakage constant. This can be attributed to leakage into a surface wave. This phenomenon has been presented in [64] where printed structures have been shown to have a Brillouin diagram that shows not only regions
corresponding to leakage into space but also regions corresponding to leakage into a surface wave. Large values of the leakage constant are associated with reactive effects and with degradation of the scan performance [96].

Within the range of $25 - 31 THz$, it can be noticed that the normalized phase constant has a linear variation and the normalized leakage constant is relatively constant. At points where the leakage constant is equal to or greater than the phase constant, broadside leaking is anticipated [55,61].

5.3.4 Bloch impedance

Real Bloch impedance signifies a real power flow of waves in periodic arrays. Bloch impedance for the unit cell is plotted in Fig. 5.8. It can be noted that the real impedance increases with asymmetry. The impedance for $d = 0.33\mu m$ within the scanning range ($26 - 31 THz$) is relatively constant (varies between 45-65 $\Omega$) and therefore allows broadband matching to a constant impedance (generally 50$\Omega$) port.
The fast wave region starts from around 25 THz. Within the leaking region, the real impedance lies between 30 – 90 Ω; hence, matching network to a system impedance of $Z_o = 50 \, \Omega$ is not critical. The drastic changes in impedance can be noted to occur on the transition when the waves cross to the fast wave region and also when the backward leaking wave transitions to forward leaking. Fig. 5.9 shows imaginary impedance variation with asymmetry.
The imaginary impedance changes from positive to negative values at around 30THz. Sign change is again noted at 35THz. At these frequencies, the real part is relatively constant. The first jump is at the transition where the backward leaking mode converts to the broadside leaking mode. The second one is the transition from the broadside leaking to forward leaking wave. These transitions are more pronounced as the asymmetry is increased. A real wave impedance signifies wave propagation, whereas imaginary impedance is an indication of evanescent waves [96]. The imaginary impedance peaks at the transitions regions; that is when the wave is crossing the fast wave region or when the backward leaking is crossing to forward leaking after the broadside point.
5.3.5 Beamwidth

The beamwidth narrows and is more stable over a bigger frequency range as the asymmetry is made smaller. A useful property of leaky wave antennas is that their beam can be frequency-scanned with little beam shape deterioration over relatively large sweep angles [54].

![Beamwidth variation with asymmetry](image)

Fig. 5.10 Beamwidth variation with asymmetry

The beamwidth characteristics are affected by the leakage constant (α). A large α means that the power leaked away per unit length is big, which ends up in a short effective aperture length and a large beamwidth [55]. From the plot above, it can be noted that below 25THz, we have a bounded wave; hence a beamwidth plot
is not expected within this range. Between $26 - 32 \, THz$, we have a relatively narrow beamwidth that becomes broad with asymmetry. As the asymmetry is increased, the beamwidth increases, while the bandwidth decreases. Between $32 - 36 \, THz$, there is a rapid increase in beamwidth, and this can be attributed to an increase in the leakage constant as shown in Fig. 5.7. For a fixed aperture surface, the beamwidth is primarily dependent on the dimension of the antenna aperture while the value of $\alpha$ strongly affects the efficiency of radiation [55].

5.3.6 Field plots

Circular polarization in the model is due to the asymmetry about the longitudinal axis. Fig. 5.11 shows a vector surface plot on the $X - Z$ plane for a symmetrical case. Vector plots use arrows to illustrate the magnitude of the $x$, $y$ and $z$ components of the field. The vector field distribution in Fig. 5.11 is symmetrical for both the left and right side of the figure. The edge voltages on the top patch is given by

$$ v \approx -E_z h $$

(5)

where $h$ is the thickness of the substrate and the negative sign defines the voltage from the edge conductor to the ground conductor. Due to symmetry, the voltage potentials at the extreme patch edges are equal and there is no potential difference on the patch edge along the $x$ axis; ultimately the net scalar field quantity on the patch along the $x$ axis is zero, $E_x \approx 0$. 

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For the longitudinal asymmetrical case caused by a shift along the $x$ axis, there is unequal distribution of the $E$ vector between the two sides along the $x$ axis, as shown by the circled region in Fig. 5.12.

![Fig. 5.11 E vector of longitudinal symmetrical patch](image1)

![Fig. 5.12 E vector of longitudinal asymmetrical patch](image2)

Hence a potential difference exists between these two opposite edges and effectively $E_x$ is non-zero. For transverse axis symmetry and asymmetry ($Y - Z$ plane), $E_y$ is always nonzero due to phase variation of the travelling wave in this direction [105]. $E_x$ and $E_y$ are orthogonal complex valued field components, and if neither is zero, then CP exists. Phase quadrature between $E_x$ and $E_y$ in planar rectangular geometries has been theoretically proved in [105]. The existence of quadrature phase difference between $E_x$ and $E_y$ results in circular polarization [104,105]. By tuning the dimensions, the amplitude of $E_x$ and $E_y$ can be made to be almost equal in the far field leading to CP.

