INTEGRATED FEEDS FOR ELECTRONICALLY RECONFIGURABLE APERTURES

by

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A thesis submitted in conformity with the requirements for the degree of Master of Applied Science
Graduate Department of Electrical and Computer Engineering
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Abstract

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With the increasing ubiquity of wireless technology, the need for lower-profile, electronically reconfigurable, highly-directive beam-steering antennas is increasing. This thesis proposes a new electronic beam-steering antenna architecture which combines the full-space beam-steering properties of reflectarrays and transmitarrays with the low-profile feeding characteristics of leaky-wave antennas. Two designs are developed: an integrated feed reflectarray and an integrated feed transmitarray, both of which integrate a leaky-wave feed directly next to the reconfigurable aperture itself. The integrated feed transmitarray proved to be the better architecture due to its simpler design and better performance. A $6 \times 6$ element array was fabricated and experimentally verified, and full-space (both azimuth and elevation) beam-steering was demonstrated at angles up to 45 degrees off broadside. In addition to the reduction in profile, the integrated feed design enables robust fixed control of the amplitude distribution across the aperture, a characteristic not as easily attained in typical reflectarrays/transmitarrays.
For my parents,
Christine and Grant Nicholls.
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List of Acronyms

AESA  Active Electronically Scanned Array
BW   Bandwidth
FPA  Fabry-Pérot Antenna
FSS  Frequency Selective Surface
LEO  Low Earth Orbit
LNA  Low Noise Amplifier
LWA  Leaky-wave Antenna
OMT  Orthomode Transducer
PBC  Periodic Boundary Condition
PEC  Perfect Electric Conductor
PMC  Perfect Magnetic Conductor
PSS  Polarization Selective Surface
RA   Reflectarray
TA   Transmitarray
TDD  True Time-Delay
TDU  Time-Delay Unit
TE   Transverse Electric
TEM  Transverse Electromagnetic
TEN  Transverse Equivalent Network
TM   Transverse Magnetic
TRM  Transmit/Receive Module
XPL  Cross-Polarization Level
Chapter 1

Introduction

Antennas are increasingly becoming one of the most critical components in modern wireless communications systems. In the last several decades the accelerating growth of the wireless industry and the ubiquity of internet access and connectivity has exponentially increased the demand for ever greater data bandwidths. This demand has necessitated in wireless systems the use of more and more frequency bands at higher frequencies and wider bandwidths coupled with greater spectral efficiencies. The electronics inside wireless devices, benefiting from Moore’s Law, have been able to keep up with these demands while simultaneously reducing form factor, cost and improving device aesthetics. Antennas, being fundamentally limited in size by Maxwell’s equations, have not been as lucky and face increasingly greater challenges in meeting the same constraints.

One trend in antenna design is the gradual replacement of omnidirectional antennas with directive antennas, i.e. antennas which focus most of their radiated power in a single direction or beam. These antennas today are used extensively in a wide range of applications: most importantly long distance point-to-point communications (such as satellite) and precise sensing (such as RADAR). They are becoming even more useful as they allow for greater data rates through both increased received power (larger $E_b/N_0$) over larger distances as well as greater frequency re-use (spectral efficiency). Unfortunately, since directivity (and likewise gain) is proportional to effective antenna size, higher directivity antennas require larger effective areas, directly in conflict with the desire to reduce device form factor. For example, one of the most commonly used directive antennas is the parabolic reflector antenna. Colloquially referred to as dish antennas and often seen used in satellite communications, these antennas are large, conspicuous and extremely difficult to incorporate within any other object or device (such as a satellite, vehicle or handheld device).

Another trend in modern antenna design is the use of beamforming or beam-steering
antennas which are directive antennas capable of steering their beam to communicate with specific devices and/or track moving objects. This allows for increased data bandwidths in two primary ways: increasing received power by focusing power towards a specific device and reducing interference by reducing power radiated towards or received by third party devices. Without beam-steering, these performance gains could only be achieved in fixed links using fixed directive antennas. Most modern communications systems however are dynamic, with devices constantly moving and network conditions constantly changing. Reconfigurability at the antenna level, specifically beam-steering, allows the system to adapt to these changes. For example, there has been much interest lately in using large constellations of small satellites in low Earth orbit (LEO) to provide global internet coverage. As demonstrated in Fig. 1.1, these satellites are fast moving and far away, necessitating the use of highly directive beam-steering antennas both on the ground and on the satellites.

Beam-steering antennas have been in existence for decades, but are still most often implemented in the commercial sphere by mechanically rotating directive antennas. This mechanical method of beam-steering is bulky, ungainly and quite often impractical, particularly when considered for modern commercial devices requiring small form-factors and no moving parts. Electronic beam-steering is the natural successor of mechanical steering. Unfortunately, it has traditionally been extremely expensive and complicated to design and is therefore mostly used in military and defense applications (such as aircraft RADAR and missile tracking). Much more recently, there has been interesting work in the development of novel electronic beam-steering antenna architectures. While much less complex and often lower profile, these architectures suffer from their own drawbacks.
as well. The following sections will discuss these architectures and then propose a solution for a low-profile electronic beam-steering antenna which combines the merits of multiple current designs.

1.1 Electronically Steerable Directive Antennas

There exist today multiple different forms of highly directive, electronic beam-steering antennas. Several of these can be lumped into the category of array antennas. Array antennas are unique in that they are produced by distributing individual antenna elements in a one-dimensional (linear), two-dimensional (planar), or sometimes even three-dimensional array arrangement. The array as a whole can be thought of as the discrete sampling of a larger continuous aperture, and therefore like in aperture antennas, can produce large directivities and gains. Unlike in continuous aperture antennas however, arrays have increased flexibility due to their ability to control the excitation (magnitude and phase) of the array elements individually. This unique control allows arrays to perform beam synthesis, shaping and steering by exciting the entire array with specific amplitude and/or phase distributions. For example, tapering the amplitude distribution across arrays results in greatly improved side lobe levels, while phase gradients across the array result in steered pencil beams.

Even more importantly, this individual element control can easily be performed electronically, thus allowing electronic reconfiguration of the array. This is the core principle behind several of the modern electronically steerable directive antennas: phased arrays, reconfigurable reflectarrays and reconfigurable transmitarrays. Another type of reconfigurable directive antenna is the reconfigurable leaky-wave antenna. While not an array by definition, this type of antenna possesses similar properties to array antennas including large directivities and the ability to perform reconfiguration through electronic changes to the aperture magnitude and phase distribution. The following subsections will briefly discuss these antennas.

**Phased Arrays**

Fig. 1.2 shows a schematic of a two-dimensional phased array. In the context of this thesis, phased arrays include all array antennas in which reconfigurability is provided by means of a reconfigurable beamforming network. Each element in the array is fed individually, often by means of either corporate (Fig. 1.2) or series power-division networks. The magnitude and phase excitations of each element are then controlled by reconfig-
urable components embedded in the beamforming network itself. This includes phase shifters, time-delay units (TDUs), filters and sometimes low-noise or power amplifiers (LNAs/PAs). Often these components are lumped together in so-called Transmit/Receive Modules (TRMs, shown at the back-left of Fig. 1.2).

This architecture allows almost complete control of the array radiation properties. With full electronic phase and amplitude control of each element, the array can produce highly directive beams with low side lobes scannable in full space (both azimuth and elevation). They can also produce shaped beams or multiple beams. The beamforming networks however, especially for large arrays, can become prohibitively large, complex and hence both expensive and very lossy. Phased arrays are therefore commonly seen in military and defense applications where performance is more important than cost, and not seen as often in the commercial market. [1] and [2] provide a good overview of phased arrays and their applications.

Reflectarrays and Transmitarrays

Reflectarrays, as seen in Fig. 1.3, are the array equivalent of reflectors where the reflecting surface is discretized into an array of reflecting unit cells. Like reflectors, reflectarrays are fed from a feed antenna placed slightly above the aperture surface and reflect the feed radiation in such a way as to collimate it, producing a directive beam in the far-field. Unlike reflectors however, reflectarrays do not require curved or special geometric surfaces to realize the collimation. Instead the collimation is achieved by tuning the individual
element reflection phases across the array. By using microstrip patches, reflectarrays can be made completely planar, low profile and therefore very cost effective, and are beginning to see applications in low-cost, high-gain satellite communications.

Similar to phased arrays, electronic reconfigurability can be introduced to electronically tune the cell reflection phases thereby enabling beam steering. Reflectarrays offer several advantages over phased arrays however. The first and most important is the use of a spatial feed to excite the array. In large arrays the benefits of this feature are massive as it completely bypasses the need for the large power-division networks used in phased arrays and their associated size, losses and noise. The other advantage is that the phase tuning mechanism can be completely integrated within the aperture itself. Because of this, reconfigurable reflectarrays in this thesis are referred to as being reconfigurable aperture antennas, as opposed to a phased array which uses a reconfigurable beamforming network. Reflectarrays thus offer similar beam-steering performance as phased arrays, with much reduced size, cost and design/fabrication complexity.

Transmitarrays are the transmission analog to reflectarrays. The feed antenna is instead placed behind the array, and the array collimates the radiation as it passes through it, identical to a lens. Transmitarrays offer the same benefits as reflectarrays over phased arrays, with the added benefit that with the feed behind the array, it no longer blocks some of the radiation. One disadvantage of transmitarrays over reflectarrays however is the need for additional phase-tuning circuitry to achieve full phase-tuning ranges. [3] provides a good overview of reconfigurable reflectarray and transmitarray designs.

Leaky Wave and Fabry-Pérot Antennas

Leaky-wave antennas (LWAs) and the closely related Fabry-Pérot antennas (FPAs) are another high-directivity antenna design. Instead of using an external spatial feed, the feed is integrated into the body of the antenna. Feed waves then propagate along the aperture while slowly leaking radiation. Because of this integrated feeding, leaky-wave and Fabry-Pérot antennas are even lower profile then reflectarrays and transmitarrays.

Similar to reflectarrays and transmitarrays, leaky-wave and Fabry-Pérot antennas can be made reconfigurable by introducing electronic reconfigurability into the aperture itself. The fundamental reconfiguration principle is different than that of phased arrays and reflectarrays/transmitarrays however. Rather than changing individual element excitations or reflection/transmission responses, the distributed leakage mechanism (either the leaky-wave propagation constant or aperture periodicity) is tuned to adjust the phase gradient
across the array. Fig. 1.4 shows an illustration of a reconfigurable leaky-wave antenna. While leaky-wave antennas can be extremely low profile, they have a large disadvantage in that the beam can only be steered in a direction parallel with the feed wave propagation direction. For example in Fig. 1.4, a two-dimensional leaky-wave antenna fed in the center can only produce radiation which forms an open conical beam with scannable opening angle. Therefore despite having fully reconfigurable two-dimensional apertures, modern leaky-wave antennas still cannot perform full-space pencil-beam steering. [4] provides a quick overview of the state-of-the-art in low-profile electronically reconfigurable leaky-wave antennas, both one- and two-dimensional.

1.2 Motivation and Objectives

As discussed previously, electronic beam-steering antennas have traditionally been used in high-cost, high-performance wireless and sensing applications over the past many years. There is a growing interest however in designing simpler, cheaper beam-steering antennas which can be integrated in commercial systems. Phased arrays are in most cases much too costly and complex to incorporate in many systems due to their complicated and lossy beamforming networks. Reflectarrays and transmitarrays are a much more cost-effective approach, and are beginning to see use in space-based applications. Unfortunately the need for an external feed in many applications makes them impractical, as it greatly increases the occupied volume of the antenna and makes them much more obtrusive.
Figure 1.4: Reconfigurable leaky-wave antenna with a center feed and a conical beam.

Leaky-wave antennas are also a cost-effective solution and also boast incredibly small profiles. Unfortunately reconfigurable leaky-wave antennas still suffer from only being able to steer beams in certain directions, preventing them from obtaining true full-space beam steering. There is thus the need for a cost-effective, low-profile antenna capable of full-space pencil-beam steering.

This thesis will attempt to design such an antenna by combining the best aspects of the previous designs: integrated feeding as seen in leaky-wave antennas and reconfigurable array-like phase control as seen in both reflectarrays and transmitarrays. By doing so, an electronically reconfigurable aperture antenna will be designed capable of obtaining full-space pencil-beam steering but with a low-profile integrated feed. An illustration of such an antenna can be seen in Fig. 1.5. Such an antenna could be considered a hybrid between reconfigurable reflectarrays/transmitarrays and reconfigurable leaky-wave/Fabry-Pérot antennas, or could more simply be seen as an integrated feeding mechanism for reconfigurable aperture antennas such as reflectarrays and transmitarrays.

This thesis will present two designs: one using a reflectarray aperture, and another using a transmitarray aperture. It is then hoped that this technique will be extendable to other current or future reflectarray or transmitarray designs, reconfigurable or not.

The contributions of this thesis are thus succinctly:

1. Design an integrated feeding mechanism that can be used to feed reconfigurable aperture antennas including reflectarrays and transmitarrays.

2. Design new, or modify current reconfigurable aperture designs to allow incorporation of the integrated feeding mechanism.
3. Fabricate and test a prototype reconfigurable aperture antenna using a proof-of-concept integrated feed.

1.3 Thesis Outline

The second chapter of this thesis will first present the basic background theory used to design and analyze array and leaky-wave antennas. It will then use this theory to discuss the advantages and disadvantages of modern electronic beam-steering antennas of the types mentioned above and will discuss some established designs. Lastly, it will cover two more advanced modeling techniques which are used in this thesis, namely Floquet theory for infinite periodic array unit cell analysis and transverse equivalent networks (TENs) for leaky waveguide or leaky-wave antenna design.

The third chapter will present the first design of this thesis, that of an integrated feed reflectarray. It will begin by presenting the design of this antenna as an evolution of the reflectarray and folded reflectarray. It will next present the general architecture of this design, then move into the detailed design of the leaky integrated waveguide and the reconfigurable reflectarray unit cell. Finally electronic beam-steering simulation results will be presented for a one-dimensional array design, and the results will motivate the second design of this thesis.

The fourth chapter will present this second design, that of an integrated feed transmitarray. It will again start with a discussion of the general antenna architecture and then move onto the detailed design of the transmitarray unit cell and the leaky waveguide design. Next simulated beam-steering results for both one-dimensional and two-dimensional
arrays will be presented. Finally a prototype two-dimensional $6 \times 6$ element array will be fabricated and measurement results will be presented.

The final chapter will conclude by suggesting which antenna architecture is best, and discussing this thesis’ contributions as well as future work which could be pursued.
Chapter 2

Background

Chapter 1 presented an introduction to the concept of reconfigurable aperture antennas, gave examples of their applications and provided the motivation to design simpler and lower-profile feeding architectures. This chapter will now cover the fundamentals required to understand and design reconfigurable aperture antennas including array and leaky-wave antennas. The first section will cover the basic theory behind array antennas and beam steering as well as metrics used for comparing different array antennas. The second section will then discuss modern approaches to reconfigurable aperture antennas and further demonstrate the need for lower-profile full-space scanning antennas. Finally the third section will discuss modeling techniques used in this thesis to design and analyze array and leaky-wave antennas.

2.1 Background on Antenna Arrays

Most of the reconfigurable aperture antennas discussed in this thesis (phased arrays, reflectarrays and transmitarrays) can be analyzed in the context of array antennas. Leaky-wave antennas while not array antennas by definition, can still often be modeled as array antennas. In general, antennas can be made more directive, focusing power in a tighter beamwidth, by increasing their effective aperture size. Traditionally antennas with large directivities were and still are constructed using large continuous apertures such as those found in horn or parabolic reflector antennas. Another way of providing large directivities is by sampling a continuous aperture using an array of smaller antennas. These antenna arrays can achieve similar effective aperture sizes and hence directivities as fixed continuous apertures while simultaneously enabling the means to electronically reconfigure the radiation properties. This section will discuss the basics behind array antennas.
Chapter 2. Background

2.1.1 Array Theory

An antenna array is an array of individual radiating antenna elements that when located in a particular spatial arrangement and fed with particular phases and amplitudes acts to produce an antenna with characteristics different than the individual elements themselves. These characteristics are numerous, such as increased directivity, beam-steering, beam-shaping, reduced sidelobes or even multiple beams. While arrays can be designed in any geometrical arrangement, the most common and simplest to analyze is a linear array of evenly spaced elements such as that shown in Fig. 2.1. For the following discussion isotropic elements will be assumed, and the effect of different elements will be added in later.

The isotropic elements in Fig. 2.1 are spaced with an equal distance \( d \) and are fed with equal amplitudes \( I \) and a progressive phase difference between elements of \( \delta \). The resultant far-field pattern can be analyzed by taking the superposition of the individual element far-field patterns, taking into account the path difference between elements. This pattern is given the name array factor and is calculated in the following fashion

\[
AF(\theta) = \sum_{n=0}^{N-1} w_n e^{-jk_0r_n} \tag{2.1}
\]

where \( w_n \) are the complex excitations of each element (in this case magnitude \( I \) and phase \( e^{-jnd} \)), \( k_0 \) is the free-space wavenumber and \( r_n \) is the far-field distance for each element. For the case of the array in Fig. 2.1, a far-field approximation (parallel rays) can be used yielding \( r_n = r_0 - nd \sin \theta \). Including the fact that the excitation magnitudes
are constant and equal to $I$ and the phase differences are constant and equal to $\delta$, the array factor can be expressed as

$$AF(\theta) = I e^{-j k_0 r_0} \sum_{n=0}^{N-1} e^{jn(k_0 d \sin \theta - \delta)}.$$ (2.2)

With a bit of algebra, the normalized array factor for Fig. 2.1 can then conveniently be expressed, minus a phase factor, in the form

$$AF_n(\theta) = \frac{\sin(N \frac{1}{2}(k_0 d \sin \theta - \delta))}{N \sin(\frac{1}{2}(k_0 d \sin \theta - \delta))}.$$ (2.3)

This pattern is a periodic sinc function and produces maxima when the argument

$$\frac{1}{2}(k_0 d \sin \theta - \delta) = \pm m\pi$$ (2.4)

where $m = 0, 1, \ldots, \infty$. The pattern has multiple maxima within the angular range $-\pi \leq \theta \leq \pi$ for any given $\delta$ unless $d$ is restricted to be less than half a wavelength. In this case, the only maxima is then given by the case $m = 0$. Therefore, to produce a beam at an angle $\theta_b$ the phase difference between elements need simply be specified by the following formula

$$\delta_b = k_0 d \sin \theta_b$$ (2.5)

or in terms of a phase gradient

$$\beta_b = \frac{\delta_b}{d} = k_0 \sin \theta_b.$$ (2.6)

It can now be seen that to steer or scan the beam in $\theta$, only the phase gradient $\delta$ or phase constant $\beta$ need be scanned. This can be done in a fixed case in order to produce a fixed beam at a certain angle. For this thesis however, either $\delta$ or $\beta$ are made electronically reconfigurable allowing the array to electronically steer the beam.

When $d$ is greater than half a wavelength, the multiple maxima lead to grating lobes which are generally unwanted. For the case of $d$ being exactly half a wavelength, the first grating lobe will only appear when $\delta$ is scanned such that the main beam scans to $\pm 90$ degrees (endfire) at which point the grating lobe will appear at $\mp 90$ degrees (opposite endfire). Since arrays generally do not scan much beyond 60 degrees due to performance degradation for wide angles, it is often safe to use array spacings slightly larger than half a wavelength without having to contend with grating lobes.

The above derivations only apply to the case where the element excitations maintain
an exact phase difference $\delta$. In reality, coupling between array elements or incomplete control of the exact radiated phase of each element lead to phase errors which may cause $\delta$ to differ between elements. For beam-steering scenarios these phase errors generally manifest as increased side lobe levels and a degraded pattern, and therefore attempts are normally made to minimize them.

For two-dimensional arrays, which is the focus of this thesis, the formulas and derivations for the array factor are very similar except the summation in Eqn. 2.1 is now carried over elements positioned across two-dimensions and the array factor becomes a function of both $\theta$ and $\phi$. If the array is a rectangular array lattice with spacings $d_x$ and $d_y$ and progressive phase shifts of $\delta_x$ and $\delta_y$ in $\hat{x}$ and $\hat{y}$ directions respectively, the resulting array factor can be expressed as

$$AF_n(\theta, \phi) = \left[ \frac{\sin\left(\frac{M}{2} (k_0 d_x \sin \theta \cos \phi - \delta_x)\right)}{M \sin\left(\frac{1}{2} (k_0 d_x \sin \theta \cos \phi - \delta_x)\right)} \right] \left[ \frac{\sin\left(\frac{N}{2} (k_0 d_y \sin \theta \sin \phi - \delta_y)\right)}{N \sin\left(\frac{1}{2} (k_0 d_y \sin \theta \sin \phi - \delta_y)\right)} \right].$$

(2.7)

The reader is referred to [5] for a derivation.

The key point here is that the array factor is now simply the multiplication of an array factor in $\hat{x}$ and an array factor in $\hat{y}$, i.e. $AF(\theta, \phi) = AF_x(\theta, \phi)AF_y(\theta, \phi)$. When the spacings $d_x$ and $d_y$ are chosen to be close to or smaller than half a free-space wavelength (for the same reasons as before), this total array factor results in a pencil beam, so-called because it possesses a narrow beamwidth in both directions due to the larger aperture size in both directions. To then produce a single pencil beam in the direction $(\theta_b, \phi_b)$ the phase shifts must satisfy

$$\delta_{b,x} = k_0 d_x \sin \theta_b \cos \phi_b$$

(2.8a)

$$\delta_{b,y} = k_0 d_y \sin \theta_b \sin \phi_b$$

(2.8b)

Hence a steerable pencil beam can be produced by reconfiguring the phase shifts $\delta_x$ and $\delta_y$ or equivalently the phase constants $\beta_x = \frac{\delta_x}{d_x}$ and $\beta_y = \frac{\delta_y}{d_y}$.

For the above discussion only the array pattern assuming isotropic elements has been considered, but what happens if realistic antenna elements are used? When the antenna elements are identical, and assuming no significant mutual coupling between elements (spacings on the order of half a wavelength), the total pattern produced by the array is simply the array factor multiplied by the element pattern (element factor):

$$PF(\theta, \phi) = EF(\theta, \phi)AF(\theta, \phi).$$

(2.9)
This is often beneficial in that grating lobes can be further suppressed through the position of the element factor nulls at the location of grating lobe appearance (endfire). For scanning purposes, it is also often a good idea to use an element which has a relatively isotropic pattern, otherwise when the beam is scanned the beam will have a tendency to squint towards broadside (assuming the element is directive at broadside). Typical array elements consist of dipoles, patches and apertures which often possess one or two of the previous qualities. If mutual coupling does occur, exciting one element will also cause excitation of adjacent elements, leading to an element factor slightly different than the actual element factor but which can often be determined by analyzing the element within an array environment. A method for performing that analysis will be discussed shortly.

