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Ferrite and Dielectric Antennas for Personal Communications

by

Aldo Petosa, M.Eng.

A thesis submitted to the
Department of Electronics
in partial fulfillment of the requirements
for the degree of
Doctor of Philosophy
in
Electrical Engineering

Ottawa-Carleton Institute for Electrical Engineering
Department of Electronics
Carleton University
Ottawa, Ontario

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**"FERRITE AND DIELECTRIC ANTENNAS
FOR PERSONAL COMMUNICATIONS"**

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in partial fulfilment of the requirements
for the degree of
Doctor of Philosophy
in
Electrical Engineering

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Abstract

In this thesis, several novel ferrite antennas are presented, which possess active properties based on the magnetic field dependence of the effective permeability of ferrite materials. By the application of a magnetic field, properties such as phase control, polarization agility, and frequency agility are demonstrated experimentally, using prototype antennas designed to operate at microwave and millimeter-wave frequencies.

Two classes of ferrite antennas are examined: the ferrite resonator antenna (FRA) and the ferrite rod antenna. The ferrite resonator antenna is a novel antenna which behaves similar to the recently introduced dielectric resonator antenna (DRA), with the additional feature that the polarization or the resonant frequency of the FRA can be controlled with an applied magnetic field. The ferrite rod element introduced in this thesis is a novel antenna which combines the radiation characteristics of a dielectric rod with the capability of magnetic phase control. A linear array of ferrite rod antennas is designed and fabricated to demonstrate electronic scanning using this magnetic phase control capability.

The performance of these novel ferrite antennas is demonstrated using several prototypes and is characterized in terms of such parameters as far-field patterns, radiation efficiency, and impedance bandwidth, through extensive experimentation. In addition, design equations are presented for predicting the resonant frequencies of the FRAs, the frequency shift with applied bias, and the radiation patterns of FRAs and ferrite rods. The Geometric Theory of Diffraction is used to predict the effects of finite ground planes on the radiation patterns of FRAs. Finally, several linear arrays of DRAs are investigated and their performance as potential antennas for personal communications is evaluated.

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List of Symbols

Chapter 2

dB	decibels
DRA	dielectric resonator antenna
GaAs	gallium arsenide
GHz	10^9 Hertz

Chapter 3

3.2.1.1

ϵ_r	relative permittivity
a,b,d	physical dimensions of rectangular dielectric resonator antenna
c	speed of light in free space [m/s]
f_o	resonant frequency [Hz]
k	wavenumber $2\pi/\lambda$ [rad/m]
k_o	free space wavenumber [rad/m]
k_x, k_y, k_z	separation constants in x-, y- and z-directions, respectively [rad/m]

3.2.1.2

h	height of a cylindrical dielectric resonator antenna
r	radius of a cylindrical dielectric resonator antenna

3.2.2

(θ, ϕ)	far-field spherical co-ordinate system
$ep(\theta, \phi)$	far-field normalized element pattern of a DRA
E_o	constant representing the electric field [V/m]

3.3.2

λ_z	guided wavelength in a dielectric rod
E^y_{pq}	E^y field component for mode (p,q) in a rectangular dielectric rod
a	cross-sectional width of a dielectric rod
b	cross-sectional height of a dielectric rod
k_z	guided wavenumber [rad/m]
p,q	mode numbers in a rectangular dielectric rod

3.3.3

L	length of a dielectric rod antenna
$P(\theta, \phi)$	far-field normalized radiation pattern of a dielectric rod
$sp(\theta, \phi)$	space factor of a dielectric rod

3.4.3

μ_r	relative permeability
---------------------------	------------------------------

T_c	Curie temperature [deg]
3.4.4	
β	propagation constant [rad/m]
γ	gyromagnetic ratio (2.8 MHz/ Oe)
κ	Polder permeability tensor element
μ	Polder permeability tensor element
$[\mu]$	Polder permeability tensor
	$[\mu] = \begin{bmatrix} \mu & -j\kappa & 0 \\ j\kappa & \mu & 0 \\ 0 & 0 & \mu_o \end{bmatrix}, \text{ for a z-directed bias}$
μ_{eff}	effective permeability
μ_o	permeability of free space [H/m]
$\mu_{\pm}$	circular-polarized effective permeability
ω	radian frequency [rad/s]
ω_o	gyromagnetic frequency [rad/s]
$\mathbf{b}$	magnetic flux density vector [G]
$b_{\pm}$	circular-polarized magnetic flux density [G]
$\mathbf{h}$	magnetic field intensity vector [Oe]
$h_{\pm}$	circular-polarized magnetic field intensity [Oe]
H_o	magnitude of the applied magnetic field strength [Oe]
L	length of ferrite
lh_{cp}	left hand circular polarization
M	magnetization [G]
rh_{cp}	right hand circular polarization
3.4.5	
ϵ	complex permittivity [F/m]
ϵ'	real component of complex permittivity [F/m]
ϵ''	imaginary component of complex permittivity [F/m]
$\tan\delta$	loss tangent
μ	complex permeability [H/m]
μ'	real component of complex permeability [H/m]
μ''	imaginary component of complex permeability [H/m]
j	$\sqrt{-1}$
ΔH	linewidth [Oe]
3.4.5.5	
C	a constant dependent upon the shape factor used to determine h_{crit}
ΔH_k	spinwave linewidth [Oe]
h_{crit}	critical field strength where the onset of spinwaves occurs [Oe]

$4\pi M_r$	remnant magnetization [G]
$4\pi M_s$	saturation magnetization [G]

3.4.6

B_r	$4\pi M_r$ [G]
H_c	coercive field strength [Oe]
M_r	remnant magnetization
R	$B_r / 4\pi M_s$

3.4.7

G	gauss (1 G = 10 000 Wb / m ²)
Oe	oersted (1 Oe = 1/(0.004 π) A/m)

Chapter 4

4.1

FRA	ferrite resonator antenna
-----	---------------------------

4.2.2.3

dB _i	decibels with respect to isotropic
-----------------	------------------------------------

4.4.2

η	radiation efficiency
Q	quality factor
Q_{ant}	Q factor for the ferrite resonator antenna
Q_{loss}	Q factor representing the losses in the ferrite resonator antenna
P_{rad}	radiated power [W]
P_{loss}	power loss [W]
R_{in}	input resistance [Ω]
R_{rad}	radiation resistance [Ω]
R_{loss}	loss resistance [Ω]
E	energy stored per cycle [J]
Δf_{rad}	10 dB bandwidth of the FRA [Hz]
Δf_{cap}	10 dB bandwidth of the FRA with a Wheeler cap [Hz]

Chapter 5

E_t	total electric field [V/m]
E_{go}	electric field due to geometric optics [V/m]
E_d	diffracted electric field [V/m]
$D(d, \psi, n)$	diffraction coefficient
GO	Geometric Optics
GTD	Geometric Theory of Diffraction
MoM	Method of Moments

Chapter 6

η	radiation efficiency
θ_e	half-power beamwidth in the E-plane [degrees]
θ_h	half-power beamwidth in the H-plane [degrees]
$d_{\max}$	maximum dimension of a tapered ferrite rod cross-section
$d_{\min}$	minimum dimension of a tapered ferrite rod cross-section
D	antenna pattern directivity [dB]
G	antenna gain [dB]
N	number of turns in a solenoid
I	applied solenoid current
L_d	insertion depth of ferrite rod into the waveguide
L_p	length of the solenoid
s_p	separation of Faraday suppression plates

Chapter 7

s	inter-element spacing of DRAs in linear array
DIG	dielectric image guide

1. Introduction

1.1 Overview

Due to the ever-increasing demand for portable, ground-based satellite communication services, saturation of the current frequency spectrum allocation at L-band is forecasted. Recently, frequency bands in the K/Ka spectrum (20/30 GHz) have been allocated for civilian satellite communications to alleviate the predicted congestion at L-band. Several advantages for portable satellite communications are to be gained by operating at these higher frequencies. Larger bandwidth is available, allowing data to be transferred at higher rates, thus permitting increased versatility and more elaborate features for consumers. The size of antenna elements will be significantly smaller, thus facilitating the realization of portable, personal satellite communications terminals. Finally, depolarization effects on signals due to Faraday rotation in the ionosphere are negligible above about 1 GHz¹. This removes the requirements for circular polarization at millimeter-wave frequencies, thus eliminating extra complexity in the antenna.² These advantages have sparked research aimed at the development of antenna arrays for use at these millimeter-wave frequency bands.

The design of phased arrays at millimeter-wave frequencies provides antenna engineers with many difficult challenges. To date, there has not yet been a satisfactory solution that overcomes all the difficulties. The objectives of current research are to develop a reliable phased array at a

¹ Depolarization may still occur due to hydrometeor (rain, snow) absorption and scattering in the troposphere.

² The use of linear polarization will, of course, introduce the need for alignment of the antenna in the direction of the polarized wave.

relatively low cost. The antenna must be light weight and compact to realize portability for use in a personal communications terminal. A phased array is required to allow for electronic beam scanning capability, necessary for satellite tracking. High gain is necessary to compensate for the increase in path loss. Finally, reliability and low cost are essential for commercial viability.

Current research in this field has focused on active array architectures, consisting of passive antenna elements combined with active components such as phase shifters or amplifiers. The research carried out in this thesis, on the other hand, focuses on active antenna elements which consist of ferrite materials. These elements may be considered as alternative candidates to the passive printed elements currently under investigation, for phased array applications.

1.2 Objective

The objective of the research reported in this thesis is to introduce various novel ferrite and dielectric antennas as potential alternative candidates to printed antennas for personal communication applications. The properties of these novel antennas are to be investigated by designing, fabricating, and testing several prototypes. These prototypes will be proof-of-concept models, used to demonstrate the various active properties which can be obtained by applying a magnetic field to the ferrite antennas. They will also be used to assess performance, in terms of such antenna parameters as gain, radiation efficiency, cross-polarization, impedance bandwidth, and electronic beam steering. These prototype antennas will also serve to evaluate the performance of various feeding mechanisms, by assessing their practicality and exposing any manufacturing difficulties.

The data collected by evaluating the performance of these prototype models will be used to determine the important physical parameters required to optimize the antenna performance, and to predict what limiting bounds on performance can be expected. This information will be used to evaluate the potential of these novel antennas as candidates for various applications in personal communications.

1.3 Scope of Research

Scientific journal publications in microwave and antenna topics are currently filled with an increasing percentage of papers based on numerical methods and techniques for the analysis (and to a lesser extent, the design) of antennas. It seems that fewer and fewer papers are devoted to experimental methods or practical implementations of antennas. Several reasons can be put forward to explain this trend. The primary ones can be attributed to the lack of financial resources to cover the increasing cost of test facilities and research equipment required to carry out experimental investigations. Experimental research is also time-consuming, dependent upon many factors not under the control of the researcher, and often fraught with annoying and unforeseen problems associated with the fabrication or measurement process itself. Conversely, only a fraction of the cost and time is required to undertake research based on numerical techniques, and many research institutes can readily afford the relatively inexpensive computing facilities. Although research on numerical methods and the solutions it provides represent an important contribution to this field, published results often do not provide the general reader with any useful information on the design and fabrication of antennas, or any insight into the underlying principles of their operation.

The research carried out for this thesis, unlike the general trend towards numerical methods, strongly emphasizes experimental techniques. This emphasis was made possible due to the excellent research and test facilities available to the author. This thesis undertakes to demonstrate the validity and usefulness of novel ferrite antennas based on measurements of antenna performance, and to present practical configurations. Antenna element designs were based on relatively simple, yet accurate, engineering models which provide physical insight into the behaviour of these antennas and can be used as design tools for element development. These models allowed for rapid predictions of antenna behaviour, providing first-order approximations that exhibited good agreement with measured results.

The goal of this research was to demonstrate the capability of these novel ferrite and dielectric antennas, to highlight potential applications, and to develop array configurations for use in these areas. This work was carried out in part to stimulate interest in ferrite and dielectric antennas and to promote further research in this area. The emphasis was placed on measuring the performance of several individual antenna elements and arrays of elements, giving a detailed account of their geometry, feeding mechanisms, overall performance and practical limitations. This thesis work was carried out to lay the foundation for future research of a more rigorous nature which would delve further into the optimization of these elements.

1.4 Organization

This thesis is divided into several chapters. Chapter 1 provides an overview of the thesis work and outlines the objectives and scope of the research. Chapter 2 contains background information on some of the array architectures under recent investigation. The major design challenges are detailed along with advantages and disadvantages of the various architectures. Chapter 3 contains a review of dielectric resonator antennas (DRAs), dielectric rod antennas, and microwave ferrite materials. Since the ferrite antennas in this thesis share many of the properties and characteristics of dielectric antennas, the field configurations, modes of operation, and design equations presented in this chapter are directly applicable to the ferrite antennas with little or no modification. Chapter 4 introduces novel ferrite resonator antennas, describes their behaviour in the presence of a magnetic biasing field, and presents both measured and predicted results including radiation patterns, efficiency, polarization agility, and tuning of the resonant frequency. Chapter 5 discusses the effects of finite ground planes on ferrite resonator antenna patterns. The Geometric Theory of Diffraction is used to predict these effects and agreement with measured patterns is found to be quite good. Chapter 6 describes the research on ferrite rod antennas. A novel element is introduced which combines the antenna with an analog phase shifter. Radiation patterns and phase shifter results are presented along with design equations. Chapter 7 presents several linear arrays of DRAs and ferrite rods. The performance of these arrays is investigated with design equations and measured data. Finally, Chapter 8 summarizes the research and discusses future work.

2. Review of Arrays for Personal Communications

2.1 Introduction

This chapter examines some of the recent research and development of arrays for portable satellite and personal communications applications at millimeter-wave frequencies. It is intended to underline some of the major challenges and technical difficulties in the development of such arrays, and it sets the stage for the research carried out in this thesis.

A detailed treatment of millimeter-wave antennas for various applications is given in [1]. This chapter is focused only on arrays for millimeter-wave personal or satellite communication applications. A survey of the more promising architectures is presented, with emphasis on the many challenges and difficulties which have been encountered. Most of the current research on arrays for millimeter-wave satellite communication applications can be classified into two categories: printed array architectures and non-printed architectures. Research on printed element phased arrays consists of the integration of printed antenna technology with active devices using a monolithic or hybrid process. Research on non-printed architectures involves the periodic loading of transmission guides such as microstrip lines or dielectric guides to excite radiated waves. Advances and obstacles in the development of these architectures are detailed in the following sections.