The field distributions for the dominant mode are shown in Figs. 5.13 and 5.14 with backward leaking at 28.5 $THz$. It can be noted that the $E_x$ and $E_y$,
corresponding to longitudinal and transverse electric fields respectively, are in phase quadrature in an axially asymmetrical leaky wave patch and inherently elliptically polarized [105]. Looking at Fig. 5.13, we can note that the vertical edges are at phase quadrature with each other. Similarly, the horizontal edges are in phase quadrature to each other.

![Fig. 5.13 E-field at phase](image1)

![Fig. 5.14 E-field vector plot](image2)

The phase quadrature relationship between the series and shunt radiation contributions holds for unit cells that are symmetric with respect to their transverse axis and asymmetric with respect to their longitudinal axis [105]. The $E_x$ field at far field is controlled by the degree of longitudinal asymmetry. In terms of radiation, it has been shown in [105] that the difference of the vertical edge voltages in the $y$-direction contributes to series radiation and is $y$ polarized. Similarly the edge voltage difference in the $x$-direction contributes to shunt radiation, which is $x$ polarized. The existence of $x$ and $y$ polarizations leads to circular polarization.
5.3.7 Stokes vector parameters

We have illustrated CP by plotting of Stokes vector parameters. The Stokes vector parameters give a direct physical interpretation of the state of polarization of a given EM field and the polarized emissivity is computed via calculating the power reflection coefficient using Kirchhoff’s law \[108\]. The spectral emissivity for a reflective surface is given by \[87, 108\]

\[\epsilon(\lambda, T) = A(\lambda, T) = 1 - R(\lambda, T)\] (5.8)

where \(\lambda\) is the wavelength of radiation, \(A\) and \(R\) are the absorptivity and reflectivity of the metamaterial respectively, and \(T\) is the temperature in Kelvin of the emissive structure. The Stokes parameter is given by the matrix of the form \[108\]

\[
S = \begin{bmatrix}
S_0 \\
S_1 \\
S_2 \\
S_3
\end{bmatrix} = \begin{bmatrix}
\langle \epsilon_s \rangle + \langle \epsilon_p \rangle \\
\langle \epsilon_s \rangle - \langle \epsilon_p \rangle \\
2 \langle \sqrt{\epsilon_s \cdot \epsilon_p \cdot \cos(\delta)} \rangle \\
2 \langle \sqrt{\epsilon_s \cdot \epsilon_p \cdot \sin(\delta)} \rangle
\end{bmatrix}
\] (5.9)

where \(\delta\) is the phase shift between the two orthogonal components of the polarized emissivity and \(S_0, S_1, S_2\) and \(S_3\) are the Stokes vector elements. The Stokes parameters are sufficient to characterize the magnitude and the relative phase, and hence the polarization of a wave. The Stokes parameter \(S_0\) is always equal to the total power (density) of the wave, \(S_1\) is equal to the power in the linear horizontal or vertical polarized components, \(S_2\) is equal to the power in the linearly polarized components at tilt angles \(\delta = 45^0\) or \(135^0\), and \(S_3\) is equal to the power in the left-handed and right-handed circular polarized components \[108\].
A CP antenna is orthogonal to the direction of the wave propagation and if the incident wave is CP, the handed direction of the reflected wave is inversed [121]. If any of the parameters $S_0, S_1, S_2$ or $S_3$ has a non-zero value, it indicates the presence of a polarized component in the plane wave and often the Stokes parameters are normalized by dividing by $S_0$. The Stokes vector parameters can take values of $-1, 0$ and $+1$. Fig. 5.15 illustrates a simulated variation of $TM, TE$, $S_0$, $S_1$ and $S_3$ Stokes parameters with frequency, angle, and asymmetry respectively. $S_0$ denotes the intensity of the leaked wave. It can be noted that the total intensity closely correlates with the combination of $TE$ and $TM$ for each respective case. When $d = 0$, we have a case of symmetry in the longitudinal and transverse axis. This leads to linear polarization and this is depicted by the $S_1$ plot, where the main beam has a similar trend as the $TM$ wave emissivity. When $S_1 = 1$, it depicts a linear horizontally polarized wave, and when $S_1 = -1$, it means that the wave is linear and vertically polarized. With respect to the color bar that shows blue for $-1$ and red for $+1$, the $S_1$ plot shows a linear vertically polarized wave. $S_3$ shows the plot for CP. $S_3$ can also take values of $-1, 0$ and $+1$ depicting, RHCP, unpolarized, or LHCP respectively.