2.1.2 Array Performance Metrics

As with all antennas, the performance of an array can be measured using several metrics. Some of the most important metrics are directivity ($D$), gain ($G$), aperture efficiency ($\epsilon_{ap}$), bandwidth (BW), side lobe level (SLL) and cross-polarization level (XPL). Directivity can be computed directly from the pattern factor calculated in (2.9) via the equation

$$D(\theta, \phi) = \frac{4\pi |PF(\theta, \phi)|^2}{\int_0^{\pi} \int_0^{\pi} |PF(\theta, \phi)|^2 \sin \theta d\theta d\phi},$$

and is simply a measure of how focused the pattern is in a specific direction. The maximum possible directivity $D_{max}$ produced by any aperture antenna is related to the physical area $A_p$ of the aperture and the frequency of operation and is given by the formula

$$D_{max} = \frac{4\pi}{\lambda^2} A_p.$$

This maximum directivity is achieved when the aperture is uniformly illuminated and possesses a uniform phase distribution.

While directivity represents how directed or focused the power is in relation to an antenna isotropically radiating the power, gain represents the actual increase in radiated power density along the beam maximum with respect to the power incident upon the antenna ports. Gain is always less than the maximum directivity, and the difference is accounted for by the aperture efficiency

$$G = \epsilon_{ap} D_{max}.$$
The aperture efficiency is a combination of many different efficiencies and almost always includes the following terms:

\[ \epsilon_{ap} = \epsilon_{cd} \epsilon_x \epsilon_t \epsilon_p, \]  

(2.13)

although additional unique terms are added for specific antennas [6]. \( \epsilon_{cd} \) is the radiation efficiency which takes into account both dielectric and conductive losses which exist inside the antenna. In most antennas the difference between measured gain and measured directivity is simply the radiation efficiency

\[ G = \epsilon_{cd} D. \]  

(2.14)

For reflectarrays and transmitarrays with external feeds, spillover efficiency \( \epsilon_s \) may also exist which represents a loss of power due to the aperture not catching all of the incident power. This spillover efficiency can also exist for leaky-wave antennas where not all power is radiated by the end of the array. Blockage efficiencies \( \epsilon_b \) also exist for reflectarrays and are due to the external feed blocking a portion of the radiated power. In these cases these spillover and blockage efficiencies can be added to Eqn. 2.14. If mismatches exist at the feed of the antenna, an additional efficiency term exists which is the mismatch efficiency \( \epsilon_r \). This term will also result in a reduction of gain but is usually not included as part of the aperture efficiency. The gain with mismatch taken into account is often called the realized gain.

\( \epsilon_x \) is the polarization efficiency and represents the loss of power into polarizations orthogonal to the desired polarization. This term contributes to the difference between total gain and co-polarized gain (or total directivity and co-polarized directivity). \( \epsilon_t \) is the taper efficiency and exists for antennas which do not have uniform amplitude distributions. \( \epsilon_p \) is the phase efficiency and exists for antennas which do not have uniform phase distributions or possess phase errors. In most antennas the phase and taper efficiencies can be seen as the difference between the measured directivity and the maximum directivity (assuming uniform aperture distributions) i.e.

\[ D = \epsilon_t \epsilon_p D_{max}. \]  

(2.15)

While the array factor for Fig. 2.1 and for the two-dimensional rectangular array assumed uniform amplitude excitations, uneven amplitude excitation is actually quite common. The resulting array factor generally does not possess a closed form solution but can easily be numerically computed. Tapering of the amplitude excitation (for example
using a triangular amplitude weighting) is an effective means for reducing sidelobe levels and is commonly used with array antennas. The disadvantage to tapering is that it lowers your overall taper efficiency and hence directivity, as the effective size of the aperture is reduced.

Phase errors as mentioned before also lead to a reduction in directivity due to the phase efficiency term. While phase errors are never desired, they are often tolerated, particularly in the case of discrete phasing. In this scenario elements can only be excited by a certain number of discrete phase steps and the actual phase has to be approximated. The resulting reduction in directivity can be thought of as due to a sort of “smearing” of the beam pointing angle. Phase errors can also lead to substationally increased side lobe levels such as so called quantization lobes introduced with discretized phasing.

The bandwidth (BW) of beam-steering array antennas is often taken as the width of the frequency band over which the gain at the desired beam angle does not drop below 3 dB of the center frequency gain. Array antennas can have bandwidths varying from only a few percent to several octaves. Reconfigurable antennas typically have poorer bandwidths unless the phase shifting is linear with frequency, for example by using true-time delay phase shifters. This is much more difficult and expensive to accomplish.

Side lobe level (SLL) is the difference between the peak gain of the main beam and the peak gain of the highest side lobe. In this thesis they are described as a negative decibel value. For example, a side lobe which is 10 dB below the peak of the main beam has a SLL of $-10$ dB. Side lobes are often an issue with beam-steering antennas, since they are prone to phase errors. Finally the cross-polarization level (XPL) is the difference between the peak co-polarized gain of the main beam and the peak cross-polarized gain over the entire pattern. Like SLL, it is also described as a negative decibel quantity. Lower cross-polarization levels are always better, as it indicates greater polarization purity.

Ultimately there are many different properties of array antennas which need consideration during the design process. The most important property of arrays for this thesis is Eqn. 2.5 or Eqn. 2.8 which describe the method by which beams can be steered. How this process is performed, and how the array in general is fed can lead to substantial changes in the performance of the array including how much it can scan, how accurately it can scan and how much the quality of the pattern degrades when scanning. As will be seen in the next section, modern array antennas come in many different forms, all which make use of different feeding mechanisms for different effects. Lastly, and most relevant to the goals of this thesis, feeding array antennas is the most complicated aspect and often requires extremely large and complex beamforming networks. Reducing the size and complexity of these networks while maintaining performance is the goal of this
thesis.

2.2 Modern Array Antennas

This section will now discuss modern reconfigurable array antennas and the ways in which
they provide feeding and reconfigurability. This thesis divides reconfigurable array archi-
tectures into two primary categories: arrays with reconfigurable beamforming networks
and arrays with reconfigurable apertures. Of course antennas exist which fall slightly in
both categories or outside of these categories entirely, however, these two categories can
be used as an approximate grouping for the purpose of this thesis. Within each category
there are several architectures which will be discussed. For arrays with reconfigurable
beamforming networks, the primary architecture is the phased array and its varieties.
For arrays with reconfigurable apertures, the architectures discussed are reflectarrays,
transmitarrays and reconfigurable leaky-wave antennas including reconfigurable Fabry-
Pérot antennas. A brief description of the principle of operation for each architecture
will be included along with a survey of modern designs.

2.2.1 Arrays with Reconfigurable Beamforming Networks

Phased Arrays

Phased arrays are the traditional array antenna and also probably the most high per-
forming antenna that exists. Phased arrays are generally thought of as arrays of fixed
antenna elements which are fed using complex reconfigurable beamforming networks to
provide phasing and sometimes even amplitude control. While phased arrays with re-
configurable antenna elements exists, for the purposes of this thesis we will consider any
array which uses any sort of reconfigurable beamforming network as a phased array.

Fig. 2.2 shows a conceptual image of a one-dimensional phased array. The feeding
to each element is provided via some sort of guided power division network such as
a series or corporate microstrip network. The phasing between elements is controlled
mostly by the beamforming network. Often phased arrays make use of analog or digital
phase shifters located just before each antenna element. For wideband applications, true-
time delay (TTD) as opposed to just phase shifting is required, in which case some or
all of the reconfigurability is placed in time-delay units (TDUs) located throughout the
beamforming network. For large arrays, the number of phase shifters and TDUs increases
dramatically, so often these elements are placed further back in the power division network
in a technique referred to as subarraying.
The advantage of beamforming networks in phased arrays is that they allow almost full control of every possible antenna parameter through the use of transmit/receive modules (TRMs) including phasing, tapering, noise and even gain. The disadvantage of beamforming networks is that they can become tremendously complex, particularly as the array size scales up. Corporate microstrip power division networks for two-dimensional arrays can become prohibitively large and lossy, necessitating the need for additional amplifiers throughout the array. Phased arrays are thus typically only used in scenarios where the best performance is required and money is less of a concern, such as in defense-related applications.

Phased arrays have been around for quite some time and the literature is rife with different phased array designs for specific applications. In fact phased arrays are often considered more of a system than a specific antenna design, and it is easier to find papers discussing individual portions of the array such as the elements, the power division network, phase shifters, TDUs or amplifiers. This thesis will not delve into specific designs, however [1] and [2] provide a good review of phased array basics and applications and [7] provides in-depth coverage.
2.2.2 Arrays with Reconfigurable Apertures

The following array architectures rely on reconfigurable apertures as opposed to reconfigurable beamforming networks. This allows them to do away with the large, complex, lossy and expensive networks which phased arrays require and replace them with either spatial or guided surface wave feeds where power is distributed using spatial division techniques. For this thesis this category has been subdivided into reflectarrays, transmitarrays and reconfigurable leaky-wave and Fabry-Pérot antennas. It will be shown that while the feeding for these arrays is much less complex, the requirement for the aperture itself to be reconfigurable imposes several constraints and limitations on the designs such as narrow bandwidth or limited scan performance.

Reflectarrays

Reflectarrays can most easily be thought of as the array equivalent to a reflector antenna such as the parabolic reflector. The general principle behind any type of reflector antenna is to reflect the power from a feed antenna in such a way that all the reflected rays experience equivalent electrical path lengths (or delays). This results in highly directive beams in the far-field. Parabolic reflectors achieve this by manipulating the position of the reflection for each ray by means of a curved (parabolic) surface. Reflectarrays achieve the same effect by instead altering the local reflection phase on a surface which could be of any shape but is usually chosen to be flat. The reflectarray can be thought of as sampling the reflector’s phase delay at each point on the surface and instead representing that delay through a change in the reflectarray surface impedance.

Fig. 2.3 shows a conceptual image of a one-dimensional reflectarray. The feed for the reflectarray is now located in space above the array, several wavelengths away. Compared to phased arrays, the array elements are fed with phases corresponding to the path length between each element and the feed. In most cases for two-dimensional arrays this phase distribution is somewhat parabolic/paraboloidal across the surface and may or may not wrap multiples times over the surface. The job of the reflectarray elements is then to change that phase distribution into the desired phase distribution such as a uniform phase distribution for a broadside beam or a linear phase gradient to steer a pencil beam in a certain direction.

In the one dimensional case, as with the phased array, the phase gradient $\delta_b$ needed to produce a beam in the direction $\theta_b$ is given by Eqn. 2.5. To produce this phase gradient,
the $n^{th}$ element reflection phase need obey the following relation

$$\psi_n - k_0 r_{i,n} = n\delta_b \pm 2\pi m,$$  \hspace{1cm} (2.16)

where $r_{i,n}$ is the distance between the $n^{th}$ element and the feed and $m = 0, 1, \ldots, \infty$ is introduced to allow phase wrapping. This equation ensures that the total phase accrued up to and including reflection from each element adds up to the required phase at that position due to the phase gradient. In two dimensions this expression is more complicated but can be easily deduced using numerical ray-tracing methods. The ability to have absolute control over the radiated phase gradient is what allows reflectarrays to produce pencil beams in any direction.

While the savings in terms of the feed network complexity appear marginal comparing Fig. 2.3 to Fig. 2.2 (phased array), it should be understood that most array designs are both two-dimensional and much larger in practice. Phased arrays are often made with thousands to tens of thousands of elements, necessitating huge beamforming networks in both dimensions. In reflectarrays a single feed antenna can illuminate any aperture regardless of size simply by placing the feed further away from the antenna. Regardless of size, weight and cost savings, reflectarray designs also become tremendously much easier to scale.

The original concept for the reflectarray was developed in the 1960’s with the work of
Barry et al. [8] using shorted waveguides as the reflectarray elements. It wasn’t until later with the work of Huang [9] that reflectarrays were designed using more planar structures such as microstrip. These days most reflectarrays are designed using microstrip as it offers significant cost and fabrication advantages over other elements such as waveguide or other non-planar antennas. Combined with the fact that beamforming networks are not required, the ability of reflectarrays to use microstrip elements makes them one of the cheapest, lightest and thinnest array architectures in existence.

Another significant advantage of reflectarrays over traditional reflectors is that they can easily be made electronically reconfigurable through incorporated tuning mechanisms. The reader is referred to [3] for an excellent in-depth review of modern and future reconfigurable reflectarray technologies. A brief summary of these technologies will be provided here. With all methods, electronic control of the radiated phase from each element allows full electronic control of the radiated phase gradient and enables full-space pencil-beam steering.

There are two basic approaches on how to incorporate reconfiguration in reflectarray designs. The first is to use tuned coupled resonances to provide the phase shift. This can be accomplished in many ways, one such way being to tune the resonant length of microstrip patches using devices such as varactor diodes [10], PIN diodes [11], RF Microelectromechanical systems (MEMS) [12], liquid crystal [13] or ferro-electric films [14]. The surface could also be treated as an impedance surface in which the overall impedance could be tuned using varactor diodes [15]. Another approach is to couple power from the aperture to a transmission line which is then recoupled back to the aperture. Phase shifters, TDUs and even amplifiers can be incorporated along the transmission line to institute phase shifts, time delays and amplification if required. Again there are multiple methods of providing phase shifts along transmission lines through the use of varactor diodes [16], PIN diodes [17], RF MEMS [18] or now even MMICs [19]. In fact, as shown in [19], entire integrated TRMs can be placed behind the reflectarray elements pushing the designs closer to phased arrays.

Ideally these reconfiguration methods have linear responses with frequency such that their frequency response is wideband in nature. The difficulty in reality is that most antenna elements used for reflectarrays and the reconfiguration methods used to tune them are narrowband. Even with coupled-line designs using TTD lines, the microstrip patches used to couple power are inherently narrowband resulting in narrowband beam steering. Therefore, in most designs it is sufficient to have simply 360 degree phase agility for each element at the frequency of interest. In large reflectarrays this phase range is often relaxed further in order to reduce the required biasing circuitry. This can be accomplished
using 1 bit phase shifters (i.e. switches) but comes at the price of reduced directivity and increased side lobe level. Another limitation with reflectarrays are that the reflection magnitudes and loss are often not flat with tuning, leading to unwanted aperture distributions, increased side lobes and poor aperture efficiencies. Additional downsides to reflectarrays include both feed blockage and spillover which, as with standard reflectors, lead to further reduced aperture efficiencies and hence directivities.

The largest advantage of reflectarrays as mentioned before is the fact that the power dividing feed network is replaced by a spatial feed originating from a feed antenna while maintaining absolute control of the radiate phase across the aperture. In most cases this spatial feed is beneficial as it simplifies the array design immensely. However, there exist certain applications in which the external feed is not an adequate solution such as when the antenna is required to be conformal to the surface of a car, plane or building etc. Designs have been developed to overcome this issue where the feed is actually placed in the aperture itself and the power is reflected first off a reflecting surface located just above the aperture. These designs are so-called folded reflectarrays and a conceptual image can be seen in Fig. 2.4. Most often these designs utilize polarization selective surfaces (PSSs) for the reflecting surface and can be both fixed \cite{20} and reconfigurable \cite{19}, \cite{21}, \cite{22}. While folded reflectarrays do achieve an overall planar structure, the total volume occupied by the antenna is still quite large since the reflecting surface is typically still on the order of many wavelengths away from the aperture.
Transmitarrays

If reflectarrays are thought of as the array equivalent of reflectors, transmitarrays can be thought of as the array equivalent of lenses. Lenses serve to focus or collimate radiation incident on one side of the lens producing a directed pencil beam on the other side. Lenses achieve this by altering the electrical path length seen by rays travelling through the structure, often by means of a dielectric material and varying material thickness. Transmitarrays achieve this electrical path change by introducing circuits which either add electrical delay or adjust the transmission phase through the array.

Fig. 2.5 shows a one-dimensional version of a transmitarray. Comparing this to Fig. 2.3, it is obvious that the beamforming behaviour is identical to that of reflectarrays, except the power is now incident on the opposite side of the array. Therefore Eqn. 2.16 still applies and transmitarrays are fully capable of producing pencil beams directed anywhere in full-space. While both arrays have to contend with losses in the array, one significant difference between reflectarrays and transmitarrays are that transmitarrays have to additionally contend with reflections from the array surface leading to less total power radiated. Hence transmitarrays also need to be designed with good input matching by providing a free-space input impedance on the input side. Transmitarrays have an advantage over reflectarrays in that since the feed is behind the array, they do not suffer from feed-blockage effects.

While the concept of lens antennas has existed for quite some time, transmitarrays, particularly planar transmitarrays were not fully studied until around 1996 [23]. A lot of the development of transmitarrays actually stemmed from work on frequency selective surfaces (FSSs) [24]. Work on reconfigurable transmitarrays is not as widespread and much more recent than work on reconfigurable reflectarrays. As mentioned before, [3] provides a good review of both reconfigurable transmitarray and reflectarrays designs in both the past and more recently. A brief discussion will be included here.

Similar to reflectarrays, there are two major approaches towards realizing the phase shifting unit cells of reconfigurable transmitarrays. The first approach is using tuned resonators. Unlike reflectarrays, these tuned resonator structures are often required to be more complex in order to institute full 360 degree phase tunability. This can be understood by viewing a resonator system in light of the poles in its frequency response. For reflectarrays, typically only single-pole responses are required as the inclusion of the ground plane requires waves to traverse the structure twice, effectively multiplying the number of poles and allowing a maximum adjustable phase range of up to 360 degrees. In transmitarrays the waves traverse the structure only once, which necessitates multiple pole designs. Transmitarray structures utilizing the tuned resonator approach are
Chapter 2. Background

Figure 2.5: One-dimensional cross-section of a transmitarray phased to produce a pencil beam at an angle $\theta_b$ from broadside.

Transmitarrays suffer from a lot of the same problems which reflectarrays suffer, in particular narrowband response, transmission magnitudes being coupled with phase tuning and difficulties in producing full continuous 360 degree phase ranges. In fact the phase issue with transmitarrays is more pronounced since multiple pole systems are required with tuned resonances, hence the recommendation for guided-wave transmitarray architectures. Despite these shortcomings, transmitarrays are still highly advantageous
Figure 2.6: One-dimensional cross-section of a leaky wave antenna with a phase constant $\beta$ producing a pencil beam at an angle $\theta_b$ from broadside.

over phased arrays in many applications, particularly when it comes to cost, size and weight due to the removal of a complicated beamforming network. That said, there are certain applications where the requirement for a feed antenna would be problematic, such as mounting on the roof of a car or on a plane, as the extra empty volume required increases the antenna profile.

Reconfigurable Leaky Wave and Fabry-Pérot Antennas

Another style of reconfigurable aperture antenna is that of the reconfigurable leaky-wave antenna or Fabry-Pérot antenna. These antennas are different in that they are often not formed by a discrete array of elements, but a more continuous or periodic aperture. They are also fed using guided or surface waves which propagate along or directly underneath the surface of the aperture itself, producing a natural phase gradient. Reconfigurability is granted by reconfiguration of the aperture or the way in which the feed wave propagates across the aperture. Because of this they are similar to reflectarrays and transmitarrays in that they do not require complex beamforming networks. Fig. 2.6 demonstrates the concept of a reconfigurable leaky-wave antenna.

Ignoring reconfigurability at first, the mechanism by which leaky-wave antennas produce directive radiation is through the leakage of power as the feed wave propagates along the antenna. The feed wave propagates with a phase constant $\beta$ as indicated in Fig. 2.6. The feed wave can be either fast, $\beta < k_0$, or slow, $\beta > k_0$ which is determined by the characteristics of the leaky waveguide. If the wave is fast, and assuming there is an opening between the leaky waveguide and free space, phase matching at this interface will lead to a vertical wave component $k_z = \sqrt{k_0^2 - \beta^2}$. This component along with the
horizontal component results in a beam at a corresponding angle \( \theta_b \) given by

\[
\sin \theta_b = \frac{\beta}{k_0}.
\] (2.17)

Fast-waves are said to “leak” radiation, hence the term leaky-wave antenna. Because of this leakage, the leaky propagation constant actually consists of a phase constant \( \beta \) and a leakage constant \( \alpha \). How large \( \alpha \) is determines how fast the radiation leaks, and will therefore generally control the effective size of the aperture and hence the directivity/beamwidth.

If the phase constant is slow, phase matching at the interface between guide and free-space produces evanescent modes above the guide, and the wave is essentially contained within the waveguide. Fortunately, waveguides with slow waves can be made leaky by introducing periodic perturbations in the interface between guide and free space. These periodic perturbations introduce discontinuities which enable radiation. More rigorously, periodic perturbations enable in free-space just above the aperture an infinite series of spatial modes called Floquet harmonics. These Floquet harmonics are produced by periodic phase constants related to the fundamental waveguide phase constant \( \beta_0 \) through

\[
\beta_n = \beta_0 + \frac{2\pi n}{d}.
\] (2.18)

In order to generate radiation, one of the Floquet modes must be made fast in order to satisfy the phase matching condition in Eqn. 2.17. In other words, \(-k_0 < \beta_n < k_0\) for some integer \(-\infty < n < \infty\). Most commonly the periodicity \( d \) is chosen appropriately to allow the \( n = -1 \) Floquet mode and only that mode to become fast.

The advantage of the periodic leaky-wave antenna using the \( n = -1 \) Floquet mode is obvious once scanning in introduced. First off, leaky-wave antennas are naturally beam-steering antennas due to the dispersive nature of the waveguide which feeds them. As can be seen in Eqn. 2.17, when the frequency is changed the phase constant \( \beta \) will vary nonlinearly with frequency (whilst \( k_0 \) varies linearly with frequency), thus causing \( \theta_b \) to scan. For almost all waveguides, the fundamental waveguide phase constant \( \beta_0 \) is always greater than zero, corresponding to a forward traveling wave. In this case the uniform or quasi-uniform leaky-wave antenna can only have a forward pointing beam (ie. positive \( \theta_b \)). The periodic leaky-wave antenna, while having a positive fundamental phase constant, can use the \( n = -1 \) harmonic to produce an effective phase constant which is negative, producing a backwards pointing beam. At this point it should be mentioned that with all leaky-wave antennas, the beam-pointing direction (boresight)
lies in a plane formed by the direction the leaky-wave is propagating and the normal to the aperture surface. A leaky-wave antenna cannot scan the beam orthogonal/transverse to the leaky-wave propagation direction. So for Fig. 2.6, the beam can only be scanned in the plane of the page, and never into or out of the page.

In addition to the phase constant, attention during design must also be payed to the leakage constant $\alpha$. This parameter for the most part controls the rate at which power is leaked from the surface, and therefore controls the effective aperture size. Generally leaky-wave antennas are designed such that they radiate at least 90% of the power by the end of the array. For an array of length $L$, the power remaining at the end of the antenna with a constant $\alpha$ is $e^{-2\alpha L}$. Therefore, the requirement for greater than 90% leakage necessitates $e^{-2\alpha L} \leq 0.1$ and the requirement on $\alpha$ becomes:

$$\alpha \geq \frac{\ln 10}{2L}.$$  \hspace{1cm} (2.19)

Despite a fundamentally different radiation method than the previously described array antennas, most leaky-wave antennas can still be analyzed from an array perspective, particularly when made periodic. The only difference is that there is now a fundamental amplitude and phase tapering across the array, given by the leakage and phase constants respectively.