2.2 Printed Array Architecture

2.2.1 Overview

The idea of integrating printed antenna technology with active components using a monolithic or hybrid process is a very attractive one. Printing the antenna array, r-f feed network, phase shifters, and amplifiers using a monolithic process onto a single substrate layer is desirable for many reasons: ease of manufacturing, reliability and reproducibility, low labour costs, and facility for high-volume production. As outlined in the following section, however, there are severe limitations to a single-layer monolithic process that exclude this approach as a viable solution. The direction of research has thus shifted to a hybrid technique, which consists of several planar substrate layers of various configurations. Although multi-layered architectures overcome many disadvantages of the single substrate approach, other problems arise due to added complexity. The single- and multi-layered printed architectures are now examined in closer detail.

2.2.2 Single-Layer Monolithic Phased Arrays

The monolithic approach involves the printing of radiating elements, feed network, and active components onto a single substrate layer using a monolithic process [2,3]. In such an architecture, amplifiers are required for each printed antenna element (or sub-array of elements) to compensate for the increased circuit losses at these frequencies [4]. A monolithic process on a single substrate layer would have several manufacturing advantages, including low labour costs and a potential for high-volume production. There are, however, severe and prohibitive limitations inherent in this architecture. One disadvantage is that of limited surface

area for the placement of the radiating elements, feed network, and active components. This leads to crowding, complex routing of feed lines, and a loss in flexibility for the placement of the elements to optimize performance. Also, spurious radiation from the feed lines or active components may interfere with the desired radiation patterns. Serious as these problems are, they might be overcome using imaginative design techniques. There is, however, a prohibitive fundamental problem severely limiting the performance of the antenna array: the conflicting substrate requirements of the radiating elements and active devices. The radiating elements most suited for this application are printed dipoles or patches.¹ Due to their resonant nature, these printed elements have inherently narrow bandwidths. The bandwidths of these antennas may be improved by placing the elements on a thick substrate of low permittivity. Unfortunately, active devices perform better on a thin substrate of high permittivity, such as a GaAs substrate. These conflicting substrate requirements are irreconcilable due to the severe deterioration in performance of either the printed elements or active components. Thus with current active component technology, the single-layer printed substrate architecture does not provide a viable solution to the millimeter-wave phased array problem.

2.2.3 Multi-Layer Printed Phased Arrays

The problem of conflicting substrate requirements can be resolved by using two or more substrate layers [2-4]. In such an architecture, the top substrate would consist of a low permittivity dielectric onto which the

¹ Slot elements can also be used, but are less attractive since they require cavity backings to suppress unwanted parallel plate modes [4]. These cavities are formed by plated via holes or shorting pins, adding to manufacturing complexity, thus reducing reliability and increasing costs.

radiating elements (and possibly the feed network) would be printed. (Alternatively, a second layer could be used for the feed network.) The second (or third) layer would consist of a high-permittivity substrate containing the active components. Thus, both the low permittivity requirement of the printed elements and the high permittivity requirement of the active components would be met. This architecture also provides a certain degree of isolation between layers, reducing or eliminating the amount of spurious radiation from the active components and feed lines that would deteriorate the desired radiation patterns. A further advantage is the additional surface area available on the antenna aperture, allowing greater freedom and flexibility for the optimal placement of the elements. This solution however, comes with its own set of problems. This architecture involves much more complexity, and issues such as board alignment, maintenance of good electrical connections between layers, thermal expansion, and heat removal must be addressed, all of which add to the cost of the antenna.

Many of the problems with the hybrid architecture can be traced to the amplifiers required to overcome the increased conductor losses of the feed network and of the elements themselves at millimeter-wave frequencies. The incorporation of these components necessitates additional substrate layers, thus contributing to a significant increase in both complexity and cost of the antenna array. If there were a way of reducing these losses, the number of amplifiers could be reduced, if not eliminated altogether, thus simplifying the design and decreasing the cost significantly. To reduce these losses, alternate guiding structures and/or radiating elements are required that provide significantly lower loss than the microstrip feed networks and printed elements used in the multi-layered

approach. Some of these alternate guiding structures are investigated in the second category of research on millimeter-wave antennas: non-printed element array architectures.

2.3 Non-Printed Element Array Architecture

2.3.1 Overview

The second class of millimeter-wave phased arrays is an antenna architecture using non-printed elements or non-printed feeds or both. Examples of the most prominent of these architectures include leaky-wave antennas, dielectric aerial arrays, dielectric resonator arrays, and waveguide slot arrays. The more promising architectures are briefly discussed in the following sections.

2.3.2 Leaky-Wave Antennas

Leaky-wave antennas are a class of travelling-wave antennas consisting of a periodically loaded guiding structure. This periodic loading with discontinuities couples out appropriate levels of energy from the guided wave to form the desired beam [5,6]. The leaky-wave antenna holds significant potential, for if the appropriate guiding structure is used, there will be little loss from the feed network. Thus the need for amplification of the signal at each radiating element will be minimized.

A promising class of leaky-wave structures for millimeter-wave applications is the dielectric waveguide. Dielectric guides come under various configurations [7-12], the most common of which are shown in Figure 2.1. The common feature of these dielectric guides is the confinement of the energy within the dielectric layer. Therefore, assuming a low-loss dielectric material, the losses through the guide will be

Figure 2.1 Various dielectric waveguide configurations

significantly reduced compared to microstrip since the conductor loss component will be small. (A comparison of the losses in various guide structures is given in [7].) To create a leaky-wave antenna from a dielectric guide, discontinuities are added to perturb the confined surface wave and

excite a radiated wave. The most common techniques involve printing metal strips on the guide or cutting grooves in the guide surface. Several linear arrays have been fabricated using these techniques [13-22]. Unfortunately, for most of these arrays, the radiation beam can be steered only by frequency scanning, which is not desirable for many applications, including personal and portable satellite communications.

Some attempts have been made to scan the beam at a fixed frequency [23-27]. In one design, p-i-n diodes were added to the side walls of a strip-loaded dielectric guide antenna. The phase of the guided wave was altered by adjusting the bias current of the diodes, thus obtaining electronic beam scanning [23]. However, this method had its limitations in that only two adequate beam positions were obtained: one when the diodes were turned off completely and one when they were turned on fully. At intermediate bias current levels, the insertion loss of the diodes was unacceptably high, resulting in severe gain degradation of the radiation pattern.

Other attempts for fixed frequency beam scan involved the use of grooved ferrite and dielectric waveguides immersed in a d-c magnetic field [25-26]. The beam was shown to scan smoothly as a function of the applied d-c magnetic field. However, the required magnetic field was generated by placing the ferrite waveguide between the poles of a large electromagnet: satisfactory for laboratory tests but not suitable for practical applications. This problem was resolved with the development of a phase-scanned grooved ferrite rod linear antenna [27], where the required axial d-c magnetic field was generated by a voltage-activated switching coil wrapped around the ferrite rod, thereby eliminating the need for an external electromagnet. The ferrite rod was sandwiched between metal plates, to suppress unwanted Faraday rotation of the electric field through the

magnetized ferrite. A sector scan of approximately 15° was reported at 36.25 GHz. Over this sector however, a peak gain variation of nearly 2 dB was observed, and no mention was made of the antenna efficiency. Also, scanning was limited to a single plane.

Finally, a planar array consisting of an array of dielectric guides sandwiched between ferrite substrates has recently been reported [28]. Radiation is achieved through an array of 22×55 printed dipoles, etched on the upper ferrite substrate. Beam steering in two dimensions is achieved by the appropriate biasing of the ferrite material. The array, operating at 37.5 GHz, produces a pencil beam with $4^\circ \times 5^\circ$ beamwidths, and is capable of scanning 20° and 40° in the two orthogonal planes, with a gain of approximately 25 dBi. Reported losses were on the order of 3-4 dB. This technology appears to have significant promise.

2.3.3 Slotted Waveguide Arrays

Slotted waveguide technology is well established, dating back to the 1940s. Slotted waveguides have traditionally been used at X-band for radar applications, offering advantages such as low profile, high gain, high power handling, and low sidelobe performance. One disadvantage of slotted waveguides is their relatively narrow bandwidth (about 2%) [29,30]. Also, at millimeter-wave frequencies, manufacturing problems arise when attempting to meet the required mechanical tolerances. A 400-element slotted waveguide planar array at K-band has been reported, but again, beam steering was achieved using frequency scanning [31]. An electronically scanned K-band planar phased array using slotted waveguide was developed for spaceborn applications [32]. A $\pm 50^\circ$ beam scan was achieved with gains of around 36 dBi, by using 49 parallel edge-slotted

travelling-wave array elements, with each array having 36 slots. Beam scanning was achieved using ferrite phase shifters and a 300 MHz bandwidth was obtained. The overall array was 36.4 cm x 40.2 cm, with a weight of about 4.53 kg. Although the performance of this antenna is quite good, there are high manufacturing costs involved, limiting the antenna to space-based applications.

2.3.4 Dielectric Rod Antennas

A dielectric rod antenna consists of a rod of low-loss dielectric material, usually of rectangular or circular cross-section, which is tapered along its length to produce end-fire radiation. Dielectric rods offer several advantages at millimeter-wave frequencies including high gain and low losses [33,34]. Feeding of dielectric aeriels is usually accomplished by inserting the rod into the mouth of an open waveguide [35,36]. Feeding by slots, probes, or rings is also possible [37-39]. A 16-element linear array of rods has been reported with $\pm 30^\circ$ beam scanning, achieved using ferrite phase shifters [40]. One advantage of dielectric arrays is that fewer rods are required to attain the same gain as printed elements or slots, due to their higher directivity. Also, the phase shifters can be built into the waveguide feeds of the dielectric rods. One disadvantage with these dielectric rods is the need for waveguide feeds, which add to the weight and volume of the array.

2.3.5 Dielectric Resonator Arrays

One of the more recent antennas under investigation is the dielectric resonator antenna (DRA) [42]. DRAs have been shown to outperform microstrip patches in many cases, with advantages such as larger impedance

and pattern bandwidths, higher radiation efficiency, versatile coupling schemes, and lower losses. DRA elements are discussed in detail in Chapter 3. Most research has concentrated on individual elements, but recently, linear and planar arrays of DRAs have also been reported [41,101], which may prove suitable for portable communications applications. Several linear arrays of DRAs at 20 GHz are investigated in this thesis, with results discussed in Chapter 7. These elements have significant potential for personal communications applications, due to their compact size, wide bandwidth, and high radiation efficiency.

2.4 Summary of Survey

This chapter has reviewed the various architectures proposed for portable communication applications. The architectures were categorized into printed and non-printed arrays. In the printed array class, the multi-layered architecture shows promise, but more work on development is still required to deal with such issues as thermal stress, board alignment, and inter-connections between layers. In the non-printed class, several technologies show potential. A planar array of slotted waveguides has the ability for 2D scan at millimeter waves [32], as does the planar array of printed dipoles on ferrite substrate with dielectric guide feeds [28]. Arrays of dielectric rods also have potential, although more development is required to explore alternate feeding structures to excite the rods. Arrays of DRAs also show considerable promise, and will be discussed further in Chapter 7.

3. Review of Dielectric Antennas and Microwave Ferrites

3.1 Overview

This chapter reviews some of the recent work performed on dielectric resonator antennas (DRAs) and dielectric rod antennas. A summary of microwave ferrite properties relevant to this thesis is also presented in this chapter. Since both the ferrite resonator antenna (Chapter 4) and the ferrite rod antenna (Chapter 6) behave similarly to DRAs and dielectric rods, respectively, the modes of operation and design equations introduced in this chapter will be applicable to their ferrite counterparts with little or no modifications.

The first section of this chapter outlines the basic modes of operation of the DRA, a recently discovered antenna which outperforms microstrip antennas in many aspects. Design equations are presented for predicting the resonant frequency and radiation patterns of DRAs. With a few modifications, these equations can be adopted to predict the resonant frequency of the ferrite resonator antennas presented in Chapter 4.

The second section consists of a review of dielectric rod antennas. Their modes of operation are discussed, and design equations are presented to determine guided wavelength and radiation patterns. These equations will be used to design the ferrite rod antennas discussed in Chapter 6, and to predict their radiation patterns.

Finally, this chapter concludes with a brief review of the performance of microwave ferrites. The tensor nature of the ferrite permeability and the dependence of this permeability on an externally applied magnetic field is discussed. These properties are exploited in this

thesis to produce ferrite antennas with frequency agility, polarization agility, and magnetic phase shift control for beam steering applications.

3.2 Dielectric Resonator Antennas

As discussed in Chapter 2, printed antenna elements, such as microstrip patches or printed dipoles, have copper losses and surface wave losses which are proportional to frequency. Alternative antenna elements are thus required if these losses are to be minimized. Recent investigations have shown that dielectric resonator antennas (DRAs) offer several advantages at microwave and millimeter-wave frequencies [42]. These advantages include small size, large impedance bandwidth, high radiation efficiency, ease of manufacturing, and simple feed-coupling mechanisms (such as probe coupling, aperture coupling, or proximity coupling to microstrip lines). These advantages make dielectric resonator antennas potential candidates for personal communication applications. This section briefly reviews the major characteristics of dielectric resonators, focusing on their potential as radiating elements for personal or satellite communications.

3.2.1 Fundamental Modes of Operation

Historically, dielectric resonators have been used as components for such circuits as oscillators and filters. Their advantage lies in their compact size, ease of fabrication, and simple proximity coupling to microstrip lines. A large body of literature exists describing the behaviour of these resonators and providing analytical and numerical techniques to predict their performance [43,44]. Typically, these resonators have been enclosed in a metal case to prevent radiation.

More recently, however, attention has been given to the radiation properties of open dielectric resonators [45-50]. It has been found that these dielectric resonator antennas (DRAs) possess numerous advantages at microwave and millimeter-wave frequencies, as were mentioned above. In this section, the fundamental modes of the rectangular and cylindrical disk DRAs are presented, along with equations used to predict their resonant frequency. These fundamental modes are used in the investigation of the ferrite resonator antennas and are the modes of interest for personal communication applications.