From the plot, it can be noted that there exist some regions with circular polarization phenomena, but these regions are at points where minimal emitted power is occurring when considering the $S_0$ plot, which gives the power intensity on the surface. In a nutshell, there is no circular polarized emitted wave and hence the metasurface is linearly polarized.
Considering the other cases where we have longitudinal asymmetry, it can be noted from $S_1$ that the characteristics of linear polarization where power intensity is high (from $S_0$) diminish as the asymmetry is increased. Conversely, more and more regions with high power (from $S_0$) fall under circular polarization as shown in $S_3$ plots.

![Parametric study of asymmetry](image)

Fig. 5.15 Parametric study of asymmetry

The degree of axial asymmetry controls the amount of shunt radiation contribution and is on the transverse axis [107, 122]. This axial asymmetry is tuned to achieve a CP metasurface with the required leaking characteristics. From
Fig. 5.13 and Fig. 5.14, a quadrature phase relationship condition for CP can be noted. The second condition, equal field amplitudes, is represented by the TM and TE emissitivities. Though the amplitudes are not exactly captured, the trends depict similar behavior of TM and TE waves.

It can be concluded that increasing the asymmetry from $d = 0.1 - 0.48\mu m$ results in a substantial improvement in circular polarization. As the asymmetry is increased further, it can be noted that more regions are CP, but it can also be noted that the frequency scanning phenomena is being lost from $S_0$ plots. A further increase in asymmetry may not result in more improvement because the mid bands return loss worsens as shown in Fig. 5.3.

5.4 Designed Metasurface Characteristics

Considering the previous parametric study, a metasurface was designed to meet the following design specifications: a circular polarized wave in the LWIR ($8 - 14\mu m$), with a beam at $10.6\mu m$ leaking at $30^\circ$. A final designed metasurface is given in the following plots. From the previous parametric study in section 5.3, it has been shown that asymmetry of between $0.33\mu m$ and $0.489\mu m$ has reasonable circular polarization characteristics. Taking a $0.45\mu m$ design, the following radiation characteristics were simulated. In Fig. 5.16, the dispersion characteristics are shown. Broadside radiation is around $30THz$ and $37THz$. At broadside leaking, $\beta \approx 0$ and this frequency can be denoted as $f_0$. When the leaking frequency drops below $f_0$ i.e $< 0$, the propagating mode is directed in the backward propagation
direction [104]. From Fig. 5.16, we note a backward leaking between 25-30THz.

Below 25THz, we have a bound wave (surface wave).

![Dispersion characteristics](image)

**Fig. 5.16 Dispersion characteristics**

The normalized phase and leakage constant are shown in Fig. 5.17. The phase constant should be lower than that of free space ($\beta < k\omega$) for a leaky wave to occur. It can be noted that the normalized phase constant ($\beta$) is frequency scanned between 25-32THz. This region depicts a leaky wave region. Within this range, the leakage constant ($\alpha$) is relatively constant. Large values of the leakage constant are associated with reactive effects and with degradation of the scan performance [96]. A region where $\beta \leq \alpha$ denotes regions of broadside leaking for uniform and quasi-uniform structures [55].
Fig. 5.17 Plot of normalized phase and leakage constant with frequency

Figs. 5.18, 5.19, and 5.20 illustrate a simulated variation of $S_0$, $S_1$ and $S_3$ Stokes parameters with frequency and angle respectively. $S_0$ denotes the intensity of the leaked wave. It can be noted that the total intensity closely correlates with the frequency scanning curve of Fig. 5.17 and denotes broadside operation at 32 and 37 THz. $S_0$ has a frequency scanned beam at 10 – 12.5 $\mu$m. A patch mode is also noted at 9 $\mu$m, while in the 13.5 – 14.5 $\mu$m range, some energy attributed to the TE mode is noted.

Fig. 5.19 denotes a linear polarized wave. Linear polarization can be horizontal, $S_1 \approx 1$ or vertical $S_1 \approx -1$. In the field plot, it can be seen that LP
exists at wavelengths 12.5um and above and it is low in the other frequency ranges, though some pockets of vertical LP exist for some angles at 9 and 10 μm ranges. In Fig. 5.20, we have illustrated $S_3$ by setting a threshold of 0.5-1.0 for CP. It can be seen that the spectral range $7 - 13.5 \mu m$ is predominantly CP and within this range, the metasurface reveals a high degree of RHCP indicated by the near-unity value of $S_3$.