Leaky-wave antennas did not receive a lot of attention until the 1950’s at which point multiple different leaky designs were investigated [33]. One design in particular was the partially reflective sheet array [34]. This design is slightly different than the leaky-wave antenna concept previously discussed. This antenna in fact represents another class of antennas called Fabry-Pérot or partially reflecting surface antennas. The fundamental operational principle is in fact so similar that in many cases (as in this thesis) they are lumped under the name of leaky-wave antennas. The sole difference in their operation is that instead of a traveling wave, the feed wave is seen more as a bouncing wave between a fully reflecting surface and a partially reflecting surface (similar to a Fabry-Pérot cavity).

While frequency scanning antennas are used in a variety of applications, fixed frequency scanning is a much more desirable trait for many applications. Leaky-wave antennas can be made electronically reconfigurable at a fixed-frequency in two primary approaches, both of which are obvious through the viewing of Eqn. 2.18. In the first approach, the leaky wave antenna is either a uniform or periodic leaky-wave antenna and the fundamental phase constant $\beta_0$ is made reconfigurable. One way this approach has been achieved in the past is by the use of dielectric-image lines feeding aperture-coupled microstrip patches [35]. By adjusting the height of a movable reflector located near the
dielectric-image line, the effective phase constant was changed allowing the antenna to scan by roughly 20 degrees in the forward direction.

More recently however have been designs in both waveguide [36] and substrate-integrated waveguide (SIW) [37] which use varactor tuned high-impedance surfaces to change the effective phase constant in the guide. These designs again both achieved scan ranges of roughly 20 degrees in the forward direction. To account for the limited scan range of these designs modifications were made to provide scanning through broadside into the backward direction for a total scan range of ±25 degrees [38]. This design is fed from the center, using one side of the array to scan in the forward, the other side in the backward, and a resonance condition to produce a broadside beam. The limitations are obvious with this design in that only half of the array is ever used at one time, drastically reducing the aperture efficiency. Two-dimensional variations of these designs have also been proposed [4] but still only dedicate certain sections of the array to beams in certain directions, resulting in significant reductions in aperture efficiency.

The introduction of metamaterial antennas has allowed variable phase constant designs to achieve forward and backward scanning. One of the first such designs [39] uses a composite right/left-handed (CRLH) transmission line loaded with tunable varactors to shift the dispersion curve up and down in frequency. This allows the fundamental phase constant to be shifted from negative to positive values, and scanning from −50 to +50 degrees can be achieved. Another more recent design utilizing half-mode SIW [40] works with a similar principle and can achieve similar scan ranges. In [41] a forward scanning CRLH antenna is switched from being fed from either end to produce full-space scanning. This method has advantages over the backward to forward reconfigurable design in that it requires less extreme values for tuning capacitances thus requiring less exotic varactor diodes.

In the second approach, the leaky-wave antenna is periodic and the periodicity $d$ is made adjustable. In this approach, the $n = -1$ Floquet harmonic is used and the phase constant can be made to swing from $-k_0$ to $+k_0$, effectively allowing full scanning in both forward and backward directions. There are many different ways in which the reconfigurable periodicity can be produced; using PIN diodes [42], [43] or even a reconfigurable water grating [44] are a couple examples. While full-space scanning can be achieved, the disadvantage with these switched designs is that only discrete periodicities can be achieved, leading to discrete scanning. Another design by Sievenpiper [45] represents a sort of hybrid between the previous approach and this approach, where the phase constant is tuned but is also made periodic to allow backwards radiation.

In all of these designs, as in the reflectarrays and transmitarrays presented before,
the complicated beamforming networks are replaced with relatively low profile, reconfigurable apertures. The advantage of reconfigurable leaky-wave antennas over reflectarrays/transmitarrays is that the feed is entirely integrated within the aperture itself (ie. traveling waves). Therefore these designs are extremely low profile. The disadvantage however is that achieving even full one-dimensional scan ranges can be difficult. Almost all of the designs mentioned above are linear arrays. As mentioned in Chapter 1, two-dimensional leaky wave antennas do exist but will tend to produce conical beams, since when fed from the center the wave propagates outward. In [4] two-dimensional scannable arrays are proposed, but are formed from an array of one-dimensional scanning arrays arranged such that each array produces a beam in specific quadrant or sector. These methods drastically reduce the aperture efficiency since they only use a small part of the aperture at one time.

This Thesis

Review of the current reconfigurable array antennas reveals the advantages and disadvantages for many different architectures. To summarize, phased arrays provide full-space pencil beam steering but require extremely complex, large and costly beamforming networks. Reflectarrays and transmitarrays can do away with these beamforming networks by using spatial feeding, and integrating all of the phase control in relatively low-profile reconfigurable apertures. Unfortunately they still require external feeds which in many applications is unwanted due to the increase in volume. Reconfigurable leaky-wave/Fabry-Pérot antennas are great in that they contain all of the feeding and reconfigurability in a relatively low-profile package. Unfortunately they suffer in that it is difficult for them to achieve full-space beam steering. This thesis attempts to bridge the gap between reflectarrays/transmitarrays and leaky-wave/Fabry-Pérot antennas by designing a hybrid architecture which uses an entirely integrated feed such as in leaky-wave/Fabry-Pérot antennas but uses similar reconfiguration methodologies as reflectarrays/transmitarrays to produce full-space beam steering.

2.3 Array Modeling and Analysis Techniques

Throughout this thesis multiple different techniques will be used to design, model, simulate and analyze arrays. This section will briefly introduce these techniques and their theory such that they can be applied in the following chapters.
2.3.1 Infinite Periodic Array Theory

Rather than simulating the entire array, it is computationally much more efficient to simulate a single unit cell of an array (i.e. one reflectarray element). Elements however behave differently in an array environment than by themselves due to mutual coupling between cells which is also dependent on the phasing or tuning between cells. Therefore a method to simulate the unit cell of an array while taking into account coupling effects and the phasing between unit cells is required. The easiest way to do this is to simulate the unit cell in an infinite array environment.

Infinite Periodic Arrays

Before discussing how elements are modeled in infinite periodic array environments, a brief discussion of infinite periodic arrays will be performed. Consider the infinite periodic array in Fig. 2.7 with periodicities \( d_x \) in \( \hat{x} \) and \( d_y \) in \( \hat{y} \). Each element has a \( \hat{y} \)-directed current distribution given by \( f(x, y) \) and the phasing between elements is given by \( k_{x0} d_x \) in \( \hat{x} \) and \( k_{y0} d_y \) in \( \hat{y} \) (phase gradients of \( k_{x0} \) and \( k_{y0} \) respectively). The elements extend to infinity in all directions. It can be shown [46] that the resulting magnetic vector potential produced by the array has the following form

\[
A_y(r) = -j\mu \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \tilde{f}(k_{xm}, k_{yn}) \frac{e^{-jk_{zmn}|z|}}{2d_x d_y k_{zmn}} e^{-j(k_{xm}x + k_{yn}y)}
\]  

(2.20)

where

\[
k_{xm} = k_{x0} + \frac{2\pi m}{d_x}
\]

(2.21a)

\[
k_{yn} = k_{y0} + \frac{2\pi n}{d_y}
\]

(2.21b)

\[
k_{zmn} = \sqrt{k_0^2 - k_{xm}^2 - k_{yn}^2}.
\]

(2.21c)

and \( \tilde{f}(k_x, k_y) \) represents the fourier transform of \( f(x, y) \). The reader is referred to [46] for an excellent derivation and discussion of this result. Similar expressions can also be obtained for magnetic vector potential components due to \( \hat{x} \)- and \( \hat{z} \)-directed currents on the elements.

The primary take away from Eqn. 2.20 is that for a periodic infinite array, the fields are given entirely by an infinite summation of distinct plane waves called Floquet modes. Eqn. 2.20 is hence often given the name Floquet expansion or spectral expansion. Each Floquet mode propagates in a unique direction \( k_{xm}\hat{x} + k_{yn}\hat{y} \pm k_{zmn}\hat{z} \) with unique \( \theta \) and
Figure 2.7: Infinite periodic array of elements with $\hat{y}$-directed current distributions $f(x, y)$.

$\phi$ values given by

$$k_{xm} = k_0 \sin \theta_{mn} \cos \phi_{mn}$$  \hspace{1cm} (2.22a)
$$k_{yn} = k_0 \sin \theta_{mn} \sin \phi_{mn}$$  \hspace{1cm} (2.22b)
$$k_{zmn} = k_0 \cos \theta_{mn}.$$  \hspace{1cm} (2.22c)

Not all of these modes are propagating. If $k_{xm}^2 + k_{yn}^2 > k_0$ then $k_{zmn}$ is imaginary and the mode is evanescent. In fact, for most arrays the periodicities $d_x$ and $d_y$ are chosen to be less than half a wavelength such that only the fundamental or dominant mode with $m = n = 0$ is propagating. This is related to the choice of periodicity discussed earlier in order to avoid grating lobes. The grating lobes are in fact equivalent to Floquet modes when a finite array is extended towards infinity. As the size of an array is increased, the beamwidth of the main and grating lobes becomes narrower and eventually approaches a delta function (Floquet mode) for an infinite array. Also note that each Floquet wave may have both a $TM$ and $TE$ polarization.

**Modeling Unit Cells in Infinite Periodic Arrays**

The benefit of this analysis can be seen by considering the form the Floquet modes take in each unit cell. By shifting the magnetic vector potential by one periodicity in either of
Figure 2.8: Comparison between fields in an infinite periodic array and an array unit cell located in a waveguide with periodic boundary conditions (PBCs). The fields are identical in both cases.

the principal array directions, the resulting fields can be related to the initial fields by:

\[ A_y(x + a, y, z) = A_y(x, y, z) e^{-jk_{xm}a} \]  \hspace{1cm} (2.23a)

\[ A_y(x, y + b, z) = A_y(x, y, z) e^{-jk_{yn}b}. \]  \hspace{1cm} (2.23b)

Thus the fields between unit cells (or the fields on opposite boundaries of the unit cell) differ only by a phase shift given by \( k_{xm}a \) in \( \hat{x} \) and \( k_{yn}b \) in \( \hat{y} \). Hence if the analysis is restricted to a single unit cell with \( 0 \leq x \leq a \) and \( 0 \leq y \leq b \) the fields in this region are identical to the fields produced by an infinite periodic array assuming the proper periodic boundary conditions are enforced. These so-called periodic boundary conditions (PBCs) are given in Eqn. 2.23 with the additional conditions that the fields are continuous and smooth across the boundary. Fig. 2.8 demonstrates this behaviour for a one-dimensional cross-section of an infinite periodic array.

This behaviour can now be exploited to simulate the unit cells of arrays (Fig. 2.9). The unit cell is placed in a waveguide with dimensions the size of the unit cell and the waveguide walls are assigned as periodic boundary conditions. In Ansys HFSS the periodic boundary conditions are master/slave boundaries and the waves are excited using Floquet ports at both ends of the waveguide.

The master/slave boundaries require either the phase between boundaries \( (k_{x0}a, k_{y0}b) \) or the incidence angles \( (\theta_{inc}, \phi_{inc}) \) to be specified. When setting up the Floquet port, HFSS includes a Floquet mode calculator which uses the incidence angles (or phaseshifts) and cell periodicity to determine which modes are propagating. In most cases with a periodicity close to half a wavelength and moderate incidence angles only the fundamental Floquet modes are propagating and only these modes are simulated. For a set-up as shown in Fig. 2.9, with only the fundamental modes propagating, there are four modes
Figure 2.9: Infinite array simulation setup for a generic array unit cell using Master/Slave PBCs and Floquet waveports. Only the fundamental Floquet modes are presented along with their polarization directions, yielding a four-port simulation.

In total, fundamental $TE$ and $TM$ modes at both ends of the waveguide. The entire simulation can then be treated simply as a four port device, with two ports denoting one polarization and two ports denoting the other. For initial simulations, the incidence angle $\theta_{inc}$ is usually chosen as broadside for simplification. For later simulations, the incidence angle is usually chosen as the actual incidence angle expected in the array. Examples of these simulations will be given in the following two chapters.

2.3.2 Transverse Equivalent Network Modeling

Another modeling technique used commonly in the design of leaky-wave antennas [36] is the transverse equivalent network (TEN) [47]. This technique is an analytical method for finding the propagation constant of complicated waveguiding structures or the resonant frequency of complicated resonant cavities. For instance, in leaky-wave antennas it is common to have a waveguide with a leaky top surface which is a complicated arrangement of dielectric surfaces and printed metallic patterns. Rather than simulating the entire waveguide in a commercial simulator, or trying to find the complete analytical field solution, only a small portion (such as a unit cell) of the surface itself is simulated. The impedance or scattering parameters from this simulation are then entered into a transverse equivalent network model to find the propagation constant ($\gamma = \alpha + j\beta$) using a single transcendental equation.

To derive this modeling technique, first consider the lossless air-filled parallel-plate waveguide cross-section with height $h$ shown in Fig. 2.10(a). The $TE_1$ mode electric field
Figure 2.10: (a) TE mode in parallel-plate waveguide seen as bouncing wave and (b) corresponding TEN model.

can be found to be [47]

\[ E_y = \frac{j\omega\mu_0}{k_c} A \sin \frac{\pi z}{h} e^{-j\beta x} \]  \hfill (2.24)

where \( A \) is a constant based on the excitation amplitude,

\[ \beta = \sqrt{k_o^2 - k_c^2} \]  \hfill (2.25)

is the longitudinal propagation constant within the waveguide and

\[ k_c = \sqrt{k_y^2 + k_z^2} \]  \hfill (2.26)

is the fixed cutoff wavenumber which for this mode is equal to \( \frac{\pi}{h} \). It is easy to show that this wave can be expressed as

\[ E_y = \frac{\omega\mu_0}{2k_c} A \left[ e^{-j \left( \frac{\pi z}{h} + \beta x \right)} + e^{-j \left( \frac{\pi z}{h} - \beta x \right)} \right] \]  \hfill (2.27)

which is in the form of two planes wave traveling at oblique angles to one another with wavevectors

\[ \vec{k}_a = -\frac{\pi}{h} \hat{z} + \beta \hat{x}, \]  \hfill (2.28a)

\[ \vec{k}_b = +\frac{\pi}{h} \hat{z} + \beta \hat{x}. \]  \hfill (2.28b)

This mode can thus be interpreted as two bouncing waves which bounce off the top and bottom walls at an angle \( \theta_{bmc} \) given by

\[ \sin \theta_{bmc} = \frac{\beta}{k_0}. \]  \hfill (2.29)
As the frequency is decreased, and since the geometry is fixed, $\beta$ as seen in Eqn. 2.25 will slowly decrease to zero. This will cause the bounce angle to correspondingly go to zero (since $k_c$ is fixed). At this exact point the waveguide is in cutoff ($\beta = 0, k_z = k_c$) and the waves are seen to be bouncing directly up and down. This suggests that the waveguide can be modeled as a vertical transmission line with propagation only in the transverse (vertical) direction. Moreover, because the waves are bouncing back and forth in the transverse direction, the vertical waveguide is in resonance, and this point is called the transverse resonance condition.

The waveguide at cutoff can thus be replaced with a vertical transmission line terminated at either end by arbitrary impedances, in this case short circuits (Fig. 2.10(b)). At resonance, the currents and voltages at every point along the line must be equal. If these currents and voltages are sampled at an arbitrary point along the line, the impedances looking both up and down must be

$$Z^\uparrow = \frac{V^\uparrow}{I^\uparrow}, \quad (2.30a)$$
$$Z^\downarrow = \frac{V^\downarrow}{-I^\downarrow}. \quad (2.30b)$$

Since $V^\uparrow = V^\downarrow$ and $I^\uparrow = I^\downarrow$, the impedances must therefore satisfy

$$Z^\uparrow = -Z^\downarrow. \quad (2.31)$$

This is the transverse resonance equation and it is a transcendental equation in the cutoff wavenumber $k_c$. By treating the waveguide as a vertical transmission line, the impedances looking both up and down inside the guide can be computed via translation from the load with $k_c$ and used in Eqn. 2.31 to solve for the cutoff wavenumber. With the cutoff wavenumber known, the propagation constant $\beta$, or in the lossy case $\gamma = \alpha + j\beta$, at the frequency of interest can be found using Eqn. 2.25 or

$$\gamma = j\sqrt{k_0^2 - k_c^2} \quad (2.32)$$

for more general lossy or leaky waveguides.

As an example, Fig. 2.11(a) shows a basic dielectric filled waveguide structure between a metal ground plane and a leaky surface which can be modeled as a generalized impedance surface. It is also assumed that the propagating mode is $TE$ and possesses no field variation in the $\hat{y}$ direction such that $k_c = k_{z,g}$, the waveguide transverse phase constant. The guided wave will propagate along the structure with a complex propagation
constant \( \gamma = \alpha + j\beta \) which is unknown and also difficult to analytically solve for.

To determine the propagation constant, the waveguide can be modeled as a transverse equivalent network as shown in Fig. 2.11(b). The vertical phase constant is the cutoff phase constant \( k_v \) which is equal to \( k_{z,0} \) in the air region and \( k_{z,g} = \epsilon_g k_{z,0} \) in the waveguide region. The TE mode inside the guide has a vertical impedance \( Z_{TE,g} = \frac{\omega \mu_0}{k_{z,g}} \) while the mode in free space outside of the guide has an impedance \( Z_{TE,0} = \frac{\omega \mu_0}{k_{z,0}} \). The leaky surface has a vertical TE impedance \( Z_{TE,s} \) which would have to be determined either analytically or via simulation. For simplicity, a point is chosen just below the leaky surface to implement the transverse resonance equation

\[
Z_{TE,\uparrow} + Z_{TE,\downarrow} = 0. \tag{2.33}
\]

In this case \( Z_{TE,\downarrow} \) is given by a length \( h \) of shorted transmission line with phase constant \( k_{z,g} \) and can be found using standard transmission line formulas to be

\[
Z_{TE,\downarrow} = jZ_{TE,g} \tan(k_{z,g}h). \tag{2.34}
\]

\( Z_{TE,\uparrow} \) is given by the parallel combination of the free-space impedance \( Z_{TE,rad} = Z_{TE,0} \) and the surface impedance \( Z_{TE,s} \). Hence the transverse resonance equation for this waveguide is

\[
jZ_{TE,g} \tan(k_{z,g}h) + \frac{Z_{TE,s} \sqrt{\frac{\omega \mu_0}{k_{z,0}}}}{Z_{TE,s} + \sqrt{\frac{\omega \mu_0}{k_{z,0}}}} = 0. \tag{2.35}
\]

The solution to this equation (if it exists) can be found graphically by plotting the
equation as a function of $k_c$ over the complex space and finding its zeroes. Once $k_c$ is found it can be converted to a propagation constant via Eqn. 2.32. This method will be used to design leaky feed waveguides in the following chapters.
Chapter 3

Integrated Feed Reflectarray

As revealed in Chapter 2, there are currently many different methods of designing electronic beam-steering antennas. Reflectarrays are particularly interesting in that they possess independent element phase control allowing full-space beam steering which is all embedded in a relatively low-profile reconfigurable aperture. The downside however is the need for a bulky external feed antenna to excite the aperture from a short distance away. Folded reflectarrays offer one solution to reduce the overall occupied volume by placing the feed antenna in the aperture itself and to use an additional reflection layer to constrain the reflected feed waves closer to the aperture surface. This chapter of this thesis proposes a new solution which is to even further constrain these waves closer to the aperture through use of an integrated feed which completely guides the feed waves along the surface of the aperture.

The first section of this chapter will present the idea for the proposed antenna based on an evolution of reflectarrays and folded reflectarrays. The second section will then present the general architecture of this design and how it relates to current reflectarray and leaky-wave antenna designs. The next two sections will then discuss in more detail the design of two critical components in this antenna, namely the polarization selective leaky feed waveguide and the reconfigurable reflectarray surface. The final section will then show simulated beam-steering results for a linear section of this array. Finally, the chapter will conclude by discussing the difficulties of this proposed solution and will then set the stage for this thesis’ second design proposed in the following chapter.

3.1 Design Introduction

The concept behind the proposed design can be most simply understood by extrapolating the evolution from normal reflectarray to folded reflectarray. For the purposes of this
discussion only fixed reflectarrays will be depicted at first, and reconfigurability will be introduced later. Folded reflectarrays (Fig. 3.1(b)) were partly designed as a means to reduce the overall thickness or volume of reflectarrays (Fig. 3.1(a)). They do this by integrating the feed antenna into the reflectarray aperture itself, and use a subreflector above the reflectarray aperture to reflect and distribute the feed waves back onto and over the aperture. By placing the subreflector halfway between the aperture and the original feed, the reflectarray thickness is effectively halved, keeping all other parameters the same.

What happens now if the subreflector is placed closer and closer to the aperture? Keeping all other design parameters the same, all feed waves will still only bounce once off the subreflector and once off the reflectarray aperture. This will cause the feed waves to be distributed over less of the total aperture area (Fig. 3.1(c)). Now instead of making the reflector surface completely transparent to the aperture reflected waves, what if it is only partially transparent? This will cause the feed waves to experience multiple bounces between subreflector and aperture, and the waves will effectively travel across the aperture surface until they are mostly radiated (Fig. 3.1(d)). In fact, this design is now almost identical to that of partially reflecting surface antennas, or as discussed in Chapter 2, leaky-wave antennas. The partially reflective subreflector when placed close to the reflectarray aperture will in combination effectively form a low-profile leaky-wave antenna, with the feed waves bouncing (propagating) between the subreflector and aperture from the feed outwards. Fig. 3.1 only shows one-dimensional versions of these arrays, but in all cases these arrays would generally be two-dimensional and could look something similar to what was shown in Figures 1.3, 1.4 and 1.5 (of course with an added subreflector).

Forming a directed beam with this antenna is more difficult than with a folded reflectarray however. In a folded reflectarray the lack of multiple reflections allows the two reflecting surfaces to be effectively decoupled. This allows ray tracing to be used to determine the required phase accrual for each ray and therefore specify precisely the required reflection phase at each point on the reflectarray aperture and subreflector. In this leaky-wave/folded reflectarray hybrid antenna however, multiple reflections between subreflector and reflectarray effectively couple the surfaces together to form a waveguide. This makes it easier to analyze the antenna’s behaviour from the perspective of a travelling wave with phase constant $\beta$. If the reflection phases are kept constant along the reflectarray and subreflector, $\beta$ would remain constant and the array could be analyzed as a leaky-wave antenna. However, if the reflection phases vary, which would be the case if a reflectarray aperture is used, $\beta$ would vary throughout the guide leading not only to
Figure 3.1: Evolution of (a) reflectarray to (b),(c) folded reflectarray to (d) integrated feed reflectarray.
issues in analysis and design, but possibly disruption in the propagation of the wave (i.e. the guide going into cutoff).

Constant reflection phase surfaces could be used for the reflectarray aperture, but this would result in essentially a leaky-wave antenna and all their associated problems with beam-steering. In particular, consider now if reconfigurability is reintroduced to the reflectarray aperture. In the leaky-wave case, to steer the beam $\beta$ must be reconfigurable, and this can be accomplished by simultaneously reconfiguring the entire reflectarray aperture reflection phase in unison. This however, as was shown in Fig. 1.4, would only allow for steerable conical beams, not pencil beams. It is therefore extremely desirable to keep the local phase tunability which reflectarrays are known for. Of course, this tunability only compounds the problems of analysis, design and propagation disruption mentioned earlier with local reflection phase variability ($\beta$ variability throughout the guide). How then can this design maintain local phase tunability without causing these problems? The solution, as will be discussed in the next section, is to physically decouple the waveguiding and phase tuning components.