3.2.1.1 Rectangular DRA

The rectangular DRA in free space is shown in Figure 3.1 along with the field configuration of the lowest order TE_{111}^z mode. The equations for the fields of this mode are shown in Appendix A. In practice, the DRA is mounted on a ground plane; however, the analysis can be treated the same as the DRA in free space by removing the ground plane and doubling the height of the DRA (image theory). The resonant frequency of the DRA can be predicted using a dielectric waveguide model, where the DRA is modelled by the fields of a dielectric waveguide of rectangular cross-section. The waveguide is assumed to have perfect magnetic walls on four sides and is truncated in the propagation direction with two leaky walls [51]. A transcendental equation is derived by solving the fields with the assumed boundary conditions:

$$d = (2/k_z) \tan^{-1} (k_{zo}/k_z) \quad (3.1)$$

where:

$$\begin{aligned} k_{zo}^2 &= (\epsilon_r - 1)(2\pi f_0/c)^2 - k_z^2; \\ k^2 &= k_x^2 + k_y^2 + k_z^2 = \epsilon_r(2\pi f_0/c)^2; \\ k_x &= \pi/a; \text{ and } k_y = \pi/b. \end{aligned}$$

Figure 3.1 TE'_{111} field configuration in a rectangular DRA

The transcendental equation is solved for k_z , then the resonant frequency (f_0) can be determined. If the dimensions (a, b, d) and the permittivity (ϵ_r) of the DRA are known, eqn. (3.1) can be used to predict the resonant frequency. The achievable accuracy will depend on how well the fields are

confined within the DRA. The above equations lead to predicted resonant frequencies of between 5% and 10% of measured values.

3.2.1.2 Cylindrical DRA

The cylindrical DRA is shown in Figure 3.2. Field configurations for the $TE'_{01\delta}$, $TM'_{01\delta}$, and $HE_{11\delta}$ modes are depicted [44]. The dominant mode will depend on the height-to-radius ratio ($2h/r$). The equations for these fields are also given in Appendix A. The mode used for this work is the $HE_{11\delta}$ mode, whose far-field pattern is similar to that of a horizontal magnetic dipole. The resonant frequency for this mode can be determined using the empirically derived formula [42]:

$$k_o = \frac{2\pi f_o r}{c} = \frac{6.324}{\sqrt{\epsilon_r + 2}} \left[0.27 + 0.36 \left(\frac{r}{2h} \right) + 0.002 \left(\frac{r}{2h} \right)^2 \right] \quad (3.2)$$

For known dimensions (r , h) and permittivity (ϵ_r), the resonant frequency (f_o) of the cylindrical DRA can be determined. Again, the accuracy of the above equation is within 5% to 10%.

3.2.2 Antenna Properties of DRAs

The rectangular DRA operating in the TE^x_{111} mode and the circular DRA operating in the $HE_{11\delta}$ mode will both radiate as short horizontal magnetic dipoles as shown in Figure 3.3. The resultant equations for the far-field patterns of an x-directed magnetic dipole are:

H-Plane:

$$ep(\theta, \phi = 0^\circ) = \begin{cases} E_o \cos \theta & 0^\circ \leq \theta \leq 90^\circ \\ 0 & 90^\circ \leq \theta \leq 180^\circ \end{cases} \quad (3.3)$$

E-Plane:

$$ep(\theta, \phi = 90^\circ) = \begin{cases} E_0 & 0^\circ \leq \theta \leq 90^\circ \\ 0 & 90^\circ \leq \theta \leq 180^\circ \end{cases}$$

Figure 3.2 Lower order mode field configurations in a cylindrical DRA

Figure 3.3 Magnetic dipole model for a rectangular DRA in TE^x_{111} mode

This model assumes that the DRA is placed on a ground plane of infinite extent. However, for personal communication applications such as portable briefcase terminals or vehicle mounted antennas, DRAs are typically placed on finite sized ground planes, which have two major effects on the pattern:

- i) spillover of the radiation into the lower hemisphere ($\theta > 90^\circ$); and
- ii) diffraction ripples in the E-plane due to the finite ground plane edges.

These effects will be studied in detail in Chapter 5, using the Geometric Theory of Diffraction.

One of the advantages of DRAs is their potential for very high radiation efficiency since there are no conductor or surface-wave losses. Radiation efficiencies of higher than 98% have been reported [52]. Another advantage of DRAs is the ease of the coupling using various methods including coupling with a coaxial probe, coupling from a microstrip-fed slot aperture, and proximity coupling from a microstrip line [42].

The properties of DRAs discussed above apply equally well to the ferrite resonator antennas (FRAs) introduced in Chapter 4, since an unbiased FRA behaves like a DRA. It will be seen that the equations presented in this chapter apply with little or no modification to FRAs, and are used to design FRAs of circular and rectangular cross-section.

3.3 Dielectric Rod Antennas

3.3.1 Introduction

A second type of dielectric antenna is the dielectric rod antenna. A comprehensive overview of the research on dielectric rod antennas can be found in [33,34]. Dielectric rods, when properly excited, produce single lobe radiation patterns, directed along the rod axis. As is typical of endfire antennas, the directivity of the dielectric rod is proportional to its length.

Dielectric rod antennas are useful for applications where small apertures are required or where metal antennas are undesirable. Other advantages include low losses (no conductor or surface wave losses), compact size, and high gain. Also, the rod can be shaped for optimization of either low sidelobes or for maximum gain [53,54]. The following sections describe the field modes for dielectric rods with rectangular cross-section, and provide equations for determining guided wavelength and far-field patterns.

3.3.2 Guided Wavelength of Dielectric Rods

In order to estimate the radiation patterns of the rods, the guided wavelength for a dielectric guide with rectangular cross-section is required. (Equations for dielectric guides of circular cross-section have also been derived [33,55,56], but this thesis concentrates only on rectangular rods). Several methods are available for determining the guided wavelength of a rectangular dielectric waveguide [57-59]. One of the simplest methods, which is also fairly accurate, is based on the approximations made by Marcatili [60]. Using the Marcatili method, the fields at a given cross section are approximated as shown in Figure 3.4. The equations for these fields are found in Appendix A. The fields within the dielectric are assumed to be sinusoidal, while those outside the guide are assumed to decay exponentially. All fields are assumed to be confined to regions 1 through 5, with no fields in the shaded areas. These approximations are valid for well-guided waves. By solving the boundary conditions, the guided wavelength of the rod can be determined [60]:

Figure 3.4 Field configuration of a rectangular dielectric guide, from [60]

$$\lambda_z = \frac{2\pi}{k_z} \quad (3.4)$$

where:

$$k_z = \sqrt{k_1^2 - k_x^2 - k_y^2}$$

$$k_x = \frac{p\pi}{a} \left(1 + \frac{A_3 + A_5}{\pi a} \right)^{-1}$$

$$k_y = \frac{q\pi}{b} \left(1 + \frac{A_2 + A_4}{\pi \epsilon_r b} \right)^{-1}$$

$$A_i = \frac{\pi}{\sqrt{k_1^2 - k_i^2}}, \quad i = 2, 3, 4, 5$$

$$k_i = \frac{2\pi\sqrt{\epsilon_{ri}}}{\lambda_o}, \quad i = 1, 2, 3, 4, 5$$

and p, q are the mode numbers.

Thus, by knowing the cross-sectional dimensions of the guide (a,b), and the permittivity of the dielectric and surrounding media (ϵ_{ri}), the guided wavelength (λ_z) can be determined for the desired mode (p,q). These equations will be used to determine the guided wavelength of the ferrite rods in Chapter 6. Figure 3.5 shows the dispersion characteristics for a dielectric rod with square cross-section ($a = b$) for various modes, E_{pq}^y , using eqn. (3.4). This figure plots the normalized guided wavelength, K, (normalized with respect to ϵ_r and λ_o) versus the normalized height, B, and thus can be used for any dielectric guide with square cross-section ($a = b$). It should be noted that contrary to Figure 3.5, the E_{11}^y mode actually has no cut-off, which is not predicted by eqn. (3.4). Thus the Marcattili approximation is not valid near cut-off.

Figure 3.5 Dispersion characteristics for a square dielectric rod

3.3.3 Radiation Patterns

The dielectric rod antenna can be considered as a truncated dielectric waveguide. Radiation from the rod occurs from the ends of the rod and from any discontinuity along the rod. To minimize unwanted reflections at the dielectric-air interface, the ends of the rod are usually tapered [34].

To determine the radiation patterns of the rods, two analytical methods are presented. The first method is based on Huygen's principle and treats the dielectric rod as an array of elementary dipoles [33]. The

resultant radiation patterns are approximated by the following equations for a rod parallel to the z-axis [55].

$$P(\theta) = ep(\theta) \cdot sp(\theta) \quad (3.5)$$

where:

$$ep(\theta) = \begin{cases} \cos(\theta) & \dots\dots\dots E - Plane \\ 1 & \dots\dots\dots H - Plane \end{cases}$$

and:

$$sp(\theta) = \frac{\sin\left[\left(\frac{\pi L}{\lambda_g}\right) - \frac{\pi L \cos \theta}{\lambda_o}\right]}{\pi L \left(\frac{1}{\lambda_g} - \frac{\cos \theta}{\lambda_o}\right)}$$

The second method is known as Schelkunoff's method, and involves calculating the radiation pattern from equivalent magnetic and electric surface currents derived from the tangential components at the surface of the dielectric rod [33]. This method is best suited for rods of circular cross-section operating in the TE_{01} mode, where the fields are known exactly. Approximations are made for rectangular guides operating in the hybrid HE_{11} mode. The radiation pattern derived from these surface currents is presented below for a rod parallel to the z-axis:

$$P(\theta, \phi) = \cos\left[\frac{k_o a}{2} \sin \theta \cos \phi\right] I_1 I_2 \quad (3.6)$$

where:

$$I_1 = \frac{2b \left(\frac{\pi}{2}\right)^2 \cos\left[\frac{k_o b}{2} \sin \theta \sin \phi\right]}{\pi \left(\frac{\pi}{2}\right)^2 - \left[\frac{k_o b}{2} \sin \theta \sin \phi\right]^2}$$

$$I_2 = \frac{1}{2k_o}(A - B)$$

$$A = \frac{1}{K - \cos \theta} [1 - \cos \{(K - \cos \theta)k_o L\}] + \frac{1}{K + \cos \theta} [(K + \cos \theta)k_o L]$$

$$B = \frac{\sin \{(K - \cos \theta)k_o L\}}{K - \cos \theta} - \frac{\sin \{(K + \cos \theta)k_o L\}}{K + \cos \theta}$$

$$K = k_g/k_o = \lambda_o/\lambda_g$$

Sample patterns at 20.5 GHz using Huygen's method and Schelkunoff's method are shown in Figure 3.6 for a dielectric rod with a permittivity of $\epsilon_r = 14.8$, a length of 30 mm, and a cross-section of 2.28 mm by 2.28 mm. These equations are later compared with measured patterns of the ferrite rod presented in Chapter 6.

Figure 3.6 Theoretical radiation patterns of a dielectric rod

3.4 Microwave Ferrites

3.4.1 Introduction

Ferrite materials have long been used for various microwave components such as isolators, phase shifters, and switches. Good treatments of the behaviour of ferrites at microwave frequencies can be found in [61-70]. This section provides a brief overview of microwave ferrites, concentrating on the relevant properties of ferrites pertaining to the research carried out in this thesis.

3.4.2 Physical Properties

Ferrites are ionic compounds whose properties are determined by the crystalline structure of the compounds. The most common ferrites for use at microwave frequencies are garnets, and metal oxides of the spinel ($M^{2+}Fe^{3+}O_4^{2-}$) and perovskite ($M^{3+}Fe^{3+}O_3^{2-}$) types (where M refers to a metal ion) [62]. Ferrites typically have a high resistivity (about 10^6 times that of metals) and moderately high dielectric constant at microwave frequencies ($10 \leq \epsilon_r \leq 20$). Their physical properties include hardness, brittleness, opaqueness, and a high melting point. Ideally, ferrites are non-conductors, but due to manufacturing problems, ferrites may be semi-conductors. The physical temperature of the ferrite is also of importance, for the saturation magnetization of the ferrite is strongly dependent on temperature as shown in Figure 3.7. As the temperature rises, the magnetization of the ferrite decreases. The temperature at which the magnetization drops to zero is known as the Curie temperature (T_c). Above this temperature, the average kinetic energy of the molecules in the ferrite is enough to overcome the attractive bonds of the magnetic dipoles. This causes a random orientation of the dipoles, resulting in no net

magnetization. Typical Curie temperatures for common ferrite materials range from 85° to 600°C. Thus, for ground-based applications, one usually need not worry about loss of magnetization due to temperature variations.

3.4.3 Low-Frequency Behaviour

At d-c and low frequencies, (for temperatures below the Curie temperature), ferrites exhibit a very high magnetization, and hence high permeability ($\mu_r \geq 1000$), due to the formation of domains (relatively large areas in the crystal in which all the magnetic dipoles are aligned). These domains are initially randomly oriented in the crystal, resulting in a net cancellation of the magnetization. When a d-c magnetic field is applied, a net magnetization results, as the domains are magnetized in the direction of the field. Permeabilities of several thousands are common for ferrites at low frequencies.

Figure 3.7 Dependence of ferrite magnetization on temperature

3.4.4 Microwave-Frequency Behaviour: The Gyroscopic Model

At microwave frequencies and higher, the relative permeability of ferrites reduces nearly to unity. At these frequencies, the magnetization vectors of the domains are unable to respond quickly enough to the magnetic field. However, the individual electrons (magnetic dipoles) are able to respond to the applied magnetic field at these frequencies. Thus the behaviour of the overall magnetization vector is the net effect of all the electrons acting as separate individuals rather than each electron being influenced by its neighbour (domain behaviour), as is the case at lower frequencies. The study of the microwave behaviour of ferrites, therefore, focuses on the behaviour of a single electron in the presence of the applied fields. In this approach the electron is treated as a charged sphere with angular momentum (due to its mass) and a magnetic moment (due to its charge and spin). The spinning electron is modelled as a gyroscope as shown in Figure 3.8. The tensor nature of the permeability of the ferrite can be derived from the equations of motion describing the behaviour of the electron gyroscope in a uniformly saturated ferrite, due to the presence of a small signal r-f magnetic field [71]. In the absence of the r-f field, the electron will be aligned with the direction of the static bias field (magnetic moment parallel to the magnetic field). When a small-signal r-f field is applied in a direction perpendicular to the bias field, the electron will precess about the axis of the static field, with a precession frequency dependent on the magnitude of the static bias field. For a uniformly magnetized and saturated region of a lossless ferrite subjected to a small-signal r-f field ($\mathbf{h}$), the flux density ($\mathbf{b}$) is given by:

$$\mathbf{b} = [\mu] \mathbf{h} \quad (3.7)$$

or, decomposing into rectangular co-ordinates:

$$b_x = \mu h_x - j\kappa h_y$$

$$b_y = \mu h_y + j\kappa h_x$$

$$b_z = \mu_o h_z$$

where:

$$[\mu] = \begin{bmatrix} \mu & -j\kappa & 0 \\ j\kappa & \mu & 0 \\ 0 & 0 & \mu_o \end{bmatrix}; \quad (3.8)$$

$$\mu = \mu_o - \frac{\gamma M \omega_o}{\omega_o^2 - \omega^2}; \quad \kappa = \frac{\gamma M \omega}{\omega_o^2 - \omega^2}; \quad \text{and } \omega_o = \gamma H_o.$$

From the above equations, it is evident that $[\mu]$ is a function of frequency (ω) and applied magnetic field bias (H_o).