Fig. 5.18 Total Intensity plot $S_0$  
Fig. 5.19 Linear polarization plot $S_1$  
Fig. 5.20 Circular polarization plot $S_3$  
Fig. 5.21 Stokes parameter plot $S_3$
Linear plots of Stokes parameters normalized to $S_0$ are shown in Fig. 5.21 for a 30° incident wave. The $S_1$ plot predicts weak LP in the leaky region 26-30THz, while the $S_3$ plot depicts strong CP.

Fig. 5.22 shows frequency scanning of the leaking wave. The span angle is noted from broadside to 75° off broadside for the backward leaking wave. From the leaking angle plot it can be seen that the main beam direction is able to continuously scan from 10° – 75° as frequency is varied. Frequency scanning is a signature characteristic of a leaky mode.

In Fig. 5.23, an angular normalized $S_0$ intensity plot (emissivity) plot is shown. The results depict frequency scanning with approximate predicted angles. At 10.5um, the leaking angle is around 28°.

In this chapter, an attempt has been made to exploit the patch asymmetry to achieve circular polarization while simultaneously achieving the beam scanning advantages of a leaky wave antenna on a metasurface in the LWIR range.
Metasurfaces composed of unit cells that are symmetrical in the transverse axis and asymmetrical in the longitudinal axis are characterized by a quadrature phase relationship between their series and shunt radiation contributions [105]. This characteristic has been exploited to design a circular polarized metasurface that does not require a complex feeding network, such as a hybrid coupler for feeding. The design has considered the case of asymmetry that does not affect the resonance frequencies. Since thermal infrared from the natural applications is nonpolarized [118], a CP leaky wave could be used in imaging polarimetry, allowing tagging of objects based on their emitted state of polarization [123] since natural circular polarizers do not exist. A leaky wave condition is defined over the angular sector where most of the surface wave power is actually launched.
Chapter 6

CONCLUSIONS AND FUTURE DEVELOPMENTS

This dissertation has worked on two closely related topics. Chapter three has concentrated on phonon wave guiding. While surface plasmons is a mature area and manipulations of light at nanoscale, such as amplification and overcoming the diffraction limit, are possible \([6]\), at LWIR, surface phonons are more enhanced. The chapter has centered on obtaining characteristics of phonon waveguiding from a proposed modification of the plasmonic waveguide in embedded media developed by Oulton. The new structure allows obtaining phonon coupling by placing a wire on a low dielectric permittivity material with a metal ground. This technique is valid if the hybridization condition is met. The technique has been illustrated by obtaining and comparing waveguiding performance parameters with plasmonic waveguiding. One immediate extension of this work is to fabricate and test the performance of the waveguide at LWIR. Also, other materials that support phonon coupling (polar materials) can be used to test the waveguide performance. Another area of extension is to develop a feeding device. Light cannot be coupled directly to phonon waveguiding because of the phase mismatch. In the dissertation,
eigen modes have been applied to the waveguides and not a signal source. Another ambitious line would be to have a modulated surface phonon coupled waveguide so as to have a leaky wave phenomenon. In this way, a dielectric leaky waveguide could be designed.

Chapters four and five have concentrated on the leaky wave metasurfaces. While leaky wave theory has been known for decades, the analysis and design of planar leaky-wave antennas has received much attention in recent years, especially for applications in the millimeter-wave ranges, due their low profile, good performances, easiness in the design, and compatibility with printed circuit board technology. Several studies are currently focused on how to translate established radio wave and microwave antenna theories into the optical frequency regime. The main achievement of this dissertation in this part has been a systematic design and characterization of a leaky wave metasurface at LWIR. Leaky wave phenomena have been demonstrated for the first time at LWIR. In Chapter 4, the design of a planar leaky-wave metasurface was made by having an array of rectangular patches periodically arranged. The metasurface is linearly polarized and this metasurface can be used in an application that requires a high directive beam within the LWIR such as thermal sensors. The chapter concludes with a design to leak a 10.6 µm beam at 30 degrees, demonstrating the flexibility of the design. In Chapter 5, a circularly polarized metasurface is designed. Since a CP has inherent advantages over linear polarization, this metasurface can be used in critical applications where
the signal is low as CP enhances anti-fade characteristics. Such applications would include medical signals.

One immediate extension of this work is to design a feeding network so as to collect or feed a signal. Ideal sources have been considered so far in order to perform the theoretical studies. Implementation and measurement is another straightforward continuation of this work. To obtain metasurfaces with very wide bandwidth, it is proposed that impedance matching should be done. Applications of this research could include such novel applications as listed in [124].
REFERENCES