### 3.2 General Architecture

Fig. 3.2 shows a cross-sectional view of the proposed integrated feed reflectarray architecture. Rather than using the reconfigurable aperture as part of the integrated feed, the aperture (RA in Fig. 3.2) is moved downwards, and another fixed layer is introduced (BL in Fig. 3.2). This fixed layer, when combined with the old partially reflective subreflector (UL in Fig. 3.2), form a fixed leaky waveguide with fixed phase constant $\beta$ and leakage constant $\alpha$. The leaky waveguide constrains all of the feed-wave within it and guides them above and across the reflectarray aperture. It is made leaky on the bottom side (BL) such that it can slowly leak power downwards towards the reflectarray aperture. The spacing between the waveguide and the aperture can be made relatively small to maximize reductions in profile.

The reconfigurable reflectarray aperture then reflects the leaked power upwards, simultaneously applying a phase gradient through individual element reflection phase tuning to steer the beam in a certain direction. Each reflectarray unit cell possesses close to 360 degrees of phase tuning range to allow any arbitrary reflected phase gradient to be formed enabling full-space beam-steering. The trick with this design is allowing the reflected power to radiate through the leaky feed waveguide. Otherwise the reflected and phased waves will be reflected back by the waveguide subsequently producing multiple reflections within the structure. This is the same problem faced before in Fig. 3.1(d)
where it becomes hard to properly analyze and control the phase of the radiated waves.

To counter this problem, the feed waveguide is constructed using polarization selective surfaces (PSSs); for example surfaces which reflect $x$-polarized waves but are transparent to $y$-polarized waves. In this manner the leaky waveguide can be made to guide $x$-polarization ($TE$ modes) while simultaneously being transparent to $y$-polarization. The feed wave is then chosen to be a $TE$ wave and the reflectarray aperture, in addition to tuning the phase, rotates the polarization from $TE$ ($x$-polarized) to $TM$ ($y$-polarized) allowing it to pass through the feed waveguide with minimal perturbation.

While Fig. 3.2 is only one-dimensional, ideally the entire antenna would be a two-dimensional planar array such that it could produce full-space steerable pencil beams. Generally reflectarray apertures are two-dimensional, therefore extrapolating the one-dimensional reflectarray aperture in Fig. 3.2 to the two-dimensional case is mostly trivial. Doing the same thing for the integrated feed (in this case the leaky waveguide) requires a bit more thought. The original idea was to make the leaky-waveguide a parallel plate guide which could guide waves from any excitation point. In particular, most interesting was the case where the excitation could be at the center and produce a cylindrically propagating parallel-plate mode (Fig. 3.3(a)). The advantage of a center excitation is that it naturally provides equal power distribution to all sections of the array, as well as aids in producing more triangular-like amplitude tapers which help with sidelobe suppression.

The disadvantage of a center excitation however is that the modes are cylindrical. In the case of a $TE$ parallel plate mode then the electric field is actually circulating around the center of the array. This then requires both leaky PSS surfaces and a reflectarray aperture which either respond to all polarizations equivalently, or are physically rotated.
to match the change in polarization of the feed wave. The latter is difficult as the resulting radiation would then not all be of the same polarization, while the former is simply more difficult technologically. A solution which is used in this thesis is to still use a parallel-plate waveguide, but instead excite the waveguide from the side instead of from the center (Fig. 3.3(b)). This, if done properly, would produce a $TE$ mode with a planar wavefront, and thus all electric fields oriented in the same (linear) direction. The difficulty with this mode is simply finding a way to properly generate a planar wavefront.

The frequency of operation chosen for this array is 5.5 GHz due to the prevalence of reconfigurable reflectarray designs at and around this frequency. This integrated feed reflectarray architecture requires the design of two different structures: the polarization selective leaky waveguide and the phase shifting/polarization rotating reflectarray aperture. The following section will discuss the design of the leaky waveguide, and the section

Figure 3.3: (a) Center and (b) side excited parallel plate feeds for reflectarray.
after that will discuss the design of the reflectarray aperture.

3.3 Polarization Selective Leaky Waveguide Design

The leaky feed waveguide has two constraints. The first is that the guide be leaky, and therefore somewhat transparent. Ideally this leaky behaviour should be tunable (in a fixed sense) to allow for different leakage constants for different sized arrays. The second constraint is that the waveguide be polarization selective. The waveguide is required to guide (and leak) one polarization while being completely transparent to the other polarization. The choice of the linearly excited parallel plate waveguide shown in Fig. 3.3(b) becomes more apparent since all of the incident guided fields will be of one polarization, either $TE$ or $TM$, simplifying the design of this PSS.

To come up with a design for the polarization selective leaky surfaces, initial research was conducted into frequency selective surfaces (FSSs) of which Benjamin Munk [48] performed much of the pioneering work. One of the simplest FSSs investigated is the gangbuster array which is an array of tightly coupled dipoles arranged in an offset configuration (Fig. 3.4(a)). These arrays offer strong frequency selective behaviour when the electric field and plane of incidence are oriented parallel to the length of the dipoles (i.e. $\hat{x}$ in Fig. 3.4(a)). However when the plane of incidence is orthogonal to the length of the dipoles (i.e. $\hat{y}$ in Fig. 3.4(a)), the arrays becomes highly polarization selective. Properly designed they can simultaneously become almost fully reflective for $TE$ waves and fully transparent for $TM$ waves. Even simpler than the gangbuster array is an array of dipoles located in a rectangular lattice without the offset that gangbusters possess (Fig. 3.4(b)). This surface offers similar polarization selectivity but with an even simpler design (although lacks some of the frequency selective properties which gangbusters are good for). In both cases these dipoles are printed on a substrate, making them easy to fabricate.

FSSs and PSSs are typically used in free space applications such as filters or lenses, so how can they be used with waveguides? As was discussed in Section 2.3.2, guided modes in waveguides can largely be seen as free space waves bouncing between the top and bottom (and/or side) surfaces of the guide. In the case of $TE$ modes the bouncing waves are $TE$ waves, and with $TM$ modes they are $TM$ waves. Thus the behaviour of an FSS or PSS as the boundary in a waveguide environment should be very similar to that of the free space environment. The dipole array mentioned previously is almost completely reflective for $TE$ waves incident perpendicular to the dipoles. It is therefore logical to conclude that replacing the “bounce” walls with this PSS with dipoles oriented...
Figure 3.4: (a) Gangbuster and (b) simple dipole array polarization selective surfaces.

Figure 3.5: Comparison between parallel plate waveguide $TE_{1}$ mode and rectangular waveguide $TE_{10}$ mode. Regardless of the rectangular waveguide width, the modes are identical.

Transverse to the direction of propagation will allow the $TE$ mode to be guided (since its electric field is also transverse).

To simplify the design process, all of the work in the following sections considers only one column of the parallel plate waveguide where the column’s length coincides with the direction of feed wave propagation. It is easy to show that if such a slice of a parallel plate waveguide is taken, the $TE_{1}$ parallel plate mode and $TE_{10}$ rectangular waveguide mode are identical and only depend on the height between top and bottom surfaces. Fig. 3.5 demonstrates this concept. This column or slice is taken to have a width equal to the unit cell spacing of the reflectarray elements, which in this thesis will be slightly more than half a wavelength. The choice of the simple dipole surface is also slightly more obvious now, as the dipoles of the PSS can conveniently fit within the unit cell without crossing the boundaries.

Of course the waveguide is only meant to be leaky on the bottom side, which ne-
cessitates designing the upper and bottom PSSs differently. The top is designed to be almost perfectly reflective to the TE mode (acting as a perfect conductor) while the bottom is designed to be slightly leaky by giving it a finite transmission coefficient (but still presenting itself as close to a perfect conductor). This will enable the waveguide to leak a small percentage of its power over each reflectarray element. It will be shown in the next subsection that both of these constraints can be easily satisfied using the basic dipole array simply by manipulating the dipole spacing and length.

3.3.1 Design Process

The design of the polarization selective leaky waveguide is aided heavily by the use of transverse equivalent network (TEN) theory described in Section 2.3.2. As mentioned frequently, this feed structure is actually a leaky waveguide, and can therefore be analyzed almost entirely in terms of both its leakage constant $\alpha$ and phase constant $\beta$. By using a TEN model, only unit cells of the top and bottom surfaces need be simulated, with the surface simulations results fed into the TEN model along with waveguide geometry to find the corresponding leaky propagation constant. Typically in leaky-wave antennas a target $\alpha$ is chosen according to the length of the antenna/waveguide while a $\beta$ is chosen to determine the beam pointing angle. In the case of this array antenna, the leaky waveguide is only used to feed the reflectarray, while the reflectarray is used to control the beam pointing angle. Therefore $\beta$ need only be chosen to be close to the $\beta$ of a complete metal waveguide for matching purposes and to offer an incidence angle on the reflectarray close to broadside.

To start, the dipole array surface is modeled in an infinite array environment in Ansys HFSS as was discussed in Section 2.3.1. First of all, the design frequency is chosen as 5.5 GHz since this is the frequency of operation of the selected reflectarray unit cell [10] (discussed in the next section). The unit cell width is chosen to be 30 mm ($0.55 \lambda_0$) corresponding to the width of the reflectarray unit cell. The unit cell length is variable and is always an integer multiple of the dipole spacing (periodicity) chosen to remain as close to 30 mm as possible. Fig. 3.6 shows an example HFSS model with a dipole spacing of 4 mm, dipole width of 1.36 mm and a dipole length of 25 mm. The dipoles are printed on substrate, in this case 3.175 mm Rogers RT/duroid® 5880. Master/slave boundaries are placed on all sides of the unit cell and Floquet ports are placed at both the top and bottom. Only the fundamental Floquet modes $TE_{00}$ and $TM_{00}$ are simulated as all others are cutoff. The incidence angle is initially chosen to be 30 degrees as this roughly corresponds to the expected bounce angle inside the waveguide. The bounce angle won’t
Figure 3.6: HFSS model of dipole array unit cell.

Figure 3.7: (a) $TE$ and (b) $TM$ reflection coefficient magnitude for dipole array with varying dipole length and spacing.

actually be known until after using the TEN, therefore iteration may be required if the bounce angle deviates significantly from 30 degrees.

Simulating this unit cell gives a $4 \times 4$ $S$-matrix (2 ports $\times$ 2 polarizations each). This $S$ matrix contains most importantly the reflection and transmission coefficients for both $TE$ and $TM$ polarizations; the cross terms are not as important since they are generally quite small for this surface. Fig. 3.7 shows the simulated reflection coefficient magnitude for both modes versus both dipole spacing and dipole length. For $TM$ polarization the surface is almost entirely transmissive (low reflection coefficient magnitude) for most variations. Changes to the dipole width were not seen to significantly change reflection or transmission behaviour for either mode. It should also be mentioned that the maximum dipole length is the width of the reflectarray unit cell and the minimum spacing corresponds to the width of the dipoles. In Fig. 3.7 the variation corresponding to a dipole length of 25 mm and dipole spacing of 13 mm has a $TE$ reflection coefficient of practically 0 dB and a $TM$ reflection coefficient of $-15.7$ dB. Therefore it is almost perfectly reflective in $TE$ while relatively transparent in $TM$ making it an excellent candidate for
the top surface.

While the top surface is almost perfectly reflective in $TE$, the bottom surface needs to be slightly transparent in $TE$ to allow leakage. To pick the appropriate dipole dimensions for the bottom surface the TEN model (Fig. 3.8) is used to find $\alpha$ for varying bottom surface dipole dimensions. Each dipole surface in the TEN is modeled as a two-port network with a $2 \times 2$ $S$-matrix corresponding to the $TE$ parameters pulled from the original dual-polarization $4 \times 4$ $S$-matrix simulated above. The top surface is modeled using the two-port matrix for the case of 25 mm by 13 mm dipoles. The bottom surface two-port network is swept through different two-ports corresponding to the different dipole variations shown in Fig. 3.7. Each variation requires a separate HFSS simulation, the same simulations used to produce Fig. 3.7. The height $h$ of the waveguide is specified as the distance between substrates and for this design was chosen to be 26.825 mm. The dipoles are placed on the outside of the guide and therefore the distance between dipoles is 33.175 mm. This distance was chosen as a starting point to get a bounce angle close to 30 degrees. The top and bottom of the TEN are terminated in the free-space wave impedance mentioned in Section 2.3.2. The reflectarray surface is excluded to simplify the model. This assumption is valid as long as the reflectarray unit cell is rotating most of the power into the $TM$ mode (i.e. it acts as a matched load for the $TE$ mode).

Fig. 3.9 shows a plot of the resultant $\alpha$ and $\beta$ for the leaky waveguide as a function of the bottom surface dipole length and spacing. Also plotted is a 57.6 rad/m phase constant
contour which corresponds to a bounce angle of 30 degrees. To keep the bounce angle around this value, operating points should be chosen close to this contour. The minimum phase constant achieved is 45.8 rad/m corresponding to a bounce angle of 23.4 degrees while the maximum is 88 rad/m corresponding to a bounce angle of 49.8 degrees. It can then be seen that an extremely wide range of $\alpha$ can be chosen for bounce angles varying from 23.4 to 49.8 degrees. While it was not performed in this thesis, the height of the waveguide can also be varied to adjust these curves. For now the height is at a reasonable value as it is close to the height of WR137 waveguide (34.8488 mm) which could be used to feed the leaky waveguide.

The choice of $\alpha$ is determined strictly by the overall length of the waveguide/array. As mentioned in Chapter 2, $\alpha$ is normally chosen to enable greater than 90% power leakage. To test the results of this modeling technique, a linear leaky waveguide was created in HFSS and simulated (Fig. 3.10). The waveguide was made 8 free-space wavelengths long which is 436.4 mm. For greater than 90% power leakage an $\alpha$ of at least 0.4 Np/m should be chosen. The point with a dipole spacing of 4 mm and a dipole length of 20 mm was chosen for this test. This gives an estimated $\alpha$ by the TEN model of roughly 3.3 Np/m and a $\beta$ of roughly 55.5 rad/m (28.8 degree bounce angle). The waveguide was excited using waveports the same dimensions of the waveguide but with short metal deembedding sections. Periodic boundaries with zero degree phase difference are used on the sides of the array to simulate a periodic slice of an infinitely wide array.

The resulting normalized far-field pattern for this leaky waveguide can be seen in Fig. 3.11. The beam experiences a maximum at 27 degrees corresponding to a $\beta$ of 52.3 rad/m. This method tends to slightly underestimate the value of $\beta$ since it assumes an isotropic element factor. The $\beta$ is therefore reasonably close to that predicted by the TEN model. The return loss for the waveguide is 35.6 dB and the insertion loss
is 21.2 dB. Using the $TE$ mode $S$ parameters the $\alpha$ can be computed and is found to be 5.59 Np/m. This is larger than expected using the TEN model, but indicates that the overall design process is sound and that further optimization will bring this to the expected level.

### 3.4 Reconfigurable Reflectarray Aperture Design

This section will now discuss the design of the electronically reconfigurable reflectarray aperture, particularly the unit cell. This reflectarray unit cell is unique in that it needs to adjust both the reflection phase and polarization. Phase shifting reflectarray unit cells have been commonly developed in the past as was discussed in Section 2.2.2. Since full-space pencil beam-steering is desired, emphasis has been placed on the use of continuously tunable (analog) phase shifting mechanisms. One simple yet effective design which provides continuous phase tuning is the varactor diode loaded microstrip patch.
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Figure 3.12: Reconfigurable reflectarray cell topology presented in [10].

[10]. This patch is capable of adjusting the reflection phase by up to 360 degrees by varying a reverse bias voltage applied across two halves of the patch. This section will show how modifications to this design can allow it to simultaneously adjust phase and polarization simply by introducing additional varactor diodes across the orthogonal patch direction.

3.4.1 Unit Cell Design and Operation

Fig. 3.12 shows the electronically reconfigurable reflectarray element presented in [10]. This element consists of a microstrip patch split in half with varactor diodes positioned across the gap and at the sides of the patch. The diodes are oriented parallel to the direction of current flow on the patch, and can be seen as tunable capacitive loading to the patch. By tuning the varactors (through application of a reverse bias voltage) the effective resonant length and hence resonant frequency of the patch changes. The utility of this behaviour is that as the patch is tuned through resonance (either by changing frequency or capacitance), the reflection phase changes. This change of reflection phase is demonstrated through iconic S-curves as shown in Fig. 3.13. In this figure different capacitance states locate the resonant frequency at different positions and therefore shift the S-curves up and down in frequency. At a fixed frequency, it can be seen from Fig. 3.13 that with an appropriate choice of capacitance range (as well as patch and substrate properties), the reflection phase can be made to vary between ±180 degrees. Note that the phase range is greater than 180 degrees due to the additional pole created through the imaging of the patch across the ground plane (power can be thought of as traversing the patch twice).

To add dual-polarization and polarization control capabilities, the above patch is
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Figure 3.13: S-curves (reflection phase curves) produced by reconfigurable reflectarray element [10].

modified by incorporating a gap and varactor diodes along the other cardinal direction of the patch. Fig. 3.14 shows the modified topology of the phase shifting/polarization rotating reflectarray unit cell. The element is for the most part a tunable square microstrip patch located on a substrate with a ground plane. The unit cell is square with a width (periodicity) of 30 mm in both directions (0.55 λ₀ at 5.5 GHz). The patch itself is rotated 45 degrees within the unit cell and is 17 mm × 17 mm (0.31 λ₀). Four varactor diodes are placed at the four edges of the patch across 1 mm gaps cut along the principal patch directions. Each set of varactors, labeled A and B in Fig. 3.14, control the reflection phase of a wave polarized parallel to the patch edge which the varactors are placed along. The varactors were chosen to be Aeroflex/Metelics GaAs MGV100-20 hyperabrupt varactor diodes which possess a capacitance range of roughly 0.15 to 2.1 pF when reverse biased from 0-20 V. The substrate was chosen to be Rogers RT/duroid® 5880 and the substrate height will be chosen later.

To produce polarization rotation, isolation in the reflection response between the two pairs of varactors (polarizations) is exploited. Fig. 3.15 illustrates the process. The patch is first excited with an electric field $\vec{E}_{inc}$ which is diagonal to the patch (parallel to the edges of the unit cell). This incident field can then be decomposed into equal components along both of the principal patch directions: $\vec{E}_{inc,A}$ and $\vec{E}_{inc,B}$. Each of these components is reflected individually from the patch receiving their own unique reflection phase dependent on the tuning of the diode pair aligned along their respective directions. To rotate the field, the two diode pairs apply a phase changes which are separated by 180 degrees. For example in Fig. 3.15, $\vec{E}_{ref,A}$ receives a zero degree phase shift while $\vec{E}_{ref,B}$ receives a 180 degree phase shift, effectively flipping its direction. This
causes the two reflected components to recombine in a polarization orthogonal to the incident polarization ($\vec{E}_{\text{ref}}$) and hence be rotated by 90 degrees. Note that any absolute reflected phase between 0 and 360 degrees can still be produced, all that is important is that the phase states of the two components be separated by 180 degrees.

### 3.4.2 Principal Direction Unit Cell Simulation Results

To verify the unit cell operation the response along a single patch direction is first simulated in HFSS. The corresponding HFSS model can be seen in Fig. 3.16. As with the dipole array unit cell, master/slave boundaries are used on all four sides with a Floquet port at the very top and a ground plane at the bottom. The substrate is initially chosen to be 3.175 mm Rogers RT/duroid® 5880. For initial simulations the incidence angle is also chosen to be broadside to simplify analysis. With this setup only the fundamental
Figure 3.16: HFSS model of reflectarray unit cell.

$3 \, \Omega, \ 0.4 \, \text{nH, 0.15–2.1 \, \text{pF}}$

Figure 3.17: Model of Aeroflex/Metelics MGV100-20 varactor diode.

$TE$ and $TM$ Floquet modes need to be excited, all others are evanescent.

The varactors diodes can be modeled in two ways. In both cases the varactors are modeled as surfaces in HFSS positioned at roughly half the height of the actual varactor diodes and with metal leads connecting the surfaces to the patch. In the first modeling method, the surfaces are series RLC boundaries (a type of HFSS boundary) where each boundary represents a component (resistor, inductor, capacitor) of the varactor equivalent circuit model. For this design the varactors were modeled using a series resistance of 3 Ω, a series inductance of 0.4 nH and a variable capacitance corresponding to the reverse bias junction capacitance (Fig. 3.17). Previous work [10], [26] has shown good agreement with measurements using these or similar values. For this method, S-curves are obtained by sweeping the varactor reverse bias capacitance in HFSS (by changing the value of the corresponding RLC boundary). This requires a completely new full wave solution for each capacitance state in HFSS and is therefore quite time consuming.

To alleviate the time needed to sweep the capacitance, the second modeling method was developed which involves replacing the RLC boundary with a lumped port. For this method, only one full-wave simulation is run, however the results now include a $6 \times 6$ S-matrix including the two Floquet modes and the four lumped ports. The equivalent circuit models are then introduced across the lumped ports during post-processing in MATLAB. While this saves a tremendous amount of time, there is a small discrepancy between the first and second methods, namely the S-curves shift down slightly in capac-
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Figure 3.18: Principal plane reflection (a) phase and (b) magnitude for the reflectarray unit cell.

ite. Therefore the second method is used mostly for design investigations, while the first is used to produce the final results used for designing the array.

Fig. 3.18 shows a plot of the reflection phase and magnitude versus diode capacitance for an incident wave polarized along one principal direction of the patch at 5.5 GHz. The other direction diodes are biased at 1 pF. As can be seen, the total phase range is not quite 360 degrees but slightly less. Additionally, the reflection losses peak at a point roughly in the middle of the phase curve corresponding to resonance. This dip in reflection is due to power being absorbed by the varactors at resonance. The series resistance of the varactor diodes will turn out to be the most significant source of loss in this design.

Since capacitance range is limited, there is in fact a trade-off between the overall reflection losses and the phase range. This trade-off can be explored by changing certain patch parameters such as the substrate height. Fig. 3.19 shows a plot of the principal direction reflection properties as the substrate height is varied between 1 and 3.25 mm. For large substrate heights the phase response possesses S-curves with generally lower phase ranges and smaller losses at resonance (dips in the magnitude curve). As substrate height decreases, the phase range increases and the losses at resonance also increase. This region is the so-called over-coupled region and corresponds to small resonator $Q$ values [49]. Once the substrate height reaches a critical value, in this case close to 1.75 mm, the patch becomes over-coupled and presents an anomalous phase response (not an S-curve) with maximized losses. As the height continues to decrease the patch becomes under-coupled with an anomalous phase response, low phase range and large $Q$ values.

Other parameters such as patch length/width and substrate permittivity give similar responses, although not as pronounced. These trade-offs must be considered during the design of the patch. A limited phase range leads to phase errors throughout the
array which in turns leads to pattern distortion and heightened side lobes while loss decreases gain and also leads to pattern distortion (since loss is a function of tuning). For this design, a substrate height of 3.175 mm using the low-loss ($\tan\delta = 0.0009$) Rogers RT/duroid® 5880 with a dielectric constant of 2.2 combined with 17 mm patches produces an optimal response with relatively low loss and around 300 degrees of phase range. 300 degrees was found to be sufficient to avoid substantial side lobe degradation in the final pattern.
3.4.3 Polarization Rotation Unit Cell Simulation Results

Up to this point only the response of the unit cell to waves polarized along the principal directions of the patch was considered. These simulations gave rise to one-dimensional $S$-curves. In this subsection the response of the unit cell to waves polarized diagonally to the patch will be investigated, resulting in responses which are two-dimensional (depending on the capacitance of both varactor pairs). In particular, the ability for the patch to rotate the polarization while simultaneously providing phase tuning will be investigated.