Figure 3.8 Gyrosopic model of an electron in an applied magnetic field

For circular polarized waves (where $\mathbf{h}_x = \pm j\mathbf{h}_y$), it can be shown by manipulating the above equations, that the tensor permeability reduces to a scalar:

$$\mathbf{b}_{\pm} = \mu_{\pm} \mathbf{h}_{\pm} \quad (3.9)$$

where:

$$\mu_{+} = \mu - \kappa, \text{ and } \mu_{-} = \mu + \kappa.$$

The $\pm$ in eqn. (3.9) represents right hand circular polarization (rhcp) and left hand circular polarization (lhcp), respectively. This result leads to interesting phenomena. Since the wave propagation constant is a function of permeability ($\beta_{\pm} = \omega\sqrt{\epsilon\mu_{\pm}}$), a linearly-polarized wave (which can be thought of as a superposition of rhcp and lhcp) travelling through a length L of a uniformly saturated ferrite, will undergo a rotation in polarization over the length (Faraday Rotation) equal to $0.5L(\beta_{-} - \beta_{+})$ and a phase shift of $L(\beta_{-} - \beta_{+})$. This phase shift property, which is dependent on the applied magnetic bias, will be explored in Chapters 6 and 7, for the ferrite rod antenna.

3.4.5 Losses

3.4.5.1 Introduction

The losses in a ferrite material can be broken down into four components: dielectric losses, gyromagnetic resonance losses, low-field losses, and spin-wave losses. Not all loss mechanisms may be present, depending on the frequency of operation and power levels used. This section briefly overviews the various losses, and describes how some of them can be avoided.

3.4.5.2 Dielectric Losses

Ferrite materials, when unbiased, behave as dielectric materials and are thus subjected to dielectric losses. This can be accounted for by using the complex permittivity:

$$\epsilon = \epsilon' - j\epsilon'' \quad (3.10)$$

The imaginary component of the permittivity (ϵ'') is used to represent the dielectric losses. These losses will set a limit to the maximum average power the ferrite can handle. In practice, the dielectric losses are expressed as a loss tangent:

$$\tan\delta = \epsilon''/\epsilon' \quad (3.11)$$

Typical values of loss tangent for good microwave ferrites are smaller than 0.0005.

3.4.5.3 Gyromagnetic Resonance Losses

In the earlier treatment of ferrite permeability, it was assumed that no losses were present in the ferrite. To deal with such losses, a complex permeability is introduced:

$$\mu = \mu' - j\mu'' \quad (3.12)$$

The imaginary component of the permeability (μ'') is used to represent permeability-related losses in the ferrite. High losses occur around the gyromagnetic resonance of the ferrite (when $\omega = \omega_0$), which is a function of the applied magnetic field H_0 . Figure 3.9 plots a typical response of the losses (μ'') in an infinite ferrite slab as a function of frequency, for a given ferrite material. The $\mu''_{\pm}$ refers to the losses associated with a circular-polarized wave travelling in the forward (+) and reverse (-)

directions. As seen in Figure 3.9, waves travelling in the forward direction will be absorbed because of the high resonance peak of μ_+ , while a wave in the opposite direction will pass with little attenuation. This is the non-reciprocal behaviour of ferrites and is exploited in such devices as resonant isolators. The losses peak around the gyromagnetic resonance, and the 3 dB width of the peak is known as the line width (ΔH). The linewidth of the ferrite is believed to be a function of several material properties including crystalline structure, porosity, and surface roughness [66].

Figure 3.9 Typical dependence of permeability loss on magnetic bias

To avoid excessive losses, the ferrite should be used at a frequency band several line widths away from resonance. When the ferrite material is not infinite in extent, the peak losses may not occur at the gyromagnetic resonance but at a frequency influenced by the shape of the ferrite.

3.4.5.4 Low Field Losses

Resonance absorption losses, called low field losses, can occur in an unsaturated ferrite. In an unsaturated ferrite, the domain structure of the material can contribute to the effective demagnetization field. This loss can occur in an unsaturated ferrite below a maximum frequency [66]. Thus to avoid low-field losses, if the ferrite is not saturated, it should be operated well above the resonant frequency, where domain behaviour is non-existent.

3.4.5.5 High Power Losses

If the peak microwave power in the ferrites exceeds a threshold value, the insertion loss will rise dramatically. This non-linear behaviour is caused by the generation of spin waves when the microwave magnetic field exceeds a critical value [72-74]. Spin waves arise from the presence of periodic variations of the uniform precession through the material [66]. These variations may be in the form of standing waves or travelling waves, and provide an alternative route for the energy absorbed by the magnetic dipoles to be transferred to the lattice (resulting in high losses). The critical field strength (h_{crit}) where the onset of spin waves occurs, is related to the spin-wave linewidth (ΔH_k) by:

$$h_{crit} = C \cdot (\omega/\omega_m) \cdot (\Delta H_k) \quad (3.13)$$

where:

C is a constant dependent upon the shape factor ($C \approx 1$); and
 $\omega_m = \gamma 4\pi M_s$.

As seen in eqn. (3.13) for a given frequency, the critical field strength varies proportionally with spinwave linewidth (ΔH_k) and inversely with

saturation magnetization (ω_m). Typical values of ΔH_k range from 1-20 Oe. [66,68]. Another consequence of eqn. (3.13) is that the power handling of the ferrite can be increased by increasing the impedance level of the ferrite device. Typical peak power values for commonly used ferrites are of the order of 1 kW. The effects on the losses in a ferrite with increased power levels can be seen in Figure 3.10. Three curves are plotted corresponding to various power levels. At low power ($h/h_{crit} \ll 1$), the losses are due to the gyromagnetic resonance, and the linewidth is relatively narrow. As the power is increased towards the critical level ($h/h_{crit} \approx 1$), the resonance linewidth increases, and a secondary resonance is seen to develop. At very high power levels ($h/h_{crit} \gg 1$), the secondary resonance is seen to increase, while the resonance linewidth also increases significantly. Thus, for low loss applications, it is best to limit the peak power well below the h_{crit} level, to avoid significant broadening of the linewidth.

3.4.6 Hysteresis

Hysteresis in ferrites is the phenomenon in which the ferrite material remains magnetized after the applied magnetic field is removed. For the case of a small magnetic field bias, the domain wall motion is reversible, and there is no net magnetization when the bias magnetic field is removed. When a larger magnetic field is applied, however, larger domains form, which are not reversible; thus when the magnetic field is removed, there is a remnant magnetization (B_r) in the ferrite material. This phenomenon is represented graphically in the form of a hysteresis loop, as shown in Figure 3.11.

Figure 3.10 Typical high power losses in ferrites

Two important parameters of the hysteresis loops are the ratio, R , of remnant magnetization (B_r) to saturation magnetization ($4\pi M_s$), and the coercive field strength (H_c). Since losses are proportional to the area of the loop [64, 65], it is best to have a ferrite with a small coercive field strength. Also, for latching applications, the remnant ratio should be as high as possible. Latching of the ferrite has the advantage that the magnetization can be maintained at the desired value, without the need to maintain the bias field, thus minimizing the need for continuous bias power [75].

Figure 3.11 Typical hysteresis loop of a ferrite

3.4.7 Published Ferrite Specifications

Two ferrite materials were used in this thesis: calcium vanadium (TTVG-1850) and nickel ferrite (TT72-1300) from Trans-Tech, Inc. These ferrites were chosen in part due to their availability as well as exhibiting good material properties in terms of their relatively low loss tangents and high saturation magnetizations. The relevant published data for these materials are shown below in Table 3.1. The dependence of the permeability on applied bias and the tensor nature of the permeability is exploited in this research on FRAs and ferrite rod antennas. The phenomena described above will be used to investigate such properties as frequency agility, polarization agility, and electronic phase shift control for beam steering applications.

Material	TTVG-1850	TT72-1300
$4\pi M_s$ (G)	1850	1300
ΔH (Oe)	≤ 15	< 150
ϵ_r	14.8	17.6
$\tan\delta$	≤ 0.0002	≤ 0.0005
T_c ($^{\circ}\text{C}$)	214	210
ΔH_k (Oe)	2.0	2.1
B_r (G)	1232	898
H_c (Oe)	0.5	0.69
μ_r (@1 kHz)	388	---

Table 3.1 Published data for ferrite materials used in this thesis.

4. Ferrite Resonator Antennas

4.1 Overview

This chapter introduces the novel ferrite resonator antenna (FRA), the first of two classes of ferrite antennas studied in this thesis. FRAs possess many of the attributes of dielectric resonator antennas (DRAs), and the properties of DRAs presented in Chapter 3 all apply to FRAs. In addition to having many of the DRA attributes, some properties of the FRA can be altered by the application of a d-c magnetic field bias. Two important properties which stem directly from the tensor permeability of ferrites, and which are demonstrated in this chapter, are frequency tuning and polarization agility. Frequency tuning is explored using an FRA of rectangular geometry while polarization agility is studied using a cylindrical disk FRA. Design equations are provided for predicting the resonant frequency of the unbiased FRAs and for predicting the maximum amount of resonant frequency shift achievable. To determine the losses present in FRAs, the radiation efficiency is measured using Wheeler's Cap Method [81,82].

4.2 Frequency Tuning Property of FRAs

4.2.1 Antenna Geometry

To demonstrate the frequency tuning property of the FRA, a ferrite of rectangular cross-section was chosen. The ferrite material used was calcium vanadium (see TTVG-1850 in Table 3.1) which was machined into a rectangular cross-section of dimensions $a = 10.0$ mm, $b = 2.5$ mm, and $d = 5.0$ mm as shown in Figure 4.1. (These dimensions were chosen based on available sample sizes.) The FRA was excited with a microstrip-fed slot aperture [51] and was operated in the TE_{111}^y mode. Proper coupling to this

Figure 4.1 Geometry of aperture-fed rectangular FRA with magnetic bias

mode was achieved by placing the resonator close to one of the ends of the slot aperture. Probe coupling can also be used to excite this mode; however, the connector would have interfered with the placement of the magnet, thus slot coupling was chosen.

In the TE_{111}^y mode, the radiation pattern of the FRA is similar to a horizontal magnetic dipole aligned parallel to the y axis (similar to Figure 3.3). Biasing of the ferrite with a static magnetic field was achieved using a permanent magnet. The magnet was located beneath the ground plane, as shown in Figure 4.1. Two biasing directions were investigated, with the orientation of the magnet parallel (x-direction) and perpendicular (y-direction) to the length of the ferrite. The physical presence of the magnet caused negligible influence on the ferrite antenna performance, due to its location underneath the ground plane. For the chosen resonator parameters, the resonant frequency of the antenna was calculated to be 10.1 GHz using eqn. (3.1).

4.2.2 Frequency Tuning

4.2.2.1 Theory

The rectangular FRA was used to investigate the potential for tuning the resonant frequency of the antenna by applying a d-c magnetic field to the FRA. The potential for a shift in resonant frequency with magnetic bias can be seen theoretically, by modifying eqn. (3.1) to include the effective permeability, μ_{eff} , of the FRA material. The modified equations are shown below:

$$d = (2/k_z) \tan^{-1} (k_{zo}/k_z) \quad (4.1)$$

where:

$$\begin{aligned} k_{zo}^2 &= \mu_{\text{eff}}(\epsilon_r - 1)(2\pi f_0/c)^2 - k_z^2; \\ k^2 &= k_x^2 + k_y^2 + k_z^2 = \mu_{\text{eff}} \epsilon_r (2\pi f_0/c)^2; \\ k_x &= \pi/a; \quad k_y = \pi/b; \text{ and } \mu_{\text{eff}} = \mu \pm \kappa. \end{aligned}$$

The $\pm$ for μ_{eff} depends on the orientation of the applied bias magnetic field. For a saturated ferrite material of infinite extent:

$$\mu=1, \text{ and } \kappa \approx \frac{(4\pi M_r)}{(4\pi M_s)} \frac{\gamma(4\pi M_s)}{f} \quad (4.2)$$

By including μ_{eff} in the resonant frequency equation, the maximum shift in frequency due to an applied bias field can be predicted. Predicting the shift in resonant frequency as a function of applied bias strength is much more complex, since the required internal magnetic field within the FRA is a complex function of ferrite shape, and is difficult to determine. Also, the exact equations for μ and κ for an unsaturated ferrite are unknown. Nevertheless, the above equations are useful in predicting the maximum achievable shifts in resonant frequency, which occur when the ferrite is saturated. Equations (4.1) and (4.2) were used to predict the maximum shifts of the resonant frequency for the FRA of Figure 4.1. The results are shown in Table 4.1, and are compared with measured results in the following section.

Magnetic Bias	Resonant Frequency (GHz)
No Bias	10.1
Y-Directed Bias	9.0
X-Directed Bias	12.2

Table 4.1 Predicted shifts in the resonant frequency of the FRA due to magnetic bias

4.2.2.2 Frequency Shift Measurements

The resonant frequency of the ferrite antenna was measured as a function of bias strength for magnetization parallel to the x and y directions [76.77]. A magnetic field in the z direction was not applied, since the presence of the magnet above the FRA would have disturbed both the resonant frequency and the radiation patterns. The return loss of the unbiased FRA is shown in Figure 4.2. The unbiased resonant frequency was 10.5 GHz, 4% higher than predicted using the dielectric waveguide model approximation (eqn. 3.1). The applied magnetic field intensity was varied by controlling the spacing, s , of the magnet from the ferrite antenna. (For practical applications, an electromagnet could be used.)