The same unit cell in HFSS is now excited with a $TE$ field incident at 30 degrees with polarization along the diagonal of the patch. Fig. 3.20 shows plots of the $S_{21}$ magnitude and phase when both varactor pairs are tuned; $S_{21}$ in this case meaning the transmission coefficient from $TE$ to $TM$ polarization. Since the patch is symmetric, the plots are (mostly) symmetric other than a 180 degree phase difference between the upper and lower triangular regions of the plot (i.e. flipping the bias on the sets of varactors will cause a negative rotation of the fields compared to before). There is a deep null in the magnitude along the diagonal which is expected since no rotation will occur if the varactor pairs are biased identically.

The different biasing configurations can be determined from these plots by generating contours for each desired phase on the phase plots and picking a point on each contour which coincides with maximal rotation (largest $S_{21}$) for that phase. To do this the phase plots are divided into their upper and lower triangular regions (matrices). These matrices are then duplicated across the diagonal to create actual symmetric matrices for each upper and lower region. The diagonal elements for each matrix are interpolated from their closest neighbour using a linear, nearest neighbour interpolation scheme. Remember that the diagonal phase elements are originally just noise since the magnitude is so low. With these new matrices, smooth, continuous contours can easily be plotted through the
data for phase steps between the minimum and maximum phase at desired intervals (in this thesis one degree intervals were chosen). Along each contour the point which gives the largest $S_{21}$ is chosen as the configuration for that phase. The resulting magnitude versus phase curves for the upper and lower triangular regions can be seen in Fig. 3.21. This plot gives a rough way of seeing what the phase range and loss is for the rotation mode patch.

The interesting thing about this design is that despite the linear phase response only having 300 degrees of phase range, the extra dimension afforded by the second pair of diodes allows the missing phase to be regained. In fact larger phase ranges can 360 degrees can be produced if larger losses are accepted. In this case Fig. 3.21 demonstrates a fully achievable 360 degree phase range with an average loss of 1.6 dB and a maximal loss of around 3.4 dB. Of course the full phase range does not need to be used (in exchange for introducing phase errors in the aperture), in which case the loss could be reduced by using only configurations with say higher than 3 dB loss.

### 3.5 Linear Array Simulations

With both the leaky waveguide and the reflectarray unit cell designed and analyzed, they can now be combined and simulated as a whole. As was done with the leaky waveguide, only a single longitudinal slice of the array was simulated. Fig. 3.22 shows
the linear array model in HFSS. It is fed with waveports located at the two ends of the 
leaky waveguide. Complete metal waveguide with substrate present is used for half a 
free-space wavelength to transition from waveport (rectangular waveguide) to the leaky 
waveguide and ensure evanescent modes die off before the waveports. The reflectarray 
aperture is placed exactly half a free-space wavelength (27.27 mm) below the bottom of 
the leaky waveguide. This height was chosen as a starting point with the possibility of 
future optimization. The rationale for half a wavelength was that it was large enough to 
allow ample decoupling between the waveguide and the reflectarray aperture, but was also 
small enough to not significantly increase the size of the array. The leaky waveguide is 
eight free-space wavelengths long (436.36 mm) while the reflectarray aperture is slightly 
larger (16 cells totaling 480 mm) to ensure capture and rotation of all leaked power. 
Finally master/slave boundaries were placed on the sides with a zero degree phase shift 
to simulate transverse periodicity and a radiation boundary was placed on the top and 
ends to capture radiated power.

It should be stressed that this array is infinite in the transverse direction but finite in 
the longitudinal direction. Ideally the array would be made finite in both the longitudi-
nal and transverse directions. Unfortunately, the use of two different polarizations and 
hence both $TE$ and $TM$ modes makes it extremely difficult to provide proper boundary 
conditions for such a case. The issue is that the guided $TE$ mode requires perfect elec-
tric conductor (PEC) boundaries on the sides, whereas the $TM$ mode requires perfect 
magnetic conductor (PMC) boundaries. Therefore simply using metal walls (PEC) on 
the sides would allow proper feed wave propagation, but the radiating mode would be 
significantly perturbed and/or cutoff. This issue will be further discussed at the end of 
this chapter. In the meantime periodic boundaries with zero degree phase shifts are used 
which act as satisfactory boundary conditions for both $TE$ and $TM$ modes.
3.5.1 Phasing/Configuration Algorithm

In order to provide polarization rotation while simultaneously producing a steerable pencil beam, a method of configuring the biasing of each unit cell to provide rotation and the required phase is developed. This is done through a configuration algorithm. This algorithm takes as input the two-dimensional reflection rotation response (Fig. 3.20), the desired beam-steering direction $\theta_b$, the leaky waveguide phase constant $\beta_g$ and the array size (periodicity $d$ and number of cells $N$) and outputs the bias configuration for each cell.

To start, the required phase gradient $\beta_b$ to produce a pencil beam in the direction $\theta_b$ is computed using Eqn. 2.6. This phase gradient needs to be a sum of both the phase gradient produced by the leaky phase constant $\beta_g$ and the phase gradient introduced by the cell tuning $\beta_c$. In other words, the required phasing is the difference between the required phase gradient and leaky phase gradient

$$\beta_c = \beta_b - \beta_g.$$  \hspace{1cm} (3.1)

This phase gradient results in a phase difference between cells of $\Delta \phi = \beta_c d$. The required phase for the $n^{th}$ cell in the array is then given by $\phi_n = n \Delta \phi + \phi_0$. $\phi_0$ is a starting phase and can be any value between 0 and 360 degrees. It is used to produce several different “sets” of phases with identical phase gradients but differing absolute phases for each cell. For instance, choosing 1 degree steps for $\phi_0$ will produce 360 different sets of phase configurations. Each phase configuration will produce the same phase gradient and hence the same pencil beam angle, but will differ in the amplitude weighting of each cell since the cell amplitude is dependent upon the absolute phase of the cell (Fig. 3.21).

For each cell in each set, the required absolute phase is interpolated on Fig. 3.20(a). This produces a curve or contour over the two-dimensional plot with multiple different capacitance coordinates (diode pair A capacitance, diode pair B capacitance) yielding the required phase. To pick between these coordinates the curve is then interpolated on Fig. 3.20(b). Each coordinate then has the same phase but differing magnitudes. The coordinate with the largest magnitude is chosen. At this point each set has biasing coordinates assigned to each cell which give different amplitudes between sets. To choose which set to use, the average magnitude for each set is found (averaging over all the cells in the set) and the set with the largest average magnitude is chosen. This choice gives the largest overall radiated power and also ensures maximal rotation. Another choice could be to find the variance between cells in each set and choose the set with the lowest variance as this would give the flattest amplitude distribution.
3.5.2 Un-phased Configuration

Fig. 3.23 shows a plot of the simulated normalized total directivity in HFSS compared to a corresponding normalized pattern factor (discussed shortly). The HFSS simulation is run using a bias configuration which provides maximal rotation but applies no phase shift between elements allowing the array to radiate at its leaky angle. The beam maximum is at 27.3 degrees corresponding to a leaky waveguide $\beta$ of 52.8 rad/m. The reflection loss for the integrated feed waveguide is 19.5 dB showing a good match between leaky and full metal waveguide sections and the insertion loss is 32 dB showing that almost all power is leaked. The insertion loss corresponds to an $\alpha$ of 4.92 Np/m which is slightly less than the waveguide $\alpha$ of 5.59 Np/m. As before, it should be noted that this $\alpha$ is due to contributions both from radiation and loss. The peak directivity is 15.3 dBi, the 3 dB beamwidth is 8.3 degrees and the sidelobe level is $-10.3$ dB.

The pattern factor in Fig. 3.23 is a combination of an array factor and an element factor. The array factor is for a 16 element linear array excited using the $\alpha$ and $\beta$ extracted from the HFSS model as well as the element magnitudes and phases extracted from the configuration algorithm (which are almost perfectly uniform). The element factor is assumed to be a rectangular aperture with dimensions equal to the unit cell dimensions (30 mm by 30 mm) [6]. The peak total directivity for the pattern factor is found to be 17.1 dBi. The side lobe level is also different, evident from Fig. 3.23.

There are three main suspected causes of these discrepancies. The first is that the actual array in HFSS is not exactly 16 perfect $30 \times 30$ mm apertures. In fact the leaky waveguide, which is essentially the total aperture size, is only 14.5 elements long. This
truncated aperture will lead to a broadening of the beam (decreased directivity) and shifting of the side lobe positions. The actual array also likely possesses an element factor slightly different than that of a perfect aperture. The second cause would be reflections inside the waveguide, either off the ends, or from the leaky waveguide itself. These spurious reflections would lead to enhanced coupling between elements, leading to phase/magnitude errors and hence deteriorated side lobes and reduced directivity. The third suspected cause is that some power is getting caught below the rectangular waveguide sections and leaking out the sides of the array, as opposed to being radiated from the top. This could happen if there is incomplete rotation, causing some power to be guided in the lower section below the waveguide.

3.5.3 Leaky Waveguide Tapering

Edge effects are thought to be the main contributor for pattern degradation, particularly reflections/interactions at the start of the array where most of the power is leaked. To improve the pattern, the approach considered was to apply tapering to the leaky waveguide by changing $\alpha$ as a function of position (thereby changing the amplitude distribution). This would mostly lead to improvements in side lobe level, however, if significant power is leaking out the sides of the array, it could also increase the directivity (could think of it as an improved spillover efficiency).

Changes to $\alpha$ must be continuous in order to avoid significant perturbations to propagation. Unfortunately, continuous changes to $\alpha$ cannot be performed without also changing $\beta$ (see Fig. 3.9). Providing a continuous taper over the entire aperture would then lead to a continuously changing $\beta$ which increases the difficulty in properly phasing the elements to produce a pencil beam. Therefore it was decided to only taper the leakage from the waveguide at the start and end of the array to minimize end effects while keeping $\beta$ as constant as possible.

Referring to Fig. 3.9, it can be seen that as the leaky waveguide bottom dipole length is increased from 20 mm to 30 mm (width of the waveguide section) while keeping the spacing fixed at 4 mm the $\alpha$ decreases to zero while the $\beta$ increases and approaches the full metal waveguide $\beta$ of 65.6 rad/m. Hence by tapering the bottom dipole length from the design length of 20 mm to the waveguide width of 30 mm, the leakage can be tapered off while also moving the phase constant closer to that of the feeding metal waveguide, thereby improving the match. Since this tapering does affect the phase constant and therefore the beam pattern, it should preferably be done over a shorter distance. Experimentation in HFSS showed that a taper distance of 112 mm (28 dipoles) produced
reasonable results. This taper is then applied to the start and end of the leaky waveguide bottom surface. Fig. 3.24 shows an HFSS model of the tapered array. Fig. 3.25 shows plots of the dipole length and interpolated $\alpha$ and $\beta$ as a function of position along the array.

The simulated total gain for this tapered array can be seen in Fig. 3.26 compared to the total gain for the array without tapering. Analysis of the insertion loss gives a new average $\alpha$ of 3.0 Np/m while the beam angle remains the same and gives a $\beta$ of 53 rad/m. Overall the directivity increased by roughly 0.3 dB to 15.6 dBi, while the gain increased by 0.5 dB to 14.1 dBi. A simulation with lossless varactors was also run yielding a gain of 15.5 dB. This suggests around 1.4 dB of loss due to the varactors alone. The beamwidth was broadened from 8.3 degrees to 10 degrees and the sidelobe level improved from $-10.3$ dB to $-14.6$ dB. The return loss and insertion loss also improved to 24.5 dB and 38.8 dB respectively.

The increase in beamwidth and drop in sidelobe level are easily explained by application of the taper. The increase in directivity with simultaneous increase in beamwidth confirms the fact that power was likely being coupled out the sides of the array. That power was contributing to a reduced “spillover” efficiency. Now that the taper is added, less power is concentrated near the start and end of the array, and it is therefore likely that less power is being coupled out. This extra power being radiated would then likely increase the directivity despite the decrease in aperture size due to the taper. Small changes to $\beta$ over the taper section also probably contribute slightly to the increase in beamwidth.

Finally Fig. 3.27 shows the co-polarization $(E_\theta)$ and cross-polarization $(E_\phi)$ levels. The cross-polarization level is roughly $-26$ dB. Most of this is due to the finite leakage from the top polarization selective surface of the leaky waveguide and which can be seen
Figure 3.25: (a) Dipole length, (b) $\alpha$ and (c) $\beta$ as a function of position along the linear array.

producing a small beam at the leakage angle.

3.5.4 Electronic Beam-Steering

The configuration algorithm discussed earlier was then used to produce biasing configurations to steer the beam. Fig. 3.28 shows plots of the gain for beams between $-45$ and $+45$ degrees at 15 degree intervals. Table 3.1 shows the key beam properties for the various steering angles. Overall the beam steers quite well, with the main beam being extremely close to the design angle for all steering angles. For extreme angles the beam is pulled closer to broadside due to the element factor pattern. This could be compensated for in future beam-steering by simply trial-and-error.

The directivity fluctuates from 13.6 dBi at the extreme to 15.6 dBi at broadside while the gain fluctuates from 11.8 dBi to 14.3 dBi. The changes are again mostly due to the array element factor although the unusually low gain for the $-15$ degree beam is likely due to large phase and magnitude errors. Comparing the directivities and gains the total losses are roughly between 1.2 and 1.7 dB which are almost entirely due to the varactor diodes. The maximum possible directivity for this aperture size is roughly 17.4 dBi. The
large difference (roughly 1.8 dB) between the simulated directivity and this maximum directivity suggests low phase and taper efficiencies. As mentioned before, it could also be caused by power being coupled out the sides of the array instead of being radiated, contributing to an effective spillover efficiency.

The sidelobe level is unfortunately not as well behaved and varies quite widely between $-7.2$ dB at the extreme and $-15.3$ dB for beams close to the leakage angle. In particular the beams at $\pm15$ degrees show large sidelobes which appear to be almost mirrored across broadside. At first glance it may appear that these sidelobes are due to backwards traveling waves. However, reflected power inside the leaky waveguide should not produce beams mirrored across broadside (which is typical in leaky wave antennas) since the total phasing is a sum of both the leaky phase constant and the applied phase gradient. If the leaky phase constant is reversed (such as for a reflected backward wave), the phase
gradient would still shift the beam by the same amount and direction it would if the feed wave were moving forward. For example, a beam steered to 15 degrees applies a −15 degree phase shift to a leaky wave radiating at approximately 30 degrees. If that leaky wave were reflected, it would radiate at −30 degrees but the phasing would subtract 15 degrees to produce a reflection beam at −45 degrees.

It is possible that for those particular radiating angles, the radiation phase constant leads to excitation or coupling to unknown and unwanted modes inside the structure. For the other angles, multiple reflections inside the structure could lead to similar effects which cause pattern degradation. The problem with this design is that the use of multiple layers create many ways in which power can be reflected while traversing through the antenna. For instance, power leaked to the reflectarray surface, even if it is fully rotated, may reflect off the leaky waveguide while moving upwards then be re-phased and rotated by the reflectarray surface. The result is essentially guided waves in the lower waveguide section formed by the reflectarray surface and the bottom dipole surface. Another com-

Table 3.1: Simulated steered beam properties for each steering angle.

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<tr>
<td>−45</td>
<td>−42.1</td>
<td>13.6</td>
<td>12.1</td>
<td>−7.2</td>
<td>9.5</td>
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<tr>
<td>−30</td>
<td>−30.5</td>
<td>15.2</td>
<td>13.6</td>
<td>−14.8</td>
<td>11.1</td>
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<tr>
<td>−15</td>
<td>−14.7</td>
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<td>12.9</td>
<td>−8</td>
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<tr>
<td>0</td>
<td>0.2</td>
<td>15.6</td>
<td>14.3</td>
<td>−14.8</td>
<td>9.7</td>
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<td>15</td>
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<td>14.8</td>
<td>13.6</td>
<td>−9.3</td>
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<td>30</td>
<td>28.8</td>
<td>14.4</td>
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<td>45</td>
<td>42.4</td>
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The complicating factor is incomplete rotation by the reflectarray surface. Incompletely rotated waves will again be guided in the bottom waveguide section and be phased multiple times.

The return loss varies between 33 dB and 10 dB across all design angles. The insertion loss varies between 22 dB and 13.5 dB. The large changes in these values are likely due to the changing impedance presented by the bottom reflectarray aperture and incomplete rotation. Despite these changes, the return loss and insertion loss are always above 10 dB which for most applications is acceptable. Lastly, Fig. 3.29 shows a plot of the co- and cross-polarization levels for the broadside and extreme angle beams. At broadside the cross-polarization level is approximately $-26$ dB which is similar to that of the un-phased design analyzed earlier. As the beam is scanned to extreme angles the cross-polarization level degrades to around $-20$ dB.

### 3.6 Design Conclusions

In this chapter, the design of a new type of antenna named an integrated feed reflectarray was presented. This antenna is a hybrid between a folded reflectarray and a leaky-wave antenna. It removes the need for an external feed antenna by utilizing a planar integrated leaky waveguide to distribute power over the reflectarray aperture. The waveguide is a parallel plate waveguide made from polarization selective leaky surfaces and is placed a small distance above a reconfigurable reflectarray aperture. The reflectarray aperture is then capable of electronically reconfiguring both the phase and polarization of the leaked waves, and produces a full-space steerable pencil beam which can pass right through the
leaky waveguide.

One-dimensional linear array simulations showed that the array is capable of reliably steering a beam in at least one dimension from $-45$ to $+45$ degrees. Despite this, the array possesses several drawbacks which make it difficult to analyze, degrade performance and also make it difficult to fabricate. The need to control polarization is the most significant drawback. This requirement necessitates more complicated reflectarray unit cells and polarization selective surfaces, makes it difficult to adequately terminate the array on the sides and limits the possible polarizations which can be radiated. The requirement for the extra surfaces is also a significant drawback. It leads to both increased overall profile, and the tendency for power to be reflected and guided in unintended ways. This behaviour leads to poor side lobe performance and reduced overall directivities.

The drawbacks of this integrated feed reflectarray inspire the design of a similar antenna which does not require polarization control and is even smaller in profile. The fundamental problem with the integrated feed reflectarray is that power traverses the structure in multiple directions. First it travels through the leaky waveguide, then down towards the reflectarray, then back up through the leaky waveguide. It would be simpler if instead the power traversed in a single direction the entire time. The next chapter will discuss the proposed solution which is the integrated feed transmitarray.
Chapter 4

Integrated Feed Transmitarray

The integrated feed reflectarray design presented in Chapter 3 possesses many drawbacks, particularly the need for polarization control and the use of multiple specialized polarization-sensitive surfaces. These drawbacks stem from the fact that there is a bi-directional flow of power: power travels from the feed, through the waveguide, down to the reconfigurable aperture and then back up through the waveguide but this time in a different polarization. A much more natural approach to this problem would be to instead have a uni-directional flow of power, from the feed to the aperture to free space, more akin to how a phased array with a beamforming network, or a transmitarray with a spatial feed works.

Similar to the development of the integrated feed reflectarray, existing transmitarray architectures can be modified to move their feed location to the surface of the aperture, providing a similar integrated feed mechanism. This chapter of this thesis will present the design of such an antenna: a reconfigurable transmitarray aperture with an integrated feed closely resembling a leaky-wave antenna. The first section will present an introduction to the design and its similarities to a leaky-wave antenna. The second section will then discuss the transmitarray unit cell which the transmitarray aperture will be based off of. The third section will then discuss the design of the integrated feed, modifications to the transmitarray unit cell to allow it to be easily fed with the new feed as well as the design and simulation of a one-dimensional linear array section of the overall antenna. The fourth section will then move on to the design of a 6 × 6 element array and will present full-space beam-steering simulation results for this array. Finally the fifth and sixth sections will present the results of fabrication and experimental measurements of the 6 × 6 array with integrated feed and then follow up with conclusions.
4.1 Design Introduction

The basic idea behind this antenna is the transmitarray analog of the integrated feed reflectarray presented in Chapter 3. In its most general form, this antenna consists of a discrete array of transmitarray elements excited by a traveling wave propagating within an integrated feed which is now attached below the transmitarray aperture. A cross-section of this design can be seen in Fig. 4.1 and should be compared against the previous reflectarray design in Fig. 3.2. In this particular case the design and operation is even more similar to that of a leaky-wave antenna (discussed in Chapter 2). The traveling wave propagates through the integrated feed, gradually leaking, or in this case coupling power to the transmitarray aperture (TA in Fig. 4.1). Each transmitarray element then phase shifts the coupled power and radiates it from the top. Just like with the reflectarray, each transmitarray unit cell possesses around 360 degrees of continuous phase range to allow continuous beam-steering.

Fig. 4.1 is only one-dimensional however. To produce full-space steerable pencil beams a two-dimensional array is required. Just like with the integrated feed reflectarray, the integrated feed requires a bit more thought when moving from one to two dimensions. The original idea to make the integrated feed two-dimensional was to use a parallel plate waveguide with a center feed as shown in Fig. 4.2(a). Again, a center feed would offer a simplistic feed mechanism with optimal power distribution across the array. The propagating mode could be a $TE$ wave or any other cylindrically propagating mode. Just as before, the problem with cylindrical modes is that the polarization also rotates about the center of the array. This necessitates a transmitarray which can respond to any polarization or a transmitarray with rotating unit cells, both of which complicate the design and may limit performance. A solution to this and what is used in this thesis is to instead use a planar wavefront mode which propagates linearly from one side of the
array to the other as seen in Fig. 4.2(b).

To elucidate the differences between this architecture and a leaky-wave antenna, consider the following discussion. If the transmitarray aperture in this design were “transparent”, i.e. does not alter the phase or magnitude of the wave as it passes through the aperture, then the behaviour would in fact be identical to a leaky-wave antenna (stemming from the leaky feed). The problem with leaky-wave antennas however are that they steer the beam by tuning the leaky-wave phase constant, limiting them to steering only in the direction with which the leaky-wave propagates. So in Fig. 4.2(a) for example, if the design was that of a reconfigurable leaky-wave antenna (tuning the phase constant of the feed wave), the beam would be a cone and steering would only result in a widening or narrowing of the cone (see Fig. 1.4). Likewise in Fig. 4.2(b), the antenna would only be able to steer a pencil beam in a single plane, despite having a two-dimensional reconfigurable aperture.

Another way to state the problem with leaky-wave antennas is that the phase gradient across the aperture is inherently tied to the leaky-wave phase constant, and the phase
Figure 4.3: Phase gradients and phase propagation direction of both the feed wave (blue) and radiated wave (red) for a center-fed leaky-wave antenna (left) and an integrated feed transmitarray (right).