Figure 4.2 Return loss of the unbiased rectangular FRA

The field intensity in air was measured using a gaussmeter, whose probe was placed at the position where the centre of the ferrite antenna was to be located. The resultant shift in resonant frequency is shown in Figure 4.3. The upper curve corresponds to the frequency shift due to the x-directed applied bias field while the lower curve is due to the y-directed bias field. A total frequency shift of 16% was achieved. The 10 dB impedance bandwidth was 5.75% for the unbiased ferrite. Table 4.2 lists the impedance bandwidth when maximum bias was applied. The bandwidth is seen to decrease to a minimum of 5% when maximum magnetic bias strength was applied. The return loss at the resonant frequency consistently remained better than 15 dB as the magnetic biasing was applied. Comparison between the theoretical and measured maximum shifts in resonant frequencies of the FRA shows fairly close agreement. Discrepancies may be attributed to the finite size and shape of the ferrite (which introduce an unknown demagnetization factor), non-uniform magnetization within the ferrite, and the original assumptions of field confinement made by the dielectric waveguide model equations (3.1).

Magnetic Bias	f_o (measured)	f_o (calculated)	Bandwidth (measured)
none	10.5	10.1	5.75%
y-directed bias	9.35	9.0	5.0%
x-directed bias	11.45	12.2	5.3%

Table 4.2 10 dB impedance bandwidth of the FRA with max. applied bias

Changing the polarity of the bias field did not alter the direction of the change in resonant frequency. Only a change in direction (x or y) caused a change in the direction of the frequency shift.

Figure 4.3 Resonant frequency vs. magnetic bias of the rectangular FRA

4.2.2.3 Pattern Measurements

The far-field radiation patterns of the FRA were measured for the unbiased and biased cases. The antenna was mounted on a 12" diameter circular ground plane and measurements were carried out in an anechoic chamber with pattern cuts in both the E-plane and H-plane over a frequency range from 8-12 GHz. Figure 4.4 shows the measured H-plane and E-plane patterns of the unbiased ferrite antenna, at 10 GHz. As previously mentioned, the antenna has a radiation pattern similar to that of a horizontal magnetic dipole. The ripples in the E-plane pattern are caused

Figure 4.4 Measured radiation of the unbiased rectangular FRA at 10 GHz

by diffraction of the fields from the edges of the finite ground plane and from the coaxial-to-microstrip launcher. The cross-polarized levels, for the unbiased FRA were, on average, 15 dB below the co-polarized levels. Figure 4.5 shows the measured patterns of the y-biased FRA at 9.4 GHz while Figure 4.6 shows patterns of the x-biased FRA at 10.8 GHz.

Biasing in the x-direction resulted in an interesting observation. The co-polarized levels were similar to those of the unbiased case while the cross-polarized levels increased significantly to an average of 10 dB below the co-polarized peak. This result is not unexpected since the permeability of the ferrite is a tensor with non-zero off-diagonal components [62]. This means that a linearly polarized magnetic field can generate a component of magnetization in an orthogonal direction (e.g. an x-directed field can produce magnetization in the y direction). This effect suggests the possibility of obtaining polarization-agile antennas which will be investigated in the following section.

4.2.2.4 Discussion

This section demonstrated the capability of magnetic tuning of the resonant frequency of an FRA of rectangular cross-section. The maximum obtainable range of frequency tuning will depend on the saturation magnetization and the physical dimensions of the FRA. The higher the saturation magnetization of the ferrite material, the larger the frequency tuning range should be. Also, for a given saturation magnetization, as the frequency is increased, μ_{eff} will decrease (since κ in eqn. (4.2) is inversely proportional to frequency), therefore, a smaller amount of frequency tuning is possible.

Figure 4.5 Measured radiation of the y-biased rectangular FRA at 9.4 GHz

Figure 4.6 Measured radiation of the x-biased rectangular FRA at 10.8 GHz

The resonant frequency chosen for the above experiments was determined, in part, by the saturation magnetization of the ferrite material available. In order for the FRA to be used at higher frequencies, ferrite materials with higher saturation magnetizations will be required to produce a similar frequency tuning range. Nickel ferrites with saturation magnetizations of up to 5000 G are now commercially available which, if linear scaling is assumed (based on eqn. (4.2)), should allow for better than $\pm 10\%$ tuning of the resonant frequency at 20 GHz, which is in the frequency band of interest for portable satellite communications.

4.3 Polarization Agility of FRAs

4.3.1 Antenna Geometry

To demonstrate the ability of altering the polarization with the applied magnetic field, a cylindrical disk FRA was used, similar to the DRA shown in Figure 3.2. With circular symmetry, the $HE_{11\delta}$ mode is degenerate in the x-y plane. This means the orientation of the polarization is solely a function of the slot orientation. The FRA was made of a nickel ferrite material (see TT72-1300 in Table 3.1) which was machined into a disk of circular cross-section with a diameter of 16.0 mm and a height of 3.0 mm as shown in Figure 4.10 [78]. The cylindrical disk FRA was excited using microstrip slot coupling and was operated in the $HE_{11\delta}$ mode. When unbiased, the permeability of the ferrite is scalar, thus the FRA behaves as a cylindrical disk dielectric resonator antenna. The $HE_{11\delta}$ modes of the dielectric disk resonator are degenerate and mutually orthogonal. The two degenerate $HE_{11\delta}$ modes radiate like x- and y-directed magnetic dipoles [42]. However, in the feeding scheme shown in Figure 4.7, the slot excites only the mode which radiates like a y-directed magnetic dipole.

Figure 4.7 Geometry of the cylindrical FRA

4.3.2 Return Loss Measurements

The return loss of the unbiased FRA is shown in Figure 4.8. The resonant frequency was measured to be 7.3 GHz. For the FRA disk parameters given in Figure 4.7, the resonant frequency was calculated to be 6.7 GHz using the empirically derived formula in eqn. (3.2): a discrepancy of about 8%. A 1.2 kG permanent magnet placed parallel to the x-axis was used to bias the FRA, as shown in Figure 4.7. When the FRA is biased, the resonant frequencies of the degenerate $HE_{11\delta}$ modes split. The resonant frequency of the mode with magnetic field parallel to the y-direction decreases while the resonant frequency of the mode with magnetic field parallel to the x-direction increases similarly to the behaviour of the

Figure 4.8 Return loss of the cylindrical FRA without/with magnetic bias

rectangular FRA in the previous section [76,77]. Figure 4.8 shows the return loss response of the biased FRA in which the two split resonant frequencies are clearly observed. Both orthogonal modes are excited because of the tensor nature of the permeability of a biased ferrite, which results in cross-coupling between the two modes [62]. It is known that if an antenna is operated at a frequency in between the resonant frequencies of two nearly degenerate, mutually orthogonal modes, circular polarized radiation can be obtained. This principle has been used in the past to design circular polarized microstrip antennas with a single feed [79]. The circular polarized operation of the FRA disk antenna is based on the same principle.

4.3.3 Pattern Measurements

The far-field radiation patterns of the ferrite antenna were measured for the unbiased and biased cases. Again, the antenna was mounted on a 12" diameter ground plane and measured inside an anechoic chamber. Figure 4.9 shows the normalized co-polarized and cross-polarized H-plane patterns of the unbiased FRA at 7.2 GHz. The cross-polarization level is about 10 dB below the co-polarized level at boresight. Figure 4.10 shows the radiation pattern of the biased FRA. A spinning source antenna was used to measure the axial ratio. A 3 dB axial ratio was measured at boresight at 7.2 GHz. The 3 dB axial ratio at boresight is maintained over a 0.1 GHz bandwidth (7.15-7.25 GHz). Outside this range, the axial ratio degrades and eventually becomes linearly polarized.

Figure 4.9 Measured H-plane pattern of the unbiased cylindrical FRA at 7.2 GHz

Figure 4.10 Measured H-plane pattern of the cylindrical FRA at 7.2 GHz

4.3.4 Discussion

A novel ferrite resonator antenna has been found to produce a circular polarized pattern when an appropriate magnetic field is applied. A minimum measured axial ratio of 3 dB was demonstrated without any attempt at optimization, and would be an acceptable level for several polarization diversity applications. The return loss of the FRA was very sensitive to the position and orientation of the magnet, and the optimal position was found experimentally to obtain the two offset resonant frequencies required to produce circular polarization. Minimization of the axial ratio may be performed by optimizing such parameters as the FRA aspect ratio, material properties, magnetic field strength, and the size of the finite ground plane on which the FRA is mounted. This antenna may prove extremely useful in personal communication and mobile satellite communication applications requiring polarization agility, since it offers the novel capability of magnetically switching between linear and circular polarization.

4.4 Radiation Efficiency of FRAs

4.4.1 Introduction

In personal communication applications, it is important to determine or estimate the losses of the antenna to meet required G/T specifications. There are several methods for determining these losses [80,81]. For small antennas, the radiation efficiency (and hence the losses) of an antenna can be determined using a method proposed by Wheeler [82]. Wheeler's method is used in this section, with some modifications, to determine the radiation efficiency of FRA without and with magnetic bias.

4.4.2 Theory

The losses in the ferrite resonator antenna can be determined by measuring the radiation efficiency. The radiation efficiency of an antenna can be defined as:

$$\eta = \frac{P_{rad}}{P_{rad} + P_{loss}} = \frac{R_{rad}}{R_{rad} + R_{loss}} \quad (4.3)$$

The real part of the input impedance at the input terminals of the antenna can be decomposed into:

$$R_{in} = R_{rad} + R_{loss} \quad (4.4)$$

The input resistance (R_{in}) can be measured with a network analyzer. Wheeler [82] proposed that by enclosing the antenna with a grounded metallic cap, the antenna would cease to radiate, which is equivalent to short-circuiting the radiation resistance ($R_{rad} = 0$). A measurement of the input resistance with such a cap covering the antenna would then be a measure of R_{loss} , and the radiation efficiency could be determined using eqn. (4.3). An attempt was made to determine the radiation efficiency by measuring the input impedance of the FRA with and without a metal cap. The results were quite sensitive to the input reference plane chosen (since the feed launcher was located several wavelengths from the slot aperture) and to the position of the cap. Thus, to measure the radiation efficiency more accurately, the procedure described in [52] was adopted. Instead of measuring the input impedance, the quality factor, Q , of the antenna was measured with and without the Wheeler cap.

The Q of the antenna can be defined as:

$$Q_{ant} = \frac{\omega E}{P_{rad} + P_{loss}} \quad (4.5)$$

The Q representing the losses in the antenna can be defined as:

$$Q_{loss} = \frac{\omega E}{P_{loss}} \quad (4.6)$$

Taking the ratio of Q_{ant}/Q_{loss} and substituting into eqn. (4.3), the radiation efficiency can be expressed as:

$$\eta = 1 - \frac{Q_{ant}}{Q_{loss}} \quad (4.7)$$

4.4.3 Measurements

The experimental setup used to measure the radiation efficiency is shown in Figure 4.11. The Wheeler's cap used to short circuit the radiation resistance was a brass box (4 cm x 4 cm x 4 cm) open on one face, which was placed over the FRA. The Q_{ant} (without the cap) and the Q_{loss} (with the cap) were determined by measuring the 10 dB frequency bandwidth of the return loss. The radiation efficiency was then determined using eqn. (4.7). Figure 4.12 shows the measured return loss of the unbiased FRA without and with the Wheeler's cap. The radiation efficiency was measured for the unbiased FRA, as well as for x-directed and y-directed magnetic biases of 1.2 kG field strengths. The results are shown in Table 4.3 [77].

Magnetic Bias	Δf_{rad} (GHz)	Δf_{cap} (GHz)	Radiation Efficiency
None	0.78	0.018	98%
Y-Directed Bias	0.828	0.036	96%
X-Directed Bias	0.75	0.050	93%

Table 4.3 Radiation efficiency of the FRA

Figure 4.11 Radiation efficiency measurements of the FRA

4.4.4 Discussion

In order to ensure that the radiation resistance is short-circuited, care must be taken to ensure good contact of the Wheeler cap with the ground plane, otherwise, the measured efficiency will be lower than the actual efficiency. Also, detuning effects were noted when the walls of the cap were brought into close proximity with the FRA. The slight decrease in efficiency with applied magnetic field may be attributed to gyromagnetic resonance losses.

4.5. Summary

This chapter introduced novel ferrite resonant antennas and demonstrated their capabilities. The effects of magnetic bias on the resonant frequency, radiation efficiency, and radiation patterns of a ferrite resonant antenna were investigated. FRAs of rectangular and circular cross-section were used to study these effects. A prototype rectangular

Figure 4.12 Unbiased FRA return loss with/without Wheeler's cap

FRA demonstrated a shift in frequency of $\pm 8\%$ of the unbiased resonant frequency, depending on the direction of the magnetic bias. The frequency tuning range should be significantly increased by using ferrite materials with higher saturation magnetization. The radiation patterns of the biased antenna were similar to those of the unbiased antenna. Increases in the cross-polarized levels for the x-directed bias suggested the possibility of magnetic control of the polarization.

A cylindrical disk FRA was used to demonstrate polarization agility with applied magnetic bias. An unbiased disk FRA which produced a linearly-polarized pattern was shown to produce a circular-polarized pattern when a d-c magnetic field was applied. The polarization could thus

be electronically switched from linear to circular, making this antenna a potential candidate for polarization diversity applications.

Finally, the efficiency of the FRA antennas was measured using a modified Wheeler's cap method. The radiation efficiency dropped slightly with applied bias, with a minimum value of 93%. It is suspected that this drop in efficiency is due to the increase in the gyromagnetic resonance frequency with applied bias.

In addition to having fairly wide impedance bandwidth (about 5%) and high radiation efficiency, surpassing printed antennas, FRAs are active antennas which have demonstrated the capability of resonant frequency tuning or polarization agility, making them attractive candidates for personal communication applications.

5. Finite Ground Plane Effects

5.1 Introduction

This chapter demonstrates the use of a simple model for the dielectric resonator antenna combined with the Geometric Theory of Diffraction (GTD) to predict radiation patterns of dielectric resonator antennas on finite ground planes.

Many practical applications involve the placement of antennas on finite ground planes. Some prominent examples are brief-case terminals for portable satellite communications, and vehicular-mounted antennas for mobile satellite or cellular communications. The placement of antennas on finite ground planes usually has significant effects on the radiation patterns, due to interference of the diffracted fields emanating from the edges of the ground planes. These effects are evident in some of the measured patterns presented in Chapter 4. As previously mentioned, the major effects of finite ground planes are to introduce ripples in the ideal pattern and to generate back lobes due to radiation spilling over the finite edges. It is important to be able to predict the magnitude of these ripples and the amount of power spilling into back lobes, since these effects will influence the gain of the antenna.