Gradient is what determines the shape and direction of the beam. How the proposed integrated feed transmitarray solves this issue (and also how the integrated feed reflectarray worked as well) is by using the transmitarray aperture to completely decouple the phase gradient from the traveling/leaky-wave phase constant by altering the phase of the wave as it passes through the aperture. By giving each unit cell of the transmitarray 360 degree control of the transmitted phase, any phase gradient can be produced regardless of the phase constant of the traveling wave. In other words, it does not matter what the excitation phase is of each transmitarray element, the radiated phase for that element can be electronically tuned to any arbitrary value. This also means that the feed point of the parallel-plate waveguide can actually be at any point in the guide (at least from the perspective of the phase gradient).

Fig. 4.3 demonstrates this concept for the case of a center-fed, two-dimensional leaky-wave antenna and integrated feed transmitarray. In the leaky-wave antenna case, the center feed yields a cylindrically propagating phase constant and hence a cylindrical phase gradient (producing a conical beam). In the integrated feed transmitarray, despite a cylindrically propagating phase constant, the phase gradient can be made linear (to produce a pencil beam). Furthermore, it can be made linear in any direction, allowing it to produce a pencil beam at any elevation or azimuth angle (i.e. full-space steering).

It is also insightful to consider this design from the point of view of a transmitarray. For a transmitarray, the excitation is provided by a spatial wave which is generated from a feed antenna located a short distance away from the aperture (see Fig. 2.5). Similar to
what was done in the discussion of the design of the integrated feed reflectarray, one could imagine moving the feed closer and closer to the transmitarray aperture until it is located directly below or incorporated directly into the aperture itself. The waves would then propagate outwards from the feed, along the bottom surface of the aperture. If a method of guiding this wave is provided, then all of the feed antenna power can be constrained to travel along the surface of the transmitarray and provide element excitation. This is almost identical to the operation of the antenna in Fig. 3.1(d) except now this integrated feed is placed below the transmitarray aperture (while it was above the reflectarray aperture before). With this view the capability for full-space beam-steering is more obvious since it is identical to the method used in reconfigurable transmittarrays, only the method of feeding is different.

The advantage of this architecture is that it now combines the low-profile feeding method of leaky-wave antennas with the full-space beam-steering characteristics of reconfigurable arrays. It also does this in a much more compact and simpler manner than the integrated feed reflectarray. Since the waves traverse each structure inside the antenna only once (i.e. from feed through waveguiding structure then through aperture to free space) there is no longer any need for polarization control and polarization selective surfaces. This both drastically simplifies the design and reduces the number of interactions the waves experience inside the antenna. If the transmitarray aperture is made sufficiently compact, the antenna will also be lower profile since the waveguide can be placed directly adjacent to the aperture, as opposed to a distance away which was required with the reflectarray design.

Compared to a traditional phased array, this design offers similar beam-steering performance with the added advantage of scalability. Since the feed is a partially unconstrained feed, increasing the size of the aperture only requires the waveguiding structure to be extended, and additional transmitarray unit cells added on top. Scaling a phased array on the other hand requires significant additions to the beamforming network, often requiring additional subarray layers or TRM modules.

The design of this antenna requires several key components to be individually designed and integrated together. First and foremost, a reconfigurable transmitarray or similar structure is required to provide the proper radiated phase gradient to produce pencil beams. An integrated feed then needs to be designed in order to distribute the power across the bottom of the transmitarray aperture. Finally, a method of coupling power between the feed waveguide and the transmitarray needs to be developed. This coupling layer needs to ensure the waveguide properly excites the transmitarray unit cells, and also needs to control both the degree of coupling and provide sufficient phased decoupling such
that transmitarray tuning does not impact feed wave propagation. The following sections will discuss the design of each one of these components individually and then move on to the integration of the components to form a one-dimensional, then two-dimensional array.

### 4.2 Transmitarray Unit Cell

Since this thesis was not focused on the design of a reconfigurable transmitarray but instead a feeding system for such an array, it was decided early on to make use of currently existing reconfigurable transmitarray designs. As discussed in Chapter 2, reconfigurable transmitarrays are relatively recent, and there are only a handful of designs offering full 360 degree phase tunability within their unit cells. This thesis looked at two designs by the same author: a coupled-resonator design [29] and a guided-wave design [30], both of which are designed for an approximate 5 GHz frequency of operation. Both of these designs have also been fully fabricated and experimentally verified within the same lab as the author of this thesis, and therefore could easily be modified to accept an integrated feed without having to spend time re-fabricating and testing the array. Fig. 4.4 compares the unit cell of these two designs.

The coupled-resonator design (Fig. 4.4(a)) uses three resonant surfaces: patch, slot, patch. The power is coupled from free-space to the first patch, then couples through the slot to the second patch where it radiates on the other side. The resonant frequency of each patch/slot and thereby the phase response (at a single frequency) is tuned by var-
actor diodes incorporated across the patches/slot. The three tunable surfaces combined give a multipole transmission response, allowing a full 360 degree phase range. Despite this, the cell is quite lossy and the insertion loss varies significantly over the full phase range (between 15 to 2.7 dB).

The guided-wave design (Fig. 4.4(b)) on the other hand uses a phase shifting transmission line structure between two patches. Power is again coupled from free space to the first patch, which is then coupled to a transmission line. The transmission line uses a bridged-T phase shifter loaded with varactor diodes to perform the phase shift. The power is then coupled to a patch on the other side and radiated to free space. This design can achieve over 400 degrees of phase range with much less variation in insertion loss (between 5 and 2 dB). It is also much more compact, being only 0.17 free-space wavelengths thick. Therefore, since full-space scanning is desired which necessitates large phase ranges, and also since the design is to be as low-profile as possible, the guided-wave solution was chosen. This is also the design recommended by the author for future applications [32].

This guided-wave transmitarray unit cell was incorporated in a $6 \times 6$ element array and tested in the author’s current lab [32]. As mentioned before, this offers a significant advantage to this project as this array can be modified without having to redesign, re-fabricate and test it’s functionality. All future work in this chapter is therefore based on the assumption of using this $6 \times 6$ array. This constrains the frequency of the design to 4.8 GHz and also the size of the integrated feed which will be discussed soon.

4.2.1 Unit Cell Design and Operation

The chosen guided-wave transmitarray unit cell along with an exploded view can be seen in Fig. 4.5. The unit cell is a multilayer design consisting of six substrate layers and seven metallization layers. As was previously mentioned the cell is extremely thin, approximately only 10 mm or 0.17 free-space wavelengths at 4.8 GHz. The cell accepts linearly polarized radiation on one side of the antenna via stacked microstrip patches, couples the power to a differential transmission line, couples the power through apertures in a ground plane to another differential transmission line, phase shifts the signal using a varactor tuned bridged-T phase shifter, then couples the power back to stacked microstrip patches to be re-radiated in linear polarization.

At the core of the unit cell is the bridged-T phase shifter which can be seen in Fig. 4.6 along with its dimensions. This is a reconfigurable phase shifter implemented on two balanced microstrip lines which uses varactor diodes as tunable capacitances. The
Chapter 4. Integrated Feed Transmitarray

Figure 4.5: Transmitarray unit cell (left) with exploded view (right).

The phase shifting is accomplished via the electronic tuning of six Aeroflex/Metelics GaAs MGV100-20 varactor diodes in E28X packaging which load the bridged-Ts. These diodes, when tuned in the reverse voltage range of 0-20 V produce a variable junction capacitance of around 0.15 pF to 2.1 pF. One major advantage to this phase shifter design is that all six varactors are biased identically, therefore requiring only a single voltage bias. This bias is provided by thin bias lines etched on the same substrate which are loaded with 10 kΩ RF choke resistors. 150 pF RF shorting capacitors are also used to isolate the middle diodes from the outer diodes at DC but provide an RF short.

To decouple the input and output sides of the transmitarray unit cell and to improve symmetry, a ground plane is placed in the middle of the stack-up, and an unloaded
differential transmission line is placed on the other side of the ground plane. The two transmission lines then couple to each other via two apertures in the ground plane. Coupling to the patches is performed via proximity coupling to the balanced arms of the transmission lines. Stacked patches were used to increase the bandwidth; in this case the extra patches were placed on a separate substrate located 1.6 mm away from the first patch through an air gap. The use of stacked patches increases the bandwidth from roughly 2% to 10% at 4.8 GHz.

A complete vertical stack-up of the unit cell with dimensions and substrates can be seen in Fig. 4.7. The upper patch (stacked patch) uses Rogers RT/duroid® 5880 for its substrate, Rogers RT/duroid® 6002 is used as the substrate for the lower patch and Rogers RT/duroid® 6006 is used as the transmission line substrate. The upper patches have dimensions of 20.5 mm by 3 mm while the lower patches have dimensions of 17.5 mm
4.2.2 Unit Cell Simulation Results

The transmitarray unit cell is first replicated in Ansys HFSS in order to validate its phase shifting properties (Fig. 4.8). The transmitarray cell was originally designed and simulated in SEMCAD and thus the lumped element models used had to be adapted for use in HFSS. As done in [30], each varactor is modeled using a series resistance of 3 Ω, a series inductance of 0.4 nH, and a variable series capacitance (0.15 to 2.1 pF) in parallel with a 1500 Ω resistance (Fig. 4.9). This is different than what was used in the integrated feed reflectarray simulations in that now a large parallel resistance across the capacitance is used. This choice is made to better match the simulations performed in [30] which found good agreement using the parallel resistance. In HFSS the diodes are modeled using impedance surfaces for each resistance, capacitance and inductance with the total dimensions approximately equivalent to that of the varactor diodes (1 × 2 mm). Perfect conductor leads were connected between the microstrip and the surface impedances which were slightly elevated above the metallization layer by 0.25 mm (approximately half the height of the actual diodes).

As discussed in Chapter 2, the unit cells are simulated in an infinite periodic array environment using Floquet ports for excitation from the top and bottom and master/slave boundaries on the sides. For verification simulations the incidence angle of radiation is chosen to be broadside for simplicity. Fig. 4.10 shows the co-polarized transmission magnitude and phase response versus capacitance. As can be seen the unit cell possesses a roughly 400 degree phase range as the varactor capacitance is varied between 0.15 pF
Figure 4.9: Model of Aeroflex/Metelics MGV100-20 varactor diode for transmitarray.

Figure 4.10: (a) Magnitude and (b) phase response for transmitarray unit cell versus varactor diode capacitance.

and 2.0 pF. The insertion loss varies between 4.5 and 2 dB with an average value of 3 dB. Overall the response is very good, with ample phase range and an average insertion loss on par or better than most commerical phase shifters.

One downside which can be seen at this point is the variability of the insertion loss with bias voltage. These variations will effectively couple the radiated amplitude weighting of each array element with its phase, which in the final array will disrupt the amplitude taper and lead to pattern degradation and reduced taper efficiency. Because of the way the antenna in this chapter is fed, amplitude variations between steering configurations will also result in small changes to the total radiated power, further reducing gain. This behaviour will be identified and discussed later.

### 4.3 Feed Design and Unit Cell Modifications

The simulation results presented in the last section are for the transmitarray unit cell when operating as a transmitarray. These unit cells relied on a spatial feed provided by an antenna such as a horn located a certain distance away from the antenna (see Fig. 1.3). If the feed antenna is for example vertically polarized, all of the unit cells would be oriented
correspondingly with their patches vertical. This would ensure that all incident power (minus cross-pol introduced by the feed) would be coupled into and re-radiated by the array. In this thesis however a new feed mechanism will be developed which will require a new method of coupling power into the cells. Fortunately, the current unit cell design presented above is actually quite robust to the specific method of coupling as long as the incident electric field is polarized along the direction of the receiving patch. As we will see, only a slight modification is needed in order to couple power from a guided traveling wave to the unit cells.

4.3.1 Feed Design

In Chapter 2, it was briefly discussed how waveguide modes can often be interpreted as plane waves bouncing between two waveguide walls. This bouncing mechanism provides a means to excite the transmitarray unit cell as similarly as possible to a plane wave excitation. For a plane wave excitation of the transmitarray unit cell two things are required: a plane wave incidence angle not too far from broadside and a polarization aligned with the patches. The bouncing waveguide analogy with a $TE$ mode provides both of these and the concept is shown for this design in Fig. 4.11.

It can be seen that the upwards traveling “bouncing” plane waves in Fig. 4.11 appear almost identical to plane waves incident at a $\theta_{bnc}$ given by Eqn. 2.29. If the mode is also $TE$, then the electric fields are completely transverse. By orienting the patches in the same way (i.e. in the $\hat{y}$-direction in Fig. 4.11), the $TE$ mode almost completely replicates $TE$ plane wave incidence at an angle $\theta_{bnc}$. While in true parallel-plate or rectangular
waveguides the transverse electric field would be zero at the top and bottom waveguide walls, this would not be the case for a waveguide loaded with the transmitarray unit cell. The transmitarray unit cell would have a surface impedance different than a conductor, and the electric fields would not go to zero. Therefore coupling in the case of a loaded waveguide could still be provided.

Since, as was discussed in the previous section, a parallel-plate feed waveguide is desired, the optimal excitation mode would thus be a $TE_1$ parallel-plate mode. This is the lowest order $TE$ parallel-plate mode, and is the same mode used in the integrated feed reflectarray waveguide. The use of this mode further stresses the use of a side-excited parallel-plate waveguide as for any other excitation that does not result in planar wavefronts, the polarization would change as a function of position along the array. It was realized early on however that the practical construction and implementation of a parallel-plate waveguide would be difficult. Instead, as was done in Chapter 3, the parallel-plate waveguide is broken up into multiple linear rectangular waveguide sections with widths equal to one cell width and placed side-by-side. Fig. 3.5 showed that the $TE_1$ mode of a parallel-plate mode is identical to the $TE_{10}$ mode of a rectangular waveguide. Therefore with these rectangular waveguides placed side-by-side, a parallel-plate mode is replicated but with the fields contained within individual rectangular waveguide sections.

In future designs, keeping the parallel-plate design may be worthwhile as it may allow for additional functionality including feeding from multiple directions or different locations. One possibility would be feeding from two orthogonal directions to enable dual-polarization capabilities (with a different transmitarray unit cell). Another would be center-feeding (as was discussed previously). For the time being the rectangular waveguide design is used as it offers a simpler design and fabrication procedure.

The important parameter in the design of such a waveguide is the height of the waveguide. The height of the waveguide will determine both the phase constant and the frequency at which higher order modes appear. The height also needs to be chosen such that the waveguide can easily be excited by standard waveguide adaptors. The frequency of operation of the transmitarray is centered at 4.8 GHz and the feed should be designed around this frequency as well. WR137 with a height of 34.8488 mm is a good candidate. This waveguide has a $TE_{10}$ cutoff frequency of 4.3 GHz and a $TE_{20}$ cutoff frequency of 8.6 GHz allowing fundamental mode propagation but cutting off all higher order modes. The phase constant of the fundamental mode at this frequency is 44.5 rad/m corresponding to a bounce angle of 26.3 degrees. This is a good bounce angle as it is well within both large and small extremes (i.e. endfire/broadside angles) which allow room for possible fluctuations in the final design.
The width of the waveguide is already constrained to be equal to or less than the transmitarray unit cell spacing. This enforces the cutoff frequency for $TE_{0n}$ modes to be greater than 4.8 GHz even for when $n = 1$. This spacing thus also ensures that $TE_{mn}$ and $TM_{mn}$ modes are cutoff, since in both cases the necessity of having $n \leq 1$ puts those modes in cutoff.

Finally it should be noted that loading the rectangular waveguide with the transmitarray unit cell will change the behaviour of the waveguide. In particular the loading will cause the boundary condition on the top of the waveguide to change. This will then change both the phase constant and bounce angle inside the guide. It will be shown shortly that this waveguide can be modeled using a transverse equivalent network model in order to find both the phase constant and leakage constant. The next step is to provide a method of coupling between the unit cells and the waveguide that allows power to be coupled between waveguide and cell without significantly changing the propagation behaviour (phase constant) from a metal rectangular waveguide.

4.3.2 Modified Transmitarray Unit Cell Design

With the feed and transmitarray element chosen, the next step is to modify the transmitarray unit cell to control the coupling of power from the waveguide to the unit cell. The original transmitarray utilized patches to couple the energy from free space into the cell. Part of the reason waveguide is used as the feed is the fact that it can provide a transverse electric field similar to what the original array might see. This drastically reduces the degree of modification required for the unit cell and patches can still be used for coupling. However, the coupling layer also needs to present a boundary condition similar to what a conductive wall would present in the unperturbed waveguide, otherwise the guided mode may not propagate or the waveguide would be difficult to match/feed.

The first idea for a coupling layer was taken from the integrated feed reflectarray design and was a dipole array surface (Fig. 3.4(b)). The surface consisted of one coupling patch/dipole and several guiding patches/dipole. The coupling patch was in the exact same position as the upper patch in the orginal transmitarray unit cell and the guiding patches were added around the coupling patch to produce a boundary condition more similar to that of a conductive wall. After experimentation in HFSS, it was realized that a complete metal sheet could be used in place of the guiding patches, leaving only the coupling patch with small gaps between patch and sheet. Fig. 4.12 shows the modified unit cell with the rectangular waveguide feed and the coupling layer added. Fig. 4.13 shows a schematic of the coupling layer unit cell with dimensions. For now the length
and width of the patch on the coupling layer is kept the same as that of the original transmitarray unit cell (20.5 × 3 mm). The gaps are left as 4.75 mm as this was the spacing between dipoles used in the first several design iterations which produced good results. The metal sheet is directly attached to the walls of the waveguide, forming a complete rectangular guide over those sections. The complete waveguide with coupling layer on top is then best thought of as a rectangular guide with periodic gaps introduced with patches in the middle of the gaps.

Varactor Diode Capacitance Tuning

Fig. 4.14 shows both the transmission magnitude and phase response versus diode capacitance as well as the reflection magnitude and phase response versus diode capacitance of the modified unit cell when it is simulated in the same periodic infinite array environ-
Chapter 4. Integrated Feed Transmitarray

Figure 4.13: Schematic of coupling layer with dimensions: $w = 3 \text{ mm}$, $l = 20.5 \text{ mm}$, $g = 4.75 \text{ mm}$.

ment as the original unit cell and excited from below. Note that the waveguide is not included in this simulation, only the coupling layer, but the incidence angle is chosen to be the waveguide incidence/bounce angle (approximated as 33 degrees for now). The Floquet ports and periodic boundaries result in fields below the coupling layer being almost identical to the $TE_1$ mode in a parallel-plate guide.

Comparing the response to that of Fig. 4.10, it can be seen that the transmission phase is almost identical with over 400 degrees of phase range, while the transmission magnitude has a very similar form but is much lower in magnitude. Unlike in the original transmitarray, this low magnitude is actually desired. Remember that the waveguide is acting effectively as a leaky-wave antenna, and should only leak a small percentage of the power in each unit cell. The reflection magnitude is very large and the reflection phase is actually very close to 180 degrees. Again, this is desired as the unit cell should ideally produce a boundary condition on the upper waveguide wall as close to a conductor as possible to both minimize loading effects and to aid in matching to a WR137 waveguide.

The loading effect (percentage leakage and boundary condition) are much better represented through analysis of the waveguide propagation constant, specifically the phase constant $\beta$ and the leakage constant $\alpha$. As was done with the integrated feed reflectarray, a TEN model can be developed to determine the propagation constant in the waveguide loaded with the modified unit cell. Fig. 4.15 shows a schematic of the TEN model. In this case the model uses a short circuit at the bottom to represent the bottom conductive wall of the waveguide, a vertical air transmission line of height $h = 34.8488 \text{ mm}$ representing the height of the waveguide, a two-port network $[Z_{TA}]$ representing the transmitarray unit cell and then finally another air transmission line terminated in $Z_{RAD}$ which is the
Figure 4.14: Modified unit cell (a) transmission magnitude and (b) transmission phase as well as (c) reflection magnitude and (d) reflection phase. Cell is fed with a 33 degree incidence angle from the bottom.

free space impedance at the Floquet excitation angle (Floquet port impedance). \([Z_{TA}]\) is the impedance matrix of the modified unit cell at a specific incidence angle and the polarization of interest (TE) and is obtained directly from the previous simulation. It is a two-port network with the ports being the top and bottom TE Floquet port modes de-embedded to the top and bottom surface of the modified unit cell respectively.

Inputting the simulation data used for Fig. 4.14 into the TEN yields plots of \(\alpha\) and \(\beta\) versus diode capacitance which are shown in Fig. 4.16. First off note that these propagation values are technically for an infinitely long array where all cells are biased identically. It can now be seen that the waveguide has a leakage constant \(\alpha\) in the range of 1.2 to 2.4 Np/m, possessing relatively large variations. As mentioned in Chapter 2, this leakage constant is what determines how much power is radiated by the array and is normally chosen such that 90% of the power is leaked by the end of the array.

For the time being what is important is not the absolute value of \(\alpha\) but the variation in it. Ideally, tuning the modified unit cells should not change \(\alpha\) as changes to \(\alpha\) for one specific cell will result in changes to the excitation amplitudes of all other down-
Figure 4.15: Transverse equivalent network model for integrated feed transmitarray with:

\[ h = 34.8488 \text{ mm and } [Z_{TA}] \text{ derived from modified unit cell simulations in HFSS.} \]

Figure 4.16: (a) \( \alpha \) and (b) \( \beta \) of the integrated feed waveguide unit cell as the modified unit cell diode capacitances are tuned.

stream cells in the waveguide. Unfortunately, these variations are tied to the changes in input impedance/reflection coefficient of the modified unit cell which are a result in the changing capacitance and loss in the diodes. It is similar to how the original transmitarray experiences changes in insertion loss (transmission magnitude) as the cell is tuned. These variations will end up resulting in undesired changes in the aperture amplitude distribution (taper) which will manifest as a degraded beam pattern and reduced taper efficiency (reduced directivity). Changes to the absolute average value of \( \alpha \) will be discussed shortly.

Despite the variations in \( \alpha \), changes to \( \beta \) as the cell is tuned are actually quite minimal; the overall variation being roughly 2.6\%. This behaviour is evident in Fig. 4.14 since \( \beta \) is intrinsically tied to the reflection phase of the modified unit cell (i.e. the boundary condition) and that phase (Fig. 4.14(d)) is seen to be relatively constant. This
Figure 4.17: Transmission (a) magnitude and (b) phase through modified transmitarray unit cell versus diode capacitance for varying coupling patch length (CPL).

is very good as the primary purpose of this array is for beam-steering and changes to the phase constant as the cell is tuned would result in the phase of downstream cells being perturbed, altering the phase gradient and hence beam-pointing direction. For this waveguide, if all modified unit cells in the array are biased the same and tuned in unison, this 2.6% change in phase constant would result in less than a degree of beam squint. In a practical case however all cells will be tuned differently, meaning the changes in $\beta$ will be local (differ from cell to cell) and will likely average out to result in no beam squint, perhaps only a slight reduction in beam shape and increased sidelobes due to phase errors.

**Coupling Patch Length Tuning**

An additional degree of freedom the modified unit cell possesses is the length of the patch used on the coupling layer (henceforth named the coupling patch). The previous discussions had kept the coupling patch length fixed at 20.5 mm: the length used in the original transmitarray unit cell. Fig. 4.17 shows the transmission magnitude and phase response versus diode capacitance of the modified unit cell when the coupling patch length is adjusted between 20.5 and 21.5 mm.