Several methods are available to predict finite ground plane effects. The two most common are the Method of Moments (MoM) and the Geometric Theory of Diffraction (GTD). MoM is a numerical technique which is useful when dealing with arbitrary shapes. However, due to the large amounts of memory storage required, MoM is best suited for small geometries on the order of a wavelength in size. For large geometries, GTD is better suited.

To date, most models used to predict the far-field radiation patterns of DRAs were based on simple magnetic or electric dipoles for the lower order modes. These models are adequate when dealing with DRAs in free space, or when they are placed on an infinite ground plane. For most practical applications, including personal communications and portable satellite communications, however, DRAs are placed on ground planes of finite dimensions. For these situations, the simple models cannot predict the effects of these ground planes on the DRA radiation patterns. This chapter investigates the use of GTD in combination with these simple models to predict the finite ground plane effects.

The Geometric Theory of Diffraction has been successfully used to predict the radiation patterns of various antennas including apertures, monopoles, microstrip patches, and microstrip disks, mounted on finite ground planes [83-86]. GTD can be used to solve complicated problems with no exact solution, yielding fairly accurate results. This method offers the advantage of being easy to use, and computationally fast. It is an attractive technique for analyzing larger structures, where numerical methods are limited due to memory storage and computation time. GTD has also been successfully combined with other methods such as MoM [87,88].

The next section provides a brief overview of GTD and explains how it was used to model the DRA patterns. Comparisons between measured and predicted radiation patterns are provided along with a study on the effect of varying the ground plane size on the DRA patterns.

5.2 Theory

The Geometric Theory of Diffraction was first introduced by Keller to compensate for the inability of Geometric Optics (GO) to accurately predict

the behaviour of light in the vicinity of finite boundaries [89]. GO is a ray-tracing technique which accounts for direct, reflected, and refracted rays. However, this technique could not predict the behaviour of light when it impinged on a finite edge, or grazed along a boundary surface. To account for this behaviour, a diffracted ray was introduced. Diffracted rays are produced when incident rays hit edges, corners or vertices of boundary surfaces, or when they graze such surfaces. The magnitude of the diffracted field is a function of the distance from the source, the polarization of the incident field, and the angle of incidence. The phase of the diffracted field is proportional to the electrical length from some reference point. Once computed, the diffracted field is combined with the GO fields to produce the total field. A good introduction of the fundamentals of GTD can be found in [83,90,91].

The geometry used to formulate the GTD solution for DRAs on finite ground planes is shown in Figure 5.1. Some of the fields radiating from the DRA, will graze along the ground plane and impinge upon the edges, producing diffracted fields. These fields are responsible for the ripples in the pattern as well as for spill-over radiation into the back lobes. The total radiated field (E_t) can then be expressed as a vector sum of the GO fields (E_{go}) and the diffracted fields (E_d):

$$E_t = E_{go} + E_{d1} + E_{d2} \quad (5.1)$$

The strength of the diffracted field can be determined by calculating the diffraction coefficient. The appropriate diffraction coefficient for this application is well known and is given in Appendix B [83].

Figure 5.1 DRA on finite ground plane geometry for calculating GTD rays

The diffracted fields from each edge are then calculated as:

$$\mathbf{E}_{d_i} = 0.5 \cdot \mathbf{E}_{go}(\theta = \pm 90^\circ) \cdot D(d_i, \psi_i, n) \cdot \frac{e^{-jk d_i(1 \pm \sin \theta)}}{\sqrt{d_i}} \cdot e^{-jkr} \quad (5.2)$$

where: $i = 1, 2$, corresponding to the two edges;
 $D(d_i, \psi_i, n)$ is the diffraction coefficient;
 θ, r, d_i and ψ_i are shown in Figure 5.1;
 n is the wedge angle number ($n=2$ for a plane wedge);
and the $\pm$ refers to $i=1$ and $i=2$, respectively.

These equations assume that the finite edges are linear and extend to infinity in the orthogonal direction (i.e. out of the page in Figure 5.1). The equations can also be used to approximate diffraction from circular ground planes, as long as the radius of curvature is large enough to approach linear ground

planes. By substituting the appropriate E_{go} model in eqn. (5.1), this technique was used to predict the radiation patterns of the DRAs when mounted on finite ground planes. Source codes written in C to calculate the diffraction coefficients are found in Appendix B.

5.3 DRAs on Finite Ground Planes

To verify the accuracy of the GTD model, several DRAs were mounted on finite ground planes and their radiation patterns were measured. Figure 5.2 shows the geometry of two configurations used to determine the validity of GTD as a model for DRAs. Figure 5.2(a) shows a cylindrical ring DRA mounted on a square ground plane. The DRA is excited in the $TM'_{01\delta}$ mode by a coaxial probe of height 6.5 mm. The probe itself does not radiate, since the impedance presented by the DRA is much lower than that of free space. The probe thus couples its energy into the DRA which then radiates. Figure 5.2(b) shows a cube DRA mounted on a circular ground plane. The DRA is excited in the $TM'_{11\delta}$ mode by the same coaxial probe. Radiation patterns of these antennas were taken in an anechoic chamber over a range of frequencies corresponding to the impedance bandwidth of the resonators.

GTD was used to predict the behaviour of the above antennas when mounted on the finite ground planes. For the cylindrical ring DRA operating in the $TM_{01\delta}$ mode, the radiation pattern of a short electric monopole (0.01λ) on an infinite ground plane was used for the GO fields for the GTD model. The GO fields for a short monopole are:

$$E_{\phi}(\theta, \phi) = \begin{cases} \sin(\theta) & 0 < \theta < 90 \\ 0 & 90 < \theta < 180 \end{cases} \quad (5.3)$$

Figure 5.2 Probe-fed DRAs on finite ground planes

For the cubic DRA operating in the $TM_{11\delta}^c$ mode, the radiation pattern of a short horizontal magnetic dipole (0.01λ) on an infinite ground plane was used for the GO fields. The GO fields for a short horizontal magnetic dipole are:

$$E_{go} = \begin{cases} \cos(\theta) & 0 < \theta < 90 \\ 0 & 90 < \theta < 180 \end{cases} \quad (5.4)$$

A comparison of the predicted and measured E-plane patterns is shown in Figures 5.3 and 5.4. The patterns have been normalized to their peak values to facilitate the comparison. For both the cylindrical and cubic DRA geometries, it can be seen that the GTD predictions are in close agreement to the measured patterns. GTD predicts fairly accurately the number of ripples, the magnitude of the ripples, the location of the pattern peak, and the level of radiation at the horizon. There is less agreement below the horizon due to scattering from the complex structure of the antenna positioner, which was not accounted for in the model.

GTD was also used to predict the pattern of the rectangular FRA measured in Chapter 4. As shown in the geometry of Figure 5.5, the launcher block, which is used to hold the coaxial connector, is located only 2.5" from the FRA, and is physically larger than the FRA. The block thus will act as a scatterer and will impact on the radiation pattern. This scattering effect is visible in the E-plane pattern (Figure 4.5) of the FRA and manifests itself as large, unsymmetrical ripples. (Ripples in the H-plane due to the finite ground plane are negligible because of the DRA pattern null at the ground plane elevation, resulting in a small diffracted field.) This asymmetry arises because the distance from the launcher block to the FRA is different than that from the FRA to the edge of the ground plane.

Figure 5.3 Measured vs. GTD patterns for the cylindrical DRA at 6.5 GHz

Figure 5.4 Measured vs. GTD patterns for the cube DRA at 3.7 GHz

Figure 5.5 Slot-fed rectangular FRA on finite ground plane geometry

Thus, the magnitude and phase of the diffraction field from the block is different than those from the edge of the ground plane. A very simple model was used to try and predict this diffraction. In the model, the launcher block was treated as a simple ground plane edge, so that the same diffraction coefficient could be used. The resulting GTD pattern is shown in Figure 5.6, where the theoretical pattern has been normalized to the peak of the measured pattern to facilitate comparison. GTD predicts the number of ripples and their location fairly accurately, but the actual ripple depth is several decibels deeper than predicted. To obtain more accurate results, a more complex model would have to be used, which would account for the second order diffraction from the various edges of the block. Nevertheless, use of this simple model does allow one to get a reasonable estimate of the ripple magnitude.

Figure 5.6 Measured vs. GTD patterns for the slot-fed FRA at 10 GHz

5.4 GTD-Simulated Finite Ground Plane Effects

This section uses GTD simulations to investigate the effects of varying the size of the ground plane on the radiation patterns of DRAs. Two parameters of importance were looked at: the magnitude of the ripples, and the location of the beam peak. Ripple magnitude is important in most applications where an estimation of antenna gain is required. The finite ground plane can introduce large-magnitude ripples in the radiation pattern, which may adversely affect the gain in the desired coverage region. The GTD code listed in Appendix B was used to predict these effects on DRAs. Figure 5.7 shows the effect on the ripple magnitude of the E-plane pattern for a DRA antenna operating in the $TM'_{11\delta}$ mode for various ground plane sizes. Figure 5.8 plots the ripple magnitude as a function of ground plane radius. As can be seen, to achieve a 1 dB ripple, a ground plane with about a 7.5λ radius is required. As the ground plane radius is increased beyond 10λ , the accompanying decrease in the ripple magnitude is seen to be quite small.

An estimate of the beam peak angle is also important for low elevation-angle coverage applications such as mobile satellite communications. Figure 5.9 shows the effects on the location of the E-plane beam peak for various ground plane sizes for a DRA operating in the $TM'_{01\delta}$ mode. The finite ground plane size has a significant impact on the location of the beam peak. For very small ground planes, the location of the beam peak is nowhere near the horizon ($\theta=90^\circ$) as is expected for a short monopole antenna on an infinite ground plane. Figure 5.10 plots the beam peak location as a function of ground plane radius. Varying the ground plane size from a radius of 1λ to 10λ can lower the beam peak from about 54° to about 74° . These effects should be taken into consideration when using DRAs for low elevation-angle coverage.

Figure 5.7 Effect of the ground plane radius on ripple magnitude

Figure 5.8 Ripple magnitude vs. ground plane size for a horizontal magnetic dipole

Figure 5.9 Effect of the ground plane radius on beam peak

Figure 5.10 Beam peak angle vs. ground plane size for a short monopole

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5.5 Discussion

The Geometric Theory of Diffraction has been shown to predict reasonably accurately the effects of finite ground planes on dielectric resonator antennas operating in the $TM'_{01\delta}$ and $TM'_{11\delta}$ modes. Good agreement was found in the ripple magnitude, beam peak location, and radiation level at the horizon between measured and simulated patterns. By combining simple models (short magnetic dipole or short electric monopole) with GTD, several advantages over a full-wave rigorous solution are obtained: this technique is independent of the DRA parameters (their size, shape, and dielectric constant); as long as the appropriate DRA mode is excited, the GTD model can be used to predict the radiation pattern; GTD is fast, computationally efficient, and requires negligible memory storage; execution time on a personal computer is on the order of a few seconds; finally, GTD could be applied to DRAs operating in higher-order modes with simple modifications, by using the GO field appropriate to the desired DRA mode.

Using GTD, graphs were compiled showing the effects of the ground plane size on ripple magnitude and beam peak location. These graphs are quite general, applicable over a broad frequency range and are independent of the DRA parameters. They are useful for predicting gain variations and also for low elevation angle coverage applications such as mobile satellite communications.

6. Ferrite Rod Antennas

6.1 Introduction

This chapter focuses on the ferrite rod antenna: the second class of ferrite antennas to be studied in this thesis. Unlike the FRA, ferrite rod antennas are travelling wave antennas, which produce end-fire radiation patterns. Ferrite rod antennas exhibit similar behaviour to dielectric rod antennas, which were discussed in Chapter 3. As was the case for the FRA, the permeability of the ferrite rod can be controlled by applying a d-c magnetic bias. For FRAs, this property was used to shift the resonant frequency and to change the polarization of the antenna. For the ferrite rod, this property is used to control the relative phase along the radiator. This magnetic phase control makes it possible to produce an array with electronic beam steering capabilities, which is a useful feature for portable satellite communications. Also, the ferrite rod antenna, being a travelling wave antenna, can be made to have a higher gain than printed elements by increasing the length of the rod, and it does not suffer from conductor losses. By using these rods in an array, it should be possible to reduce the array size and the number of active devices. This chapter introduces a novel ferrite rod which is designed to be a combined antenna and phase shifter element for operation at K band.

6.2 Ferrite Rod Geometry

The geometry of the ferrite tapered rod is shown in Figure 6.1 [92]. The ferrite material used was TTVG-1850 (see Table 3.1). The ferrite was machined into a rod of square cross-section, consisting of a 30 mm length with a uniform cross-section and a 30 mm length with a tapered cross-section. WR42 waveguide was used to feed the ferrite rod. A solenoid coil

Figure 6.1 Geometry of ferrite antenna/phase shifter element

was placed around the length of uniform cross-section to provide the required d-c magnetic field bias. To suppress unwanted Faraday rotation, discussed in Chapter 3, the rod was placed between two metal plates as shown in Figure 6.1.¹ This antenna element combines the concept of the ferrite rod antenna described in [93] with a suppressed-rotational reciprocal phase shifter described in [94-99] to produce a radiating element with magnetic phase control.

6.3 Design of Tapered Rod

The rod dimensions were designed for operation at 20 GHz, which is in the band of interest for K-band satellite communications. A rod with medium gain of between 10 and 12 dBi was desired, so that in the eventual array application, fewer elements would be required than a printed array of microstrip patches. To achieve this element gain, the tapered length was chosen to be two free-space wavelengths. The tapered section was used to reduce the standing wave amplitude on the rod, induced by the impedance mismatch at the ferrite-to-air boundary. The maximum and minimum cross-sections were based on the equations for dielectric rod antennas [100]:

$$d_{\max} = \frac{\lambda_0}{\sqrt{\pi(\epsilon_r - 1)}} \quad (6.1)$$

$$d_{\min} \approx \frac{\lambda_0}{\sqrt{2.5\pi(\epsilon_r - 1)}}$$

These equations have been used for rods with linear taper lengths of down to about $2\lambda_0$ with good results [33]. The values for $d_{\max}$ and $d_{\min}$ were

¹ These plates, called Faraday suppression plates, can be thought of as short-circuiting the component of the electric field parallel to the plates, thus only the perpendicular E-field component remains, preventing Faraday rotation. These plates become more effective as their separation from the ferrite rod is decreased [94-99].

evaluated using eqn. (6.1) for operation at 20 GHz and are shown in Figure 6.1. The rod was excited by inserting it into the centre of a WR42 metal waveguide, operating in the TE_{10} mode. The insertion depth (L_d) of the ferrite rod into the waveguide was optimized experimentally to provide a good impedance match. An aluminum fixture (of thickness $t=3/16"$) was used to support the ferrite rod and to seal the open end of the waveguide in order to prevent unwanted radiation.

The applied magnetic field strength was estimated using the well-known equation for a solenoid:

$$H = NI/L_p \quad (6.2)$$

where:

N is the number of turns;
 I is the applied current; and
 L_p is the length of the solenoid.

The solenoid was 20 mm in length with approximately 150 turns. The solenoid current was varied from 0 mA to 300 mA, which corresponds to a magnetic bias varying from 0 Oe to 28 Oe.

6.4 Experimental Results

6.4.1 Phase control vs. Applied Bias Field

The phase shift versus applied magnetic bias was measured on a network analyzer. The ferrite rod assembly was connected to one port and a pyramidal horn antenna to the other port. The S_{21} phase was measured as the bias current was increased from 0 mA to 300 mA. This procedure was repeated for various spacings (s_p) of the Faraday suppression plates. The results are shown in Figure 6.2. The phase shift curves have been normalized by the length of the phase shifter section. The amount of phase shift is seen to increase with plate separation. A relative phase shift of up to

$103^\circ/\lambda_0$ was achieved for the plates separated by 5.0 mm. The change in the return loss with magnetic bias is shown in Figure 6.3. The return loss is somewhat sensitive to the applied bias, but a return loss of better than 10 dB was obtained for a plate separation of $s_p=4.25$ mm. Figure 6.4 depicts the effect of frequency on the phase shift. The phase shifts were measured over a range of 20 GHz to 21 GHz. The change in phase with frequency remains small over the linear region of the curves but increases somewhat in the saturation region.

Figure 6.2 Phase change vs. applied magnetic field for various plate separations

Figure 6.3 Change in return loss vs. applied magnetic field

Figure 6.4 Changes in phase shift vs. applied field at 20/21 GHz

6.4.2 Radiation Patterns

The radiation patterns of the tapered rod were measured in an anechoic chamber. Patterns were measured in both principal planes for various magnetic biases and plate separations. Typical patterns are shown in Figure 6.5 for the unbiased ferrite and Figure 6.6 for the biased ferrite. The co-polarized patterns did not change significantly with the applied magnetic bias. The gain and the cross-polarization levels at boresight versus magnetic bias are shown in Table 6.1 for the plates separated by 5.0

Figure 6.5 Measured radiation patterns of the unbiased ferrite rod at 20.5 GHz.

Figure 6.6 Measured radiation patterns of ferrite rod with 300 mA bias at 20.5 GHz

mm at 20.5 GHz. There was only a small change in boresight gain with the applied magnetic field, while the cross-polarization level degraded by almost 6 dB when the maximum field was applied.

Bias Current (mA)	Magnetic Field (Oe)	Gain (dBi)	Cross-Pol Level Below Peak (dB)
0	0	11.1	20.29
100	9.4	11.5	18.34
200	18.8	11.4	17.31
300	28.2	11.3	14.4

Table 6.1 Gain and cross-pol levels of the ferrite rod vs. applied bias for $s_p = 5.0$ mm, at 20.5 GHz.

Increasing the plate separation from 5 mm to 6.3 mm caused a shift of the gain peak from 20.5 GHz to 20.9 GHz (without a significant change in mismatch loss), while the cross-polarization level at boresight degraded to 11 dB below the peak value. This is not unexpected, since moving the plates further apart allows more Faraday rotation, thus causing an increase in the cross-polarization component.

The equations for the radiation patterns of the dielectric rods (eqns. (3.5) and (3.6)) were used to predict the patterns of the ferrite rod antenna and are shown in Figure 6.7 along with the measured patterns. The guided wavelength at 20.5 GHz was calculated from Marcattili's equations (3.4) to be 15.1 mm, using the maximum cross-sectional dimension (2.28 mm) of the tapered rod. The theoretical patterns have been normalized to the peak of the measured patterns. For the Schelkunoff equations (3.6), effective cross-sectional dimensions of three times the physical cross-section were used to obtain reasonable patterns. The agreement between the measured and theoretical patterns is fair, with the main beam and first nulls being adequately predicted. Several factors contribute to the discrepancies

between the patterns. The theoretical equations were developed for rods of uniform cross-section, while the measured rod had a tapered cross-section to minimize reflections at the ferrite-to-air interface. As well, the equations do not take into account the feeding mechanism nor the Faraday suppression plates, which also radiate to some degree and contribute to the pattern, especially to the sidelobes. The Huygen equations (3.5) are more useful in predicting patterns than are the Schelkunoff equations (3.6), since one need not introduce effective cross-sectional dimensions in order to obtain reasonable agreement with measured patterns. This equation gives a fairly good idea of the expected beamwidth and the location of the first nulls. To obtain closer agreement, more accurate modelling techniques would have to be considered, which would take into account the tapered cross-section of the rod as well as the feed mechanism. This would entail a much greater degree of complexity and increased computation time.

6.4.3 Radiation Efficiency

The radiation efficiency (η) of the ferrite rod was estimated using the following formula:

$$\eta = 10^{\frac{(G - D)}{10}} \quad (6.3)$$

where:

G is the measured gain (in dB) of the ferrite rod

D is the estimated directivity (in dB) of the ferrite rod.

The directivity D of the ferrite rod was estimated using [105]:

$$D = 10 \log \left(\frac{32400}{\theta_e \cdot \theta_h} \right) \quad (6.4)$$

where:

θ_e , θ_h are the half power beamwidths (in degrees) in the E-plane and H-plane, respectively.

The above directivity formula is most accurate for pencil beam antenna patterns with low sidelobes and is used in the case of the ferrite rod antenna as an approximation of the actual directivity. The estimated efficiencies for various magnetic bias strengths were calculated and are tabulated in Table 6.2. The efficiency is seen to remain fairly constant with bias current.

Bias Current (mA)	Measured Gain (dBi)	E-plane Beamwidth (deg)	H-plane Beamwidth (deg)	Directivity (dB)	Radiation Efficiency
0	11.1	43.5	48.1	11.9	85%
100	11.5	42.5	46.4	12.2	85%
200	11.4	44.1	47.7	11.9	89%
300	11.3	44.1	47.1	11.9	87%

Table 6.2 Estimated radiation efficiency of ferrite rods vs. applied magnetic bias

6.4.3 Slot Coupling

In the above sections, the ferrite tapered rod was excited by inserting the rod into the centre of the mouth of a WR42 waveguide (Figure 6.1). The insertion depth of the rod was determined experimentally to provide a good return loss. The measured radiation pattern of this configuration (without Faraday Suppression plates) is shown in Figure 6.8. The patterns have a gain of approximately 15 dBi, with low sidelobe levels, and cross-polarization levels about 12 dB below the co-polarized peak at boresight. The resultant patterns indicate that this configuration is an effective way of exciting the ferrite rod. Although the waveguide feed results in effective excitation, it is quite bulky and not compatible with planar technology. An alternative method for exciting the ferrite rod is thus desirable. An experimental investigation of a microstrip line coupled slot was carried out to determine its potential as a feed for the ferrite rod. The microstrip line coupled slot configuration is shown in Figure 6.9. The ferrite rod was

Figure 6.8 Measured waveguide-fed patterns of ferrite rod (no plates)
at 20 GHz

Figure 6.9 Slot-coupled feed for ferrite rod antenna

placed over the slot and its position was varied experimentally to provide maximum return loss. Also, several slot dimensions were used, again to optimize the return loss. The return loss of the optimized slot feeding the ferrite rod is shown in Figure 6.10. Radiation patterns in the principal planes were then measured and shown below in Figure 6.11. As can be seen from these figures, the radiation patterns of the slot-fed ferrite rod are greatly deteriorated from the waveguide-fed patterns. The patterns show a deterioration in gain, an increase in sidelobe and backlobe levels and a high cross-polarized component. The patterns themselves are also quite distorted. These results indicate that the slot itself must be radiating with deleterious effects on the radiation pattern of the rod. Instead of efficiently coupling the energy from the slot to the rod, the ferrite loads the slot with a high impedance, causing the slot to radiate. This is unlike the case of the slot-fed FRA, where the FRA presents a lower impedance than free space so that the slot couples energy efficiently into the FRA. Unfortunately, it will be difficult to find a method to reduce the spurious radiation from the slot. One way is to use a smaller slot size, however, preliminary results indicate that an acceptable return loss cannot be achieved using smaller slot sizes. Thus the microstrip slot fed aperture does not appear to be a good candidate for exciting the ferrite rod antenna.

6.5 Discussion

In this chapter, a novel tapered ferrite rod element with magnetic phase control was introduced, and its performance was evaluated using a prototype antenna. Phase shifts of over $100^\circ/\lambda_0$ were achieved at 20.5 GHz with little change in the gain and co-polarized patterns, and radiation efficiency of about 85%. One method of increasing the maximum attainable phase shift is by increasing the separation of the Faraday suppression

Figure 6.10 Return loss of slot-fed ferrite rod

plates. As the plates are moved further apart, they are less effective in suppressing the Faraday rotation, which results in a larger phase shift. The disadvantage of this approach is that it leads to increased cross-polarization levels due to the increased Faraday rotation. Optimization of material parameters ($4\pi M_s$ and ϵ_r), and aspect ratio should also lead to increases in the maximum achievable phase shift per unit length. The return loss was seen to be sensitive to the applied magnetic bias through the ferrite rod. This is due to the change in the effective permeability of the ferrite, which in turn changes the impedance of the rod. The sensitivity of the return loss should be minimized by either tapering the section of the rod inserted into the waveguide, or by attaching a short section of dielectric rod in front of

Figure 6.11 Measured radiation patterns of slot-fed ferrite rod at 20.3 GHz

the rod to act as a transformer [94].

The ferrite rod was designed using a combination of Marcatili's design equations for the guided wavelength in a rectangular dielectric waveguide (eqn. (3.4) with eqn. (6.1)) to calculate the tapered rod cross-sectional dimensions. Huygen's equations (3.5) were found to predict the main beamwidths and first nulls of the ferrite rod radiation patterns to within about 10-15%. These equations allow for a quick, first order design for these ferrite rods, which can then be optimized using more rigorous (but time consuming) analytical or numerical techniques.

The metal waveguide feed was found to offer an effective way of exciting the ferrite rod, resulting in medium gains, fairly high efficiencies, and good cross-polarization levels. Metal waveguides, however, are bulky, costly, and not amenable to integration. Alternative feed mechanisms are required to make ferrite rods more attractive as array elements for K-band satellite communications.

7. Linear Arrays of DRAs and Ferrite Rods

7.1 Introduction

To achieve the radiation patterns and gains desired for most satellite communication applications, linear or planar arrays are required. The previous chapters focused on the characterization of individual ferrite radiating elements. This chapter presents preliminary results on several linear arrays, designed and fabricated using these novel antenna elements. Measurements on the return loss and radiation patterns were carried out to provide an indication as to the viability of these antennas in an array environment. This chapter looks at several linear arrays of dielectric resonator antennas (which behave similarly to FRAs), designed to operate at 20 GHz, which is in the band of interest for K-band satellite communications. Various feeding structures for exciting these linear arrays are investigated and their performance is evaluated. Linear arrays of ferrite rod antennas are also examined, and their electronic scanning ability is evaluated, which is of interest for satellite tracking applications.

7.2 Linear Arrays of Dielectric Resonator Antennas

7.2.1 Overview

To date, most research on DRAs has been limited to single elements. Little work has been carried out on arrays [101,103,104]. This section investigates the performance of linear arrays of dielectric resonator antennas (DRAs) fed by various methods [102]. A linear array of DRAs fed by a dielectric image guide (DIG) is first examined. DIGs have the potential of being very low-loss transmission lines, which is advantageous at millimeter-wave frequencies where copper losses are high. The second feed method is microstrip feed line. The microstrip line is more lossy at

millimeter-wave frequencies than the DIG, but it is much easier to fabricate and does not require a bulky transition. An array of parasitically coupled elements is also investigated. This array has the advantage of a simple feed structure, thus providing the potential for lower losses.

7.2.2 Dielectric Image Guide Fed DRA Linear Array

A schematic of the linear array of DRAs fed by a DIG is shown in Figure 7.1. DRAs are excited through electromagnetic coupling from the DIG. The coupling amplitude can be controlled by varying the distance between the DIG and the DRAs. The dimensions of the DIG were designed for operation at 20 GHz using Marcatili's equations (3.4). The 12" guide was fabricated from Rexolite ($\epsilon_r=2.55$) and using eqn. (3.4), the cross-sectional dimensions were determined to be $a = 8$ mm and $b = 4$ mm (as shown in Figure 3.4). A 1" linear taper was cut at each end of the rod to provide an impedance transition from the DIG to the metal waveguide launchers. The length of the taper is a compromise between the compactness of the transition and the performance of the impedance transformer. Non-linear tapers (such as exponential) could be used to make the transition more compact and improve the impedance transformation, but at the expense of increased manufacturing complexity. The return loss of the DIG terminated in a short circuit is shown in Figure 7.2. At 20.2 GHz, the return loss is 3 dB. The insertion loss can be determined by noting that the return loss represents the loss in the transmitted power as it travels to the short circuit and is reflected back to the source. Thus, the insertion loss of the DIG (which includes the radiation losses from the waveguide-to-dielectric interfaces) was $0.075 \text{ dB}/\lambda_0$.

Figure 7.1 Linear array of DRAs fed by a dielectric image guide

Figure 7.2 Return loss of DIG terminated in a short circuit

To measure the spurious radiation from the waveguide-to-DIG transitions, pattern measurements were carried out on the unloaded DIG terminated in a short circuit. Figures 7.3 and 7.4 show the level of spurious radiation of the DIG, without and then with absorbing material around the transitions. It is evident that without the absorber, there is a significant amount of spurious radiation. The addition of absorber removes a large component of the spurious radiation, but will contribute to a degradation in the gain of the DRA array.