As can be seen, tuning of the coupling patch length allows for the transmission magnitude to be adjusted quite smoothly without significantly changing the transmission phase of the unit cell. This is therefore the primary means of controlling the coupling to the transmitarray unit cell and therefore the $\alpha$ of the feed waveguide. It allows the coupling for each unit cell to be individually specified while keeping the exact same transmission phase response. As will be shown in the next subsection, this feature is perfect for providing amplitude tapers across the array. The small changes to the phase response
are almost inconsequential, especially considering that each cell in the final array will need to be characterized with its own phase versus capacitance curve anyways. Despite little change to the transmission phase, the reflection phase and therefore the boundary condition does change with coupling patch length. If we again make use of the TEN model, the changes to $\alpha$ and $\beta$ as a function of the coupling patch length over a larger range can be seen in Fig. 4.18.

The coupling patch length has a significant impact on the leakage constant and a modest impact on the phase constant. Again it should be noted however that this model assumes all unit cells in the array possess the same coupling patch length. The leakage constant is seen to vary from almost 0 to over 15 Np/m with patch lengths between 18 and 28 mm. Because of this drastic variability, this unit cell could be used anywhere from very short arrays to relatively long arrays and still prove the proper $\alpha$ to leak 90% of the power.

While the changes to the leakage constant are desired, changes to the propagation constant are undesired. Similar to with capacitance tuning, changes to $\beta$ result in changes to the phase gradient and ultimately lead to beam squint or phase errors. Fortunately, since changes to coupling patch length are fixed and not tunable, they can easily be accounted for through characterization of the response of each unit cell.

Overall the results of using the integrated feed with this coupling layer are quite good. Full phase range is achieved with the same amount of transmission magnitude variation as the original transmitarray unit cell. The phase constant in the guide is relatively decoupled from the capacitance tuning of the unit cells making it simple to tune and control the entire array phasing. The biggest downside is the changes in leakage constant with tuning, since this will result in changes to the array aperture distribution for different scanning angles. For the purposes of this thesis, this behaviour is acceptable. If however
for a certain application the sidelobe levels turn out to be an issue, another unit cell would need to be chosen which has less $\alpha$ variation with tuning. Lastly, the ability to change the coupling patch length provides a powerful means of specifying the fixed amplitude distribution (taper) along the array.

### 4.3.3 One-Dimensional Array Design

With the general design for the integrated feed unit cell complete using infinite array analysis, the next step is to design an actual finite linear array fed with a single waveguide. During this process other details will be worked out such as fine-tuning the coupling layer to provide the both the required $\alpha$ as well as any sort of amplitude tapering necessary. Since the integrated feed is to be developed for the already fabricated $6 \times 6$ transmitarray, the linear array was designed with 6 elements. The linear array is modeled in HFSS and can be seen in Fig. 4.19.

Waveports are used to feed the waveguide at one end and capture all remaining power at the other end. The waveguide walls are given a thickness of 1.5 mm reducing the internal waveguide width to 27 mm. The waveguide height is 34.8488 mm (WR137 height). The rectangular waveguide sections are extended away from the actual array section in order to remove excited evanescent modes at the waveports. As shown in Fig. 4.19, periodic cylindrical waveguide perturbations appear at the corners of every unit
cell along the guide. These perturbations are actually thickening of the waveguide wall to allow bolts to pass through the waveguide from top to bottom. These bolts are used in the fabrication of the array to fasten the transmitarray aperture to the waveguide. Because the perturbations are small, are located halfway between patches and only shorten the waveguide width, they have a negligible impact on feed wave propagation and unit cell performance. They are included here for completeness.

Initial simulations are first used to verify the basic leaky-wave behaviour of the antenna. Fig. 4.20 shows a plot of the normalized total directivity produced by simulation in HFSS with all unit cells biased to 1 pF such that the beam is produced solely by the leaky-wave feed behaviour. Also included in Fig. 4.20 is a corresponding pattern factor. The pattern factor is a combination of an array factor and an element factor. The array factor is a 6 element array factor phased using an $\alpha$ and $\beta$ extracted from the linear array simulation in HFSS. The $\alpha$ is extracted by analysis of the insertion loss through the waveguide and thus includes both radiation and losses. The extracted $\alpha$ is 2.79 Np/m. $\beta$ was found simply through back calculation using Eqn. 2.17 and the angle of the beam maximum $\theta_b$ found in HFSS. Since the maximum beam angle is 27.9 degrees, this corresponds to a $\beta$ of 47.0 rad/m. Other than changes between element excitations due to $\alpha$ and $\beta$, the phase and amplitude excitations were assumed uniform (i.e. no amplitude or phase errors were applied). The element factor was assumed to be a microstrip patch with dimensions of the top radiating patch ($20.5 \times 3$ mm). The expressions used for the element factor fields were taken from [5].

Firstly, the beam angle is close to what was expected from infinite array analysis. The simulated beam angle is 27.9 degrees corresponding to a $\beta$ of 47.0 rad/m while the
infinite array $\beta$ can be deduced from Fig. 4.16 (1 pF tuning) and is roughly 50.4 rad/m corresponding to a beam angle of 30.1 degrees. This beam squint however can be partly accounted for by considering the element factor. Because the patches possess a cosine-soidal element pattern, they would actually cause a slight reduction in the maximum beam angle for beams scanned off broadside. In fact, when plotting the pattern factor in Fig. 4.20, the assumed leaky angle was actually shifted upwards from the extracted value of 27.9 degrees from HFSS. The reason being that adding the element factor reduced the pointing angle by approximately 1.6 degrees. The leaky angle which gives the best correlation between pattern factor and HFSS simulation is actually 29.5 degrees corresponding to a $\beta$ of 49.5 rad/m. This is actually much closer to the expected value from infinite array analysis.

The correlation between simulated pattern and array factor is now quite good. There is a slight discrepancy in both the sidelobe level and the beamwidth. The increased sidelobe level is likely due to edge effects at both the start, end and sides of the array since the array is relatively short. The changes in beamwidth are most likely due to the use of an approximation for the element factor as a single patch. The unit cell is actually stacked patches and therefore the true element factor may be different (possibly more directive) than a single patch. Imperfect amplitude distributions due to differences between unit cells may also lead to beamwidth broadening and side lobe levels.

The return loss for the linear array is 21.5 dB and the insertion loss is 4.4 dB. While the return loss is very good, the insertion loss implies that 36% of the power passes right through the waveguide to the other end. This means that the remaining two-thirds or so of the power is either radiated or absorbed as loss. In a best case scenario with no losses the array is still only radiating two-thirds of the incident power. This suggests that the $\alpha$ needs to be properly tuned in order to increase the leakage up to over 90%.

Additionally, Fig. 4.21 shows a plot of the average surface current density on each radiating patch along the array. This average current density should roughly correspond to the excitation amplitude of that cell. As expected, the radiation amplitude of each unit cell is not the same. In a perfect leaky-wave antenna, the fields and excitations would be expected to decay exponentially according to the leakage constant $\alpha$. While the current densities do not exhibit a perfect exponential trend, they do demonstrate some decay. The deviation from a perfect exponential is likely due to the mismatches along the array when transitioning from array to rectangular waveguide at the start and end of the array as well as the fact that one third of the power passes right through the array.
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Amplitude Tapering

Overall the combination of both incomplete radiation and the exponential amplitude taper demonstrate the necessity to adjust the leakage constant. As was shown previously, this adjustment can be performed locally (in each cell) by tuning the coupling patch length. In order to maximize aperture efficiency and improve overall leakage, the coupling patch length is then simultaneously adjusted overall and taper is applied to produce an even amplitude distribution with maximal insertion loss through the waveguide. It was found through experimentation that linearly increasing the length of the coupling patches from 20.5 mm at the start to 21.5 mm at the end (0.2 mm per cell) produced over 90% power leakage with a better amplitude distribution. Fig. 4.22 shows the new average surface current densities on each element compared to the previous, un-tapered design.

The new return loss with this design is 20.2 dB and the insertion loss is now 9.2 dB. This corresponds to roughly 88% power leakage. This of course is an ideal value, and only works for this specific case where all cells are biased at 1 pF. If all cells were biased to say 0.6 pF, as can be seen in Fig. 4.17, the overall leakage would be lower. Even more importantly, if the array is tuned to create a beam at another angle, all the cell coupling magnitudes would be different, producing a different and likely unpredictable overall power leakage. Therefore tapering this array is only an approximate means of increasing the leakage and improving the aperture distribution, and should be considered for the “average” case.

Fig. 4.23 shows a plot of the new simulated total directivity with the tapered array.
compared to the total directivity of the un-tapered array. It can be seen there is a 1 dB reduction in the side lobe level, almost no change to the peak directivity and a 3 degree shift in pointing angle to 30.9 degrees. This shift in pointing angle can be accounted for by the changes to \( \beta \) introduced by the changes in the coupling patch lengths (Fig. 4.18(b)).

**Electronic Beam-Steering**

Lastly the array is configured to steer its beam from \(-45\) to \(+45\) degrees and the results can be seen in Fig. 4.24. To do this the same beam-steering algorithms are used as in Section 3.5.1. These algorithms use the transmission magnitude and phase responses in Fig. 4.14 to create phase gradients along the array while maximizing the transmission magnitude. It should be noted that only the curves for the unit cell with a 20.5 mm coupling patch are used (in this specific case). This would result in phase errors for the cells with different coupling patch lengths since the phase response does vary slightly depending the coupling patch length. Overall the beam-steering works quite well up to around 30 degrees off broadside in both directions. The gain starts to drop off significantly for beams closer to 45 degrees off broadside. More detailed analysis will be performed in the next section for the two-dimensional array beam-steering results.

### 4.4 Two-Dimensional Array Design and Simulations

With the individual linear arrays designed, the next step is combining multiple linear arrays to form a full two-dimensional array capable of full-space beam-steering. As
mentioned many times previously, one goal of this thesis is to make use of pre-existing transmitarray apertures. In the author’s lab a $6 \times 6$ reconfigurable transmitarray has already been developed [32]. Therefore six, six-element waveguides will be combined in a side-by-side fashion to form an integrated feed for this $6 \times 6$ array.

### 4.4.1 Distributed Waveguide Feed

Feeding the adjacent linear waveguides is accomplished through the use of a simple waveguide E-plane taper from an original single WR137 waveguide. Fig. 4.25 shows a top-down schematic of this array waveguide structure with finalized dimensions. The WR137 waveguide feed can be seen on the left side followed by the taper into six identical linear array waveguides. Fig. 4.26 shows the electric fields inside the waveguide when excited from the WR137 port and with the six waveguides terminated with waveports (matched loads). For the first few design iterations the entire top of the waveguide was closed with a metal wall, creating perfect metal waveguides. Also included are the perturbations in the waveguide walls to allow for future bolt placement.

Ideally this taper would feed all six waveguides with identical magnitudes and phases. Because of the taper however there is inevitably some degree of uneven power and phase distribution across the waveguides. There are several key parameters which are optimized in order to minimize this uneven amplitude and phase distribution. The first and most important parameter is the length of the taper section. Long tapers provide the flattest phase front, most equal power distribution and also the best return loss but would require more material and hence cost and weight. A compromise is chosen to reduce the taper to
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Figure 4.24: Beam-steering gain of linear array for beam angles from $-45$ to $+45$ degrees in 15 degree steps.

A length which is manageable in terms of cost and weight, but still provides a reasonable power distribution and return loss. The second parameter(s) is the separation between the end of the taper and the start of the actual array along with the starting position of the individual waveguide walls. The starting position of the waveguide walls coupled with the position of the taper end increase or decrease the effective waveguide openings and hence the power coupling to each waveguide. Fig. 4.25 shows the optimized positions. The center waveguide wall length can actually be adjusted to any value, and is later made longer than in Fig. 4.25 to improve structural support.

The resulting return loss at the WR137 port is 12.4 dB and the insertion loss through each waveguide is 8.0 to 8.1 dB, relatively equal. In the final design, with the transmitarray placed on top of the waveguide and all cells biased to 1 pF, the return loss is 10.2 dB and the insertion loss varied from 14.6 dB at the center to 15.1 dB to 16.4 dB at the edge waveguides. Introducing the transmitarray therefore results in a slightly less even power distribution. Based on this the total radiated plus absorbed power is equal to roughly 73% of the input power. The rest is either reflected at the input port or absorbed at the end of the waveguide.

As can be seen in Fig. 4.26 there is also a phase taper across the waveguides. This phase taper results in a 33 degree phase difference between the first (edge) and second waveguides and a 14 degree difference between the second and third (center) waveguides. This phase taper is inconsequential as it is fixed and is easily accounted for when the individual unit cells are experimentally characterized. In simulation, this phase difference simply needs to be known beforehand and can be incorporated into the phasing algorithms.
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Approximate Coupling Patch Locations

Figure 4.25: Top-view of entire integrated waveguide feed for $6 \times 6$ array with optimized dimensions.

4.4.2 Full Array Beam-Steering Simulations

The full array including waveguide feed and $6 \times 6$ array surface is constructed in HFSS and can be seen in Fig. 4.27. The integrated feed waveguide is kept the same except for an extension of the middle waveguide wall further into the taper section to add structural support for future fabrication. The entire array can be simulated in HFSS on a computer with 64 GB of RAM.

To determine the phasing of each unit cell for beam-steering, the same phasing algorithm is used which was used for the linear array and the integrated feed reflectarray with a few key differences. First of all, since this array is now two-dimensional and can scan in both azimuth and elevation, two-dimensional phase gradients need to be generated using the two-dimensional phase shift requirement for pencil beams Eqn. 2.8. This now takes into account steering in both $\theta_b$ and $\phi_b$. The two phase gradients in both directions are then superimposed on one another (added vectorally) to create the resulting phase gradient producing a beam at $\theta_b$ and $\phi_b$.

Another difference between the phasing algorithm used for this array and the linear array before is that different phase and magnitude curves are used for each cell depending on its specific coupling patch length (Fig. 4.17). Finally, since the waveguides are all fed with slightly different phase excitations, these differences are taken into account and compensated for by simply offsetting the chosen phase states between linear waveguides.
Figure 4.26: Integrated waveguide feed electric field pattern for optimized waveguide feed.

Fig. 4.28 shows the simulated beam-steering total realized gains (includes feed mismatch) at 15 degree intervals between $+45$ and $-45$ degrees for scan planes E, H and D corresponding to the E-, H- and D-planes of the radiating patches. In other words the H-plane corresponds to the longitudinal plane of the array (i.e. direction of waveguide propagation), E-plane the transverse plane of the array and D-plane the plane 45 degrees between E and H-planes. Overall the beam steers quite well from negative $+45$ and $-45$ degrees in all planes and therefore full-space.

In the H-plane the boresight gains are relatively constant for all scan angles while in the E- and D-planes the boresight gain decreases with scan angle. This is likely due to the element factor of the patches possessing a more isotropic pattern in the H-plane than E-plane. This is possible if the patches behave more similarly to dipole than patches, which is likely in this design since the patches are narrow and stacked. Another thing to notice is that the 30 degree beam in the H-plane is roughly 1 dB higher than all other beams. This can be explained by the fact that the 30 degree beam is close to the waveguide leakage angle. This then requires little phase difference between elements allowing them all to be phased at closely spaced points on their phase curves with maximal amplitude. This maximal amplitude increases leakage and hence the overall radiated power.

In terms of sidelobes the H-plane has the highest in-plane sidelobe level of $-8.6$ dB, E-plane has a $-9.9$ dB sidelobe level and D-plane has the lowest at $-19.5$ dB. The reason for the drastic reduction in in-plane sidelobe level for the D-plane is because the array is rectangular and produces a rectangular window effect creating the largest sidelobes in the principal E- and H-planes. Scanning in the D-plane tends to produce larger overall sidelobe levels which will be shown in the following measurements section.
A table of the simulated losses for a broadside beam can be found in Table 4.1. Based on an aperture size of $180 \times 180$ mm the maximum possible directivity is 20.2 dBi. The directivity at broadside simulated by HFSS is 20.0 dBi. This difference corresponds to a taper and phase efficiency of roughly 96.4%. A return loss of 10.5 dB yields a feed mismatch loss of 0.4 dB. The insertion loss through each individual waveguide varies from 14 to 14.5 dB with an average of 14.3 dB yielding an end leakage loss of 1.1 dB. Varactors contribute to most of the loss and in the broadside case contributed 3.3 dB. Finally substrate losses contributed approximately 0.2 dB in all cases and conductor losses were not considered in these simulations (to reduce simulation time) but are likely quite small. This results in a final boresight gain of 15.1 dB and a total loss of approximately 4.9 dB. Taking phase errors and amplitude taper into account the total gain is 5.1 dB below the maximum possible directivity yielding an aperture efficiency of 30.9%.

Other than varactor loss the end leakage contributed the most (1.1 dB) to the reduction in aperture efficiency. This suggests that further work can be performed on improving the amplitude taper to allow even more leakage while maintaining a flat amplitude distribution. The phase and taper efficiency also turned out to be much higher than expected. It is likely however that the taper and phase efficiency is less than what was found, and that the uncertainties in the simulation are on the order of the directivity loss due to these efficiencies. Overall the results of these simulations are quite good and
Figure 4.28: Simulated total gain patterns at 4.8 GHz for beams steered between +45 and −45 degrees at 15 degree steps.
<table>
<thead>
<tr>
<th>Losses</th>
<th>dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>WR137 Feed Mismatch</td>
<td>0.41</td>
</tr>
<tr>
<td>End Leakage</td>
<td>1.1</td>
</tr>
<tr>
<td>Varactor Loss</td>
<td>3.24</td>
</tr>
<tr>
<td>Substrate Loss</td>
<td>0.19</td>
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<tr>
<td>Conductor Loss</td>
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</tr>
<tr>
<td><strong>Total Loss</strong></td>
<td><strong>4.94</strong></td>
</tr>
<tr>
<td>Phase and Magnitude Errors</td>
<td>0.16</td>
</tr>
<tr>
<td><strong>Total Directivity Reduction</strong></td>
<td><strong>5.1</strong></td>
</tr>
</tbody>
</table>

Table 4.1: Losses in simulated array.

it was at this point decided to fabricate and experimentally test the array.

4.5 Array Fabrication and Measurements

The final chapter of this thesis discusses fabrication and experimental validation of the full array with integrated feed waveguide. Both the integrated feed waveguide and the coupling layer required manufacturing. The reconfigurable transmitarray already existed, therefore only modifications were required to swap the bottom layer of the array with the new coupling layer. Other than that no other modifications were made to the transmitarray itself.

4.5.1 Fabrication

The final integrated feed waveguide was designed as seen in Fig. 4.29. Bolt holes were provided on the top of the array to allow both a top metal plate as well as the transmitarray to be bolted onto and completely cover the top of the waveguide. The waveguide ends were left open however holes were provided on the end to allow a metal end plate to be bolted over the waveguide openings effectively short circuiting them. This allowed the option of either shorting the waveguides, leaving them open, or filling/covering them with microwave absorber. The choice of what to do would be determined based on how much power was leaked and whether that leaked power affected the measurements. The final section of the waveguides were narrowed to allow room for the threaded bolt holes. On the far left of Fig. 4.29 can be seen a WR137 coaxial waveguide adaptor feed. This feed was not manufactured, but was used to provide the feed for the waveguide. It was fastened to the WR137 waveguide opening at the start of the taper via two bolts.
The waveguide was machined out of a 1.75 inch thick solid aluminum block by the University of Toronto MIE Machine Shop. A picture of the fabricated waveguide with attached top plate and coax feed adaptor can be seen in Fig. 4.30. A mounting bracket was bolted to the back of the waveguide to allow it to stand upright during measurements.

The reconfigurable transmitarray [32] was modified simply by removing the bottom layer and replacing it with a fabricated coupling layer of identical size. A picture of the fabricated coupling layer can be seen in Fig. 4.31. It was etched by Candor Industries Inc. on 1.575 mm Rogers RT/duroid® 5880 substrate; the same substrate used for the original bottom layer of the transmitarray. Bolt holes were added to the coupling layer to allow it to be bolted through the waveguide walls and tightened tight against the top of the waveguide. Copper was left on almost all of the substrate such that it would act as the conductive waveguide top in non-array sections. The coupling layer was fastened to the transmitarray in the same way as the original layer using three nylon screws in each unit cell. These nylon screws protrude 1-2 mm into the feed waveguides and were not expected to cause any interference to the feed or array operation.

Finally, the fully assembled array with integrated feed waveguide can be seen in Fig. 4.32. Also shown is the patch-rotated coordinate system and E-, H- and D-planes used in measurements. A 40 pin cable connector is attached to a bus at the top of the array which supplies each of the 36 unit cells with an analog control voltage. The total
4.5.2 Voltage Control

The full array requires 36 0-20 V bias voltages to bias the varactor diodes. A new voltage controller array was developed to interface easily with MATLAB. Three Measurement Computing USB-3114 Analog Output Devices with 16 analog outputs each were used to produce analog outputs in the range of 0-10 V. The devices interface to the computer via USB and are easily controlled using MATLAB. A 36 channel op-amp array was then developed to amplify the 0-10 V signals to 0-20 V. Each channel uses a simple non-inverting op-amp with a 10 kΩ resistor in the feedback path. The PCB was fabricated by Alberta Printed Circuits and components were hand-soldered. Fig. 4.33 shows a picture of the fully fabricated op-amp array.
4.5.3 Characterization

In order for beam-steering to be conducted and the array to be properly phased, each unit cell had to be experimentally characterized with its individual phase and magnitude response. Various phenomena in the final design lead to variations in each cell’s response. These phenomena include edge effects, mutual coupling, fabrication differences (soldering, varactor placement etc.), variability in varactor junction capacitance and most importantly the position of the cell within the waveguide. Unlike in simulations where the phasing was computed based on prior knowledge of the waveguide phase constant, in the fabricated array this phase constant is unknown and possibly varies throughout the guide. Rather than measuring the phase constant, characterization of each unit cell provides the absolute phase radiated by each cell with the phase constant incorporated. In other words characterization of two cells spaced longitudinally by $d$ should yield phase curves which are offset by roughly $\beta d$.

To characterize each unit cell, the array was set up in front of an NSI near-field scanner. The scanner probe was placed less than a centimeter away from the radiating surface of the array, and aligned with the patch polarization. The array was then excited, and each unit cell biased identically. The probe then moved from patch to patch to measure the radiated field from each patch and only that patch for one specific bias voltage. The bias voltage was then changed and the probe repeated this process until all bias voltages had been swept through. Placement of the probe as close to the patch as possible was critical to ensure that the probe did not pick up the fields from other patches. The resulting data was analyzed and put into the magnitude and phase curves
Figure 4.32: Fully fabricated array with patch-rotated coordinate system and principle planes (E, H, D) used in measurements.

seen in Fig. 4.34.

Each phase/magnitude plot pair in Fig. 4.34 corresponds to a transverse row of elements i.e. elements which are positioned the same distance longitudinally from the start of the waveguide. These pairs should therefore have similar magnitude and phase plots. The pair at the top of Fig. 4.34 corresponds to elements positioned furthest from the waveguide feed, while the bottom corresponds to the closest. Overall the phase curves for each row are in good agreement with each other and experience the same phase offset. The magnitudes are a bit less predictable, but still experience similar trends for each row.

4.5.4 Beam-Steering Measurements

With the unit cells characterized the array can now be phased for beam-steering. A similar algorithm is used as was used for phasing the full array in simulation except that now only the measured unit cell characterization curves are used. These curves represent the actual radiated phase by each element, and therefore no knowledge of the phase constant inside the waveguide is needed. The required phase gradient is computed and
then the required phase for each cell is interpolated on its respective phase curve to find the proper bias voltage. The phase gradient which produces the largest average radiation magnitude across all cells is chosen, just as before.

For measurements the array was set up 180 mm in front of the same NSI near-field scanner. Measurements were made over a near-field scan window covering −60 to +60 degrees in both azimuth and elevation in the scanner coordinate system (which is simply rotated by +90 degrees in $\phi$ with respect to the simulation coordinate system). Gain comparisons were made against a standard gain horn measured in the exact same setup. Fig. 4.35 shows measured co-polarized realized gains compared against simulated co-polarized realized gains at 15 degree intervals between $+45$ and $-45$ degrees for the H-, E- and D-planes at 4.8 GHz (planes shown in Fig. 4.32). The co-polarization direction was taken as the $\hat{\theta}$ unit vector in a coordinate system with the $\hat{z}$-axis aligned along the long axis of the radiating patches (patch-rotated coordinate system also shown in Fig. 4.32). Realized gain means that mismatch or return loss is included. Fig. 4.36 shows $u - v$ plane (directional cosine) plots of the total gain patterns for all extreme steering angles plus broadside, better showing the full-space scanning capabilities. The other D-plane has been left out as measurements have not been taken for that specific plane. Table 4.2 is a table showing the desired and actual steering angles ($\theta_b, \phi_b$), co-polarized directivity (D), co-polarized gain (G), side lobe level (SLL) and cross-polarization levels (XPL) for each steering angle in all three scan planes.

Overall, the beam can be seen to steer very well in all three planes. The agreement with simulations in Fig. 4.35 is also very good. All measured patterns including the
Figure 4.34: Characterized magnitude and phase response for all array elements.
Figure 4.35: Measured and simulated co-polarized realized gains at 4.8 GHz. Beams are steered from $-45$ to $+45$ degrees in 15 degree steps.
Figure 4.36: Directional cosine plots of measured total realized gain patterns at 4.8 GHz.
The impact of these ripples is minor in the overall patterns, however it slightly skews the measurement of the maximum directivities and gains as well as the pointing angles. For example in Fig. 4.36(d) it is obvious that the beam maximum is approximately directly at broadside, however ripples in the pattern cause the maximums measured in Fig. 4.35 for the broadside beam to be slightly off broadside. The result is a relatively large degree of uncertainty in the beam pointing angles, and the values shown in Table 4.2

<table>
<thead>
<tr>
<th>Plane</th>
<th>Desired $\theta_b$</th>
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Table 4.2: Measured beam properties for each steering angle in each steering plane at 4.8 GHz.

pattern for the standard gain horn experienced small ripples. These ripples are not expected to be due to the antenna itself, but more likely an artifact of the measurement setup. One likely candidate is spurious reflections in the measurement chamber. Since the array is scanning relatively far off broadside, the scan window was made as large as possible. This means that it is catching a lot of extra power that may be reflecting off or coming from other directions within the room. Additional absorbers were added around the antenna to mitigate this but did not produce significant changes.
should be taken only as approximate. Despite this, the beam pointing angles \((\theta_b, \phi_b)\) are all relatively close to the desired values. As was found for the simulated linear and full arrays, as the beams scan further from broadside their pointing angles are “pulled” towards broadside due to the effect of the element factor. Mutual coupling between elements may also lead to phase errors causing patterns to get slightly skewed towards broadside.

The maximum boresight gain achieved across all beams is 15.7 dBi for a \((15, 0)\) degree beam. This is the highest most likely since it is closest to the leaky-wave radiation angle requiring minimal phase difference between elements and thus the configuration algorithm chooses close to maximum transmission amplitudes for each element. The second highest boresight gain is 15.6 dBi at broadside. The maximum directivity measured is 20.6 dBi at \((15, 0)\) degrees and the next highest is 20.5 dBi at broadside. The interesting thing to note is that the directivity at broadside is actually higher than the maximum possible directivity assuming a uniform aperture (Eqn. 2.11) which is 20.2 dBi. There are multiple possible reasons for this, the most likely and largest reason being simply the uncertainties in the measurement setup including: ripples in the patterns, the gain standardization process and the fact that the scanner is measuring near-fields over a truncated window and is therefore not capturing all the fields. The fact that this is occurring makes it difficult to get an estimate for the phase and taper efficiencies. However, the difference between measured directivity and gain can give an estimate for the aperture efficiency due to losses, mismatch and leakage effects. For the broadside beam this aperture efficiency including mismatch efficiency is \(-4.9\) dB. This suggests that the losses/efficiencies are in fact very much the same as in simulation (Table 4.1). Based on this, the aperture efficiency at broadside is estimated to be around 35%.

The side lobe levels are best found by looking at the entirety of the patterns rather than just the plane cuts, since the side lobe peaks may not occur in the cuts. On average each scan plane produced similar side lobe levels between \(-8.1\) dB and \(-12.9\) dB. Overall the worst side lobes were found when the beam was steered to extreme angles in the H-plane. E-plane scans experienced the best side lobes on average, particularly at extreme scan angles.

The cross-polarization levels were found to vary between \(-27.0\) dB and \(-11.5\) dB. Fig. 4.37 shows plots of the measured realized co-polarized gain compared to the cross-polarized realized gain for broadside and extreme angle scans. The cross-polarization was taken as the \(\hat{\phi}\) unit vector in the patch-rotated coordinate system. The H-plane experienced the best cross-polarization levels on average while the E- and D-planes experienced similar values. Just like side lobe levels cross-polarization levels were worst for extreme
scan angles. One possible source of cross-polarization for this array is stray radiation from the slots which couple power between transmission lines in the unit cell [32]. These aperture fields are oriented orthogonal to the patch fields, and hence would contribute to cross-polarization.

Fig. 4.38 shows a comparison of the measured co-polarized directivities with pattern factor directivities produced by the combination of an array factor with an element factor. The array factor was taken as a uniformly excited \(6 \times 6\) array with a perfect phase gradient given by Eqn. 2.8 to produce the desired beam angle. The element factor was again chosen to be a patch with dimensions of \(20.5 \times 3\) mm [6]. Overall the agreement between measured data and the pattern factor is very good, in some cases better than the agreement between measured data and simulations. The worst agreement is found at extreme angles, particularly in the D-plane where the beams are found to have a large degree of distortion and higher side lobes than the array factor.

Finally Fig. 4.39 shows a plot of the measured co-polarized gain patterns at both 4.65 GHz and 5.1 GHz compared to the patterns at 4.8 GHz. It can be seen that for H-plane scans, the beams experience squinting in either the backwards direction for 4.65 GHz or the forward direction for 5.1 GHz. This is expected as the leaky-waves propagate in the H-plane direction, and hence scanning the frequency will cause a change in this phase constant and therefore beam squint. In the H-plane it can also be seen that the upper bandwidth is larger than the lower bandwidth with reference to 4.8 GHz since a drop in gain is not seen until close to 5.1 GHz. For E-plane scans the beam does not squint, but does experience a drop in gain likely due to squint in the H-plane. The D-plane is a combination of the behaviour from both the H- and E-planes, and these beams experience squinting but to less of a degree than in the E-plane.

Fig. 4.40 shows a plot of the measured broadside co-polarized gain for a broadside steered beam versus frequency. Interpolating this plot at the 1 dB points below the center frequency gain (i.e. 4.8 GHz) yields a 7% 1 dB bandwidth or a 9% 2 dB bandwidth. This is only slightly less than the 10% 2 dB bandwidth of the original transmitarray. Thus the squinting behaviour of the waveguides only leads to a slight reduction in bandwidth. However, if this array were made larger, the directivity would increase and the beamwidth would decrease. The decreased beamwidth would amplify the effects of beam squint since the directivity falls off much more rapidly as the beam squints away from its desired angle. Therefore bandwidth may present more of a problem for larger arrays which are much more often encountered, and hence ways of improving the bandwidth through reduction of squint will likely need to be investigated.

Overall the measurement results compare well with the results of the original trans-
Figure 4.37: Measured cross-polarization realized gain compared to co-polarization realized gain at 4.8 GHz. Only $-45$, 0 and $+45$ degree beams are shown.
Figure 4.38: Measured and pattern factor co-polarized directivities at 4.8 GHz. Beams are steered from $-45$ to $+45$ degrees in 15 degree steps.
Figure 4.39: Comparison between measured co-polarized gains at 4.65, 4.8 and 5.1 GHz. Beams are steered from $-45$ to $+45$ degrees in 15 degree steps.
mitarray [32]. For the original transmitarray peak gains of 16 and 15 dBi were found at broadside at 4.7 and 5.0 GHz respectively. Interpolating these values to 4.8 GHz gives a broadside gain of around 15.7 dBi, almost identical to what was achieved in this work (within measurement uncertainty). The side lobe level for the integrated feed transmitarray is higher, with a side lobe level of roughly $-10$ dB at broadside compared to the $-15$ dB side lobe level for the original transmitarray. The reason for the increased side lobes is likely due to the fact that this array possesses a relatively uniform taper across the aperture, whereas the original transmitarray possesses a large taper due to the external feed horn. If desired, this array could produce a similar taper by proper design of the waveguide and element couplings. Cross-polarization levels are similar between the two arrays. The bandwidth is less for the integrated feed transmitarray. The original array quoted a 10% 2 dB bandwidth, while for this array the 3 dB bandwidth is roughly 9%. This is one drawback of using the leaky-wave feed; the beam will squint with frequency, lowering the bandwidth.

The main sources of loss for the original transmitarray were the taper efficiency (0.5 dB), spillover efficiency (2.8 dB) and varactor loss (2.2 dB). These values were listed for a broadside beam. For the integrated feed transmitarray, the main sources of loss at broadside were the varactor diodes (3.2 dB), mismatch (0.4 dB) and end leakage (1.1 dB). The increased varactor loss is likely due to the fact that the original transmitarray quotes the varactor loss at broadside, where minimal phasing is required and hence most diodes are phased at favorable points with lower insertion losses.

The great thing about the integrated feed transmitarray is that the end leakage loss and mismatch loss could likely be improved through further optimization of the waveg-
uide. This represents another significant advantage of this integrated feed design. The
original transmitarray would have a difficult time improving the taper and spillover ef-
ficiencies as they are a necessary consequence of using external spatial feeds. While it
was not listed in [32], the original transmitarray will also have to contend with loss and
efficiencies of the feed horn as well. Unlike the original transmitarray, the integrated feed
transmitarray has a good degree of control over the amplitude distribution and hence
the taper efficiency as well as the spillover efficiency. It is quite possible for the spillover
efficiency to be reduced to values much lower than 1.1 dB. This would be especially
ture for larger and longer arrays, where the required power at the last element would be
smaller (proportionally) and hence less risk of significant end leakage (spillover). Thus
the design would likely benefit from scaling.

4.6 Conclusions

This chapter presented the design of a new feeding method for transmitarray apertures
which reduces the overall profile while maintaining and possibly even improving perfor-
mance. The transmitarray makes use of an integrated feed waveguide which is located
below the transmitarray aperture. Traveling waves are guided through the waveguide
and leak/couple power into the transmitarray aperture. The transmitarray aperture
then phase shifts the waves and re-radiates them. While the design is similar to a leaky-
wave antenna, it makes uses of the 360 degree phase tunability which transmitarray cells
possess to perform full-space electronic beam-steering. Thus the design combines the
low-profile nature of leaky-wave antennas with the full-space beam-steering capabilities
of reconfigurable transmitarrays.

A $6 \times 6$ array was fabricated using a pre-existing full-space scanning reconfigurable
transmitarray. Experimental measurements showed the design was capable of pencil
beam-steering from roughly $-45$ to $+45$ degrees in the principle E-, H- and D-planes and
thus in full-space. The peak co-polarized gain for this array was 15.6 dBi at 4.8 GHz,
almost identical to the original transmitarray. The total losses/inefficiencies including
mismatch losses totaled close to 5.1 dB. Most of this loss is concentrated in the varactor
diodes (roughly 3.2 dB). The aperture efficiency was calculated to be roughly 35%. This
however could be further improved through optimization of the waveguide and element
couplings in the integrated feed. The 1 dB bandwidth was roughly 7%, which is less
than the original array but still quite good despite the use of a leaky-wave feed which
experiences beam squint.

Overall the design is only 47 mm thick which is 0.75 free-space wavelengths at 4.8 GHz,
offering a substantial improvement in the overall profile and occupied volume of standard transmitarray feeds. The array is also fully capable of specifying the amplitude taper across the array. This gives it the possibility to almost completely remove any taper and spillover efficiencies present in typical transmitarray designs.
Chapter 5

Conclusions

The overarching goal of this thesis was to develop lower-profile and lower complexity alternatives to current electronic pencil-beam-steering antennas while maintaining full-space scanning capabilities. This thesis achieved that goal by combining the low-profile reconfigurable apertures of reflectarrays/transmitarrays with the low-profile feeding mechanisms of leaky-wave antennas. The result was the design of two different integrated feeds for reconfigurable apertures: one for a reflectarray aperture and one for a transmitarray aperture.

Both designs utilize the same fundamental idea which is to distribute power across the aperture surface by using a leaky waveguiding structure. This leaky waveguide power distribution is similar to using an external feed antenna in that the power division is accomplished spatially, however, it requires only a fraction of the total volume since all the fields are constrained close to the aperture surface. Amplitude tapering across the array also becomes easier through control of the waveguide’s leakage constant. While the fundamental idea behind each feed is the same, each feed varies slightly in its implementation.

For the integrated feed reflectarray, the leaky waveguide is a parallel-plate waveguide placed half a wavelength above the reflectarray aperture which leaks power down onto the reflectarray. In order for the reflected power to escape, the reflectarray rotates the polarization by 90 degrees and the leaky waveguide is made transparent to this new polarization by using polarization selective dipole surfaces. The reflectarray aperture unit cell uses a new type of varactor diode tunable patch developed in this thesis capable of both 360 degree phase tunability and 90 degree polarization rotation. Simulations of a linear array at 5.5 GHz showed a $-45$ to $+45$ degree steering range with only a $2.2$ dB variation in gain. The side lobe level was quite unsteady, varying irregularly from $-7.2$ to $-14.8$ dB over the steering range. Despite a loss of approximately only
1.7 dB in the varactor diodes, the array possessed very low efficiencies which dropped the directivity from the maximum possible directivity by roughly 1.8 dB. The poor side lobe and directivity performance is likely due to the complex architecture involving multiple separate polarization controlling layers and the fact that power must traverse each layer multiple times. This causes power to get reflected, trapped and guided in unwanted ways.

The integrated feed transmitarray was then developed as a better alternative to the integrated feed reflectarray. For the integrated feed transmitarray, the leaky waveguide is a set of taper-fed rectangular waveguides placed directly below the transmitarray aperture which leak power up and into the transmitarray. This array possesses a unidirectional flow of power which removes the need for polarization control and additional layers, simplifying the design, making it even lower profile and preventing the problems of undesired reflection and guiding present in the previous design. For this thesis a pre-existing $6 \times 6$ element reconfigurable transmitarray was modified with the addition of a coupling layer to allow placement of the transmitarray directly on top of the feed waveguide. Tapering of the coupling layer allowed a relatively uniform amplitude distribution to be produced across the array. The array was both simulated and fabricated and near-field measurements at 4.8 GHz showed a beam-steering range of at least $-45$ to $+45$ degrees in all planes with only a maximum 2.7 dB variation in gain. Side lobe levels were much more constant in this design over the previous design with the side lobe level varying from $-8.1$ to $-12.9$ dB. The total average loss was 3.4 dB and was again mostly due to the varactor diodes. The complete aperture efficiency was approximately 35%, mostly due to incomplete radiation (spillover) and of course the varactor diodes. This spillover efficiency could likely be improved through further optimization of the waveguides and coupling layer. The 1 dB bandwidth was 7%.

Both designs demonstrated the feasibility of full-space pencil-beam steering with integrated feeds and reconfigurable apertures. In the process, the three objectives set forth in Chapter 1 have been accomplished:

1. An integrated feeding mechanism (leaky waveguide) was developed which can feed reconfigurable reflectarrays and transmitarrays in a low-profile manner while maintaining full-space pencil-beam-steering characteristics.

2. A new 360 degree phase tunable, 90 degree polarization rotating reflectarray unit cell was designed to work with the integrated feed for the integrated feed reflectarray. For the integrated feed transmitarray a pre-existing reconfigurable transmitarray was modified to incorporate the integrated feed. This demonstrates that both new and old designs can make use of this feeding architecture.
3. A prototype integrated feed transmitarray was fabricated and tested and full-space electronic pencil-beam-steering was demonstrated.

Comparing the two designs, the integrated feed transmitarray performed much better than the integrated feed reflectarray in almost all categories. First of all, the integrated feed transmitarray was lower profile, with a height of only 0.75 free-space wavelengths as opposed to the 1.17 wavelengths required for the integrated feed reflectarray. The integrated feed transmitarray was much simpler to design as it required no specialized polarization selective or polarization rotation surfaces. This also allowed it to make use of a pre-existing reconfigurable transmitarray, drastically improving its utility and practicality as it can be applied to other reconfigurable apertures with minimal modification. The radiation performance was also much better with the integrated feed transmitarray: it experienced less variation in gain and side lobes while scanning than the reflectarray counterpart, likely due to the decoupling between structures and uni-directional flow of power. The only disadvantage to the integrated feed transmitarray was the higher overall loss. This loss however can be attributed to the chosen transmitarray unit cell, and could always be replaced with a lower loss variation if desired. When considering future implementations of the integrated feed design, the integrated feed transmitarray is far superior as it only requires simple modifications of already existing apertures.

Comparing the integrated feed transmitarray to the original transmitarray, the integrated feed architecture produced almost identical gains in a much lower profile package. Both designs were limited by losses in the varactor diodes since they relied on the same phase tuning unit cell. For the original transmitarray, the other main source of gain reduction was the taper and spillover efficiencies introduced by the external feed. With the integrated feed transmitarray however, proper design of the coupling layer allows almost complete control of the taper and hence amplitude distribution across the array. This allows both taper and spillover efficiencies to be improved dramatically, another significant advantage of the integrated feed transmitarray.

Finally, it is at this point worth pointing out the advantages of this design over a traditional phased array. The true advantage of this design over a phased array lies in its feed simplicity and hence scalability and cost. To produce a larger phased array, it is not as simple as just adding extra elements. The beamforming networks need to be expanded, dramatically increasing losses, size and costs. With the integrated feed transmitarray however, the only thing which needs to be changed is the coupling magnitudes on the coupling layer. The aperture can be made as large as possible simply by adding additional elements and extending the length of the waveguides. The coupling layer would then be adjusted to get the proper amplitude distribution. With the current design the array
would also require a larger taper section. Future designs however, as discussed in section Section 5.2, could make use of a center feed, getting rid of the need for a taper altogether.

5.1 Contributions

Research on the integrated feed reflectarray was presented at the following conference:


Research on the integrated feed transmitarray is currently being written to be published in the IEEE Transactions on Antennas and Propagation.

5.2 Future Work

The author has identified three major drawbacks which limit the practicality of the integrated feed design developed in this thesis. These drawbacks pertain specifically to the integrated feed transmitarray since it has been deemed the recommended design. These drawbacks are:

1. The use of rectangular waveguide adds significant bulk and weight to the array despite the reduction in overall profile. The waveguide is also much thicker than the thickness of the transmitarray aperture.

2. The necessity of requiring a specific linear polarization to feed each transmitarray unit cell constrains the feed to a side excited, linearly propagating wave. This forces the requirement for a feed taper on the end of the array, adding additional bulk and increasing the profile.

3. The integrated waveguide is dispersive therefore naturally leading to a narrow bandwidth.

To deal with these issues and to further improve the practicality of this integrated feeding mechanism, the following future work is proposed.

The first course of future work is to further reduce the profile of the integrated feed as well as reducing the weight. This thesis used rectangular waveguide operating in the $TE$
mode to provide coupling to the transmitarray unit cells. While this worked as a proof-of-principle, the rectangular waveguide needs to be roughly half a free-space wavelength thick, which is significantly taller (almost three times more) than the transmitarray aperture. The waveguide is also constructed from a solid block of aluminum which while sturdy, is quite heavy, not easily scalable and would limit its applicability to certain applications such as satellite communications.

There are two possible solutions which could be pursued. The first is to keep the same $TE$ mode but utilize artificial surfaces such as artificial magnetic conductors (AMCs) formed with metalized substrates to shorten the height of the waveguide while simultaneously reducing the weight. Another option is to remove the need for a $TE$ wave, and instead use other forms of waveguide which are lower profile but support other modes such as $TM$ or $TEM$. This could also be combined with artificial surfaces for further reduction in weight. This option will however require different ways of coupling power to each unit cell, and is thus tied to the solution of the second drawback.

The second drawback is that the current unit cell and coupling layer only allow coupling for one type of polarization using $TE$ waves. A future course of work would be to develop an improved coupling layer which can couple from other modes such as $TM$ and also be indiscriminate to the exact direction of polarization of the wave below the unit cell (i.e. the direction of wave propagation). Allowing the coupling layer to couple power from $TM$ waves would enable different waveguides to be used such as substrate integrated waveguide, which would dramatically reduce the profile of the array. Allowing the coupling layer to couple to any polarization direction for either $TE$ or other modes would allow different excitation points to be used for the feed. Currently a side excitation with a bulky taper is used to develop a planar wavefront with the same polarization direction across the entire array. If the cell could couple any polarization, center feed points with cylindrically propagating waves and circular wavefronts could be used to feed the array without requiring additional power dividing sections (center feeds divide power equally across all angles). An example could be a circularly polarized coupling layer which couples to any linear polarization and converts the coupled power to a linear polarization to be used inside the unit cell.

The last drawback is the bandwidth of the feed. Since the rectangular waveguide currently used is dispersive/leaky, beams produced by the array will have a natural tendency to squint with frequency. While the original transmitarray unit cell has a $2\,\text{dB}$ bandwidth of $10\%$, the integrated feed transmitarray has a $3\,\text{dB}$ bandwidth of only approximately $9\%$ due to squint. To compensate for this, one avenue of future work would be to investigate leaky feeds which have slower varying dispersion curves (phase
constant changes less with frequency) or even linear dispersion curves in the bandwidth of interest. This would cause the squint due to the feed to be either slower or non-existent over the bandwidth of interest, effectively widening the bandwidth.

Finally, one interesting outcome of the integrated feed transmitarray design was its ability to control the leakage constant and hence the amplitude distribution across the array. This was done by tuning the length of the coupling patches by small amounts. If reconfigurability could be added to these coupling patches, for example by adding varactor diodes along their length, the array would be capable of both reconfigurable phase and amplitude distributions. This would allow the array to dynamically compensate for the nonuniform phase-magnitude characteristics, control the amplitude taper and control the overall size of the aperture (and hence the directivity). The ability to control the overall size of the aperture is an interesting characteristic which is not commonly seen in modern antenna designs. Coupling this characteristic with full phase and taper reconfigurability would be a very exciting research topic.
Bibliography