A 20-element array was then constructed by placing the DRAs adjacent to the DIG and spaced $s = 12$ mm apart. The DRAs were fabricated from Duroid ($\epsilon_r = 10.8$) with dimensions $a = 5$ mm, $b = 2$ mm, and $d = 1.6$ mm (Figure 7.1), determined using eqn. (3.1) to resonate at 20 GHz. The inter-element spacing (s) was chosen to be approximately one

guided wavelength. The resultant return loss is shown in Figure 7.5. Radiation pattern measurements were taken of the array and are shown in Figure 7.6. The beam peak of the pattern was found to scan with frequency. The peak occurred at boresight at a frequency of 20.4 GHz, with a measured gain of 11.4 dBi. The maximum gain of 13.2 dBi occurred at 21.8 GHz, where the beam peak is scanned to 6° . The beamwidth ranged from 3.9° at 20 GHz to 3.1° at 21.8 GHz. By treating the DIG-fed DRA array as a linear array of uniform magnetic dipoles, classic array theory was used to predict the radiation pattern [83]. The radiation pattern was calculated by multiplying the element pattern of the magnetic dipole eqn. (3.3) with the array factor of a uniformly spaced, equal amplitude linear array.

Figure 7.7 compares the measured pattern to theoretical patterns using array theory, where the theoretical pattern has been normalized to the peak value of the measured pattern for comparison. The theoretical pattern was computed by assuming a uniform element excitation. The array factor of the 20-element linear array was multiplied with the short magnetic dipole element model for the DRAs. The agreement between the measured and theoretical patterns is seen to be quite good, with the theoretical pattern beamwidth of 3.2° , compared to the measured value of 3.9° , and the theoretical sidelobe levels matching closely to those measured.

Figure 7.3 H-plane spurious feed radiation from the DIG

Figure 7.4 E-plane spurious feed radiation from the DIG

Figure 7.5 Return loss of the 20-element DRA array fed by the DIG

Figure 7.6 Radiation patterns of the 20 element DRA array

Figure 7.7 Theoretical vs. measured patterns of the 20-element DRA array

7.2.3 Microstrip Line Fed DRA Linear Array

The second feed structure used to excite an array of DRAs was a microstrip line. Although the losses in a microstrip transmission line are higher than for a DIG at millimeter-wave frequencies, microstrip offers other advantages, such as ease of fabrication and integration with planar technology. A ten-element linear array of DRAs fed by a microstrip line is shown in Figure 7.8. A 125 mm long $50\ \Omega$ microstrip line printed on 0.635 mm (.025") thick Duroid ($\epsilon_r = 2.2$) was used to feed the array. The DRA elements (with identical dimension to those used in the DIG array) were spaced 8 mm apart (approximately one guided wavelength). The measured return loss of the array is shown in Figure 7.9. The insertion loss of the microstrip line was determined to be $0.3\ \text{dB}/\lambda_0$ at 20 GHz. The radiation patterns of the array are shown in Figures 7.10 and 7.11. A peak gain of 13.2 dB was measured at 19.9 GHz with a beamwidth of 8.6° and with a cross-polarized level 23 dB below the beam peak. Again, the theoretical pattern was computed using classical array theory and the agreement with the measured pattern is seen to be quite good, as shown in Figure 7.12. The theoretical beamwidth was 9.5° compared to the measured value of 8.6° . (Again, the theoretical pattern has been normalized to the peak of the measured pattern to facilitate comparison.)

Figure 7.8 Linear array of DRAs fed by a microstrip line

Figure 7.9 Return loss of the 10-element DRA array fed by microstrip

Figure 7.10 E-plane patterns of the 10-element DRA array at 19.9 GHz

Figure 7.11 H-plane patterns of the 10-element DRA array at 19.9 GHz

Figure 7.12 Theoretical vs. measured patterns of the 10-element DRA array

7.2.4 Slot-Fed Parasitic DRA Linear Array

The third linear array configuration examined consisted of a slot-fed element with adjoining parasitic elements. Figure 7.13 shows the configuration investigated. The first element was the driven element, fed by aperture coupling, while the remaining elements were parasitically coupled to the driven element. The potential advantages of this geometry include the feed line's location underneath the ground plane, thus eliminating its interference with the radiation pattern. Also, by adding parasitic elements, a single feed line was used to excite a sub-array, thus reducing the line losses and feed complexities in a large planar array configuration. The parasitic elements were added to investigate the potential for enhancing the gain of the single-fed DRA. The spacing between elements was adjusted experimentally so as not to significantly reduce the return loss bandwidth of the single element. Also, adding a fourth parasitic element caused no noticeable change in the return loss. Figure 7.14 shows the return loss of the single element and the parasitic array. Figure 7.15 compares the pattern of an individual element to the array of parasitic elements. Table 7.1 summarizes the effects on the beamwidth by adding up to three parasitic elements to the driven element. The addition of the parasitic DRA elements resulted in a narrowing of the beamwidth, increase in gain, and squinting of the beam peak. The gain of the four-element array, measured at 20.6 GHz, was 4.3 dB higher than that of the single element. (An array of four driven elements would have a maximum of 6 dB increase in gain, assuming no feed losses and equal phasing between elements.) The beam peak was squinted 50° off boresight due to a phase progression between the elements.

Figure 7.13 Parasitic array of DRAs

Figure 7.14 Return loss of single element and parasitic array

Elements	Beam Width
1	109°
2	54°
3	39°
4	32°

Table 7.1 Beamwidth vs. number of elements

Figure 7.15 Radiation patterns of single DRA vs. parasitic array at 20.6 GHz

7.2.5 Discussion of Linear DRA Arrays

Linear arrays of DRAs were designed using various feeding structures, to operate at 20 GHz, in order to investigate the performance of DRAs as potential candidates for personal communications. It was shown that the dielectric image guide has a lower insertion loss than the microstrip fed array at 20 GHz. The DIG array suffers, however, from high levels of spurious radiation from the waveguide launchers which, although suppressed adequately by absorber material, contribute to a degradation in gain. Improvements could be made to the transition by flaring the waveguide and adopting a non-linear taper in the dielectric, but at the expense of increased complexity and size. Furthermore, an alternative to the waveguide launcher is still required for integration with phased arrays, and it will be difficult to extend the DIG geometry to a planar array, due to the difficulties in realizing power splitters in DIG.

The microstrip-fed linear array shows promise, producing patterns with reasonable gain, acceptable sidelobe levels, and good cross-polarization levels, and is easily integrated with printed structures. This array doesn't suffer from spurious radiation from bulky transitions and can be quite compact. Improvements in gain may be achieved by considering alternative substrate materials and thicknesses to reduce line losses. This architecture should easily be extended to planar arrays, by using microstrip power dividers.

The slot-fed DRA with parasitic elements also shows potential. The use of parasitic elements provides a method for increasing gain without increasing the complexity of the feed network. In addition, the beam peak is scanned towards the horizon by the addition of parasitic elements, which may prove useful for mobile satellite communication applications.

Additional work is required for this array configuration to understand the interaction between the driven DRA and the parasitic elements. The relationship between the amplitude and phase distribution with element spacing will have to be determined in order to predict the array patterns.

7.3 Linear Arrays of Ferrite Rod Antennas

7.3.1 Introduction

This section examines the performance of linear arrays of the ferrite rod elements that were investigated in Chapter 6. Two linear arrays are investigated: a two-element array, where each element is fed by a separate waveguide, and a four-element array where all four elements are fed by a single waveguide. Radiation patterns for each array are provided, along with their beam scan performance.

7.3.2 Two-Element Array

A two-element ferrite-rod array was fabricated as shown in Figure 7.16. Each rod was inserted into the mouth of a separate WR42 waveguide. Spacing between ferrite elements was 22 mm. (This spacing is larger than a wavelength due to the physical size of the WR42 flanges).

Figure 7.16 Two-element array of ferrite rods

A power divider was used to split the power to each element. The separation of the Faraday suppression plates was 5 mm. Radiation patterns for various coil biases at 20 GHz are shown in Figure 7.17. The beam-scan performance of the array is summarized in Table 7.2.

Bias Current (mA)	Beam Peak (deg)	Beam Width (deg)
0	+5.66	16.77
100	-1.60	18.28
200	-3.35	18.78
300	-4.61	18.94

Table 7.2 Summary of beam-scan performance of 2-element ferrite rod array

The initial beam scan was due to unequal cable lengths in the power divider used. As can be seen in Figure 7.17, when the magnetic bias is increased, there is a slight increase in the beamwidth and in the cross-polarization level. A maximum scan range of over 10° was achieved for the bias range.

Figure 7.17 Radiation patterns of the two-element ferrite rod array

Figure 7.17 Radiation patterns of the two-element ferrite rod array

7.3.3 Four-Element Array

The two-element array described above was used to demonstrate the capabilities of the ferrite rod in a scanning array. A disadvantage with the array is that a separate waveguide feed is required for each element. A better solution would be to feed the array of elements using a single waveguide, to reduce cost and complexity. An attempt was made to design and fabricate a four-element ferrite rod array, fed from the side wall of a single WR42 waveguide. A diagram of the complete array is shown in Figure 7.18. The four rods are inserted into square apertures cut through the side wall of the waveguide. The apertures were spaced a guided wavelength apart at 20 GHz, and the waveguide was terminated with a moveable short circuit. Coupling from the waveguide to the rods was achieved by varying the insertion depth of the rods.

For the purpose of demonstrating beam scanning, and to simplify the bias network, the first two elements of the array were left unbiased, while the last two elements were biased together in series. That is, the relative phase of the first two elements were set to zero, while the phase of the last two elements were altered together. The resulting patterns are shown in Figure 7.20. The unbiased pattern of the array is seen to have a null at broadside. This indicates that the elements are out of phase. This is attributed to the penetration depth of the rods into the waveguide. In order to get reasonable coupling, the rods had to be inserted several millimeters into the waveguide. This significantly altered the field configuration within the guide, resulting in a different phase progression than expected along the guide. The four rods were no longer in phase, resulting in a null in the pattern. When biasing was applied, the null was seen to disappear, and there was a limited amount of phase scanning.

Figure 7.18 Four-element array of ferrite rods

Figure 7.19 Radiation patterns of four-element ferrite rod array

7.3.4 Discussion of Linear Ferrite Rod Arrays

In this section, linear arrays of two and four ferrite rods were constructed to demonstrate beam scanning. The rods of the two-element array were each fed by individual waveguides, and a 10° beam scan was demonstrated over the bias range, with little degradation in the co-polarized patterns or the cross-polarization levels. The maximum scan range can be increased by increasing the phase shift range of each element. The simplest way to do this would be to extend the length of each phase shifter. For a fixed phase shifter length, the amount of phase shift can be increased by optimizing the ferrite material (using one with a higher saturation magnetization), optimizing the separation of the Faraday suppression plates, and optimizing the cross-sectional dimensions of the rod. These antennas have significant potential for analog beam-steering applications, such as satellite tracking.

The four element array consisted of a single waveguide feed, with the ferrite rods inserted into the side wall of the waveguide, and was constructed to reduce the amount of waveguide required to feed each rod. This array also demonstrated beam steering; however, the unbiased pattern contained a broadside null, indicating a problem with the phasing between elements. This is attributed to the interactions of the ferrite rods with the internal fields of the waveguide feed, setting up higher order modes internally in the waveguide. This problem might be resolved by tapering the feed ends of the ferrite rods.

8. Summary and Discussion of Future Work

8.1 Summary

This thesis explored various properties of ferrite antennas, in order to investigate their potential use as alternates to printed elements in arrays for personal or satellite communication applications.

In Chapter 2, the various printed and non-printed architectures were reviewed. This review revealed the drawback of printed architectures with respect to increased losses at microwave and millimeter-wave frequencies, which led to the need for amplifiers, resulting in more complexity and higher costs. The chapter also summarized the various non-printed architectures, of which several prototypes demonstrated significant promise, notably a planar array composed of dielectric rods, ferrite substrates, and printed dipoles. This chapter laid the groundwork for the thesis research, which was to investigate novel ferrite antenna elements, characterize their performance, and establish their potential for personal communications applications.

Chapter 3 provided the theoretical foundation for the thesis research by examining in detail the modes of operation of dielectric resonators and dielectric rods, on which the behaviour of their ferrite counterparts was modeled. This chapter also contained a brief review of the behaviour of ferrites at microwave frequencies, to provide theoretical support for the observed phenomena exhibited by the FRAs and ferrite rods.

In Chapter 4, the first of two classes of ferrite antennas was introduced: the ferrite resonator antenna (FRA). It was demonstrated that the application of a static magnetic field bias to FRAs of rectangular and cylindrical shapes altered either the resonance frequency or the polarization of the FRA. The FRAs were characterized in detail, with measurements of

such parameters as radiation patterns, return loss, and radiation efficiency, and the effect of magnetic bias on these parameters. FRAs were found to offer most of the advantages of microstrip patches, with the additional advantages of wider bandwidths, high radiation efficiency, simple coupling schemes, and active element control.

Chapter 5 studied the effects of finite ground planes on the radiation patterns of dielectric and ferrite resonator antennas. For portable communication applications, the antenna will invariably be mounted on ground planes of finite extent (ex. briefcase terminals, roof mounting on vehicles, hand-held sets) which may dramatically alter the pattern from that of free space. The Geometric Theory of Diffraction (GTD) was used to predict the finite ground plane effects. Several examples were investigated, with resulting theoretical predictions agreeing closely with measured data.

In Chapter 6, the second class of ferrite antennas was introduced: the ferrite rod antenna. A novel configuration was presented which combined the radiating element with an analog phase shifter, to produce an active element. The ferrite rod antenna was characterized in terms of its radiation patterns and phase shift performance with applied magnetic bias. This element was found to be a potential candidate for portable satellite communications, due to its relatively high gain and analog phase control which would allow for electronic beam steering in an array.

Finally, in Chapter 7, arrays of dielectric resonator antennas and ferrite rods were investigated. DRAs were chosen, since their behaviour is similar to unbiased FRAs. Linear arrays of DRAs with various feeds were designed to operate at 20 GHz, and their performance was evaluated. The linear DRA fed by a microstrip line showed significant promise, achieving good radiation patterns, and acceptable gains. This architecture combined

printed and non-printed technology and should be easily extended to 2D planar arrays. A preliminary investigation on an array of parasitically fed DRA elements also demonstrated potential. Higher gains and narrower beamwidths were achieved by adding parasitic elements, without an increase in feed complexity.

Linear arrays of ferrite rod antennas were also investigated in this chapter. A two-element rod antenna was fabricated to demonstrate the potential for beam scanning. A 10° beam scan was demonstrated, but this value could be significantly increased by optimizing the material parameters of the ferrite rod.

8.2 Discussion of Future Work

This section discusses a few of the possible avenues available for future work in the area of ferrite resonator antennas and ferrite rod antennas.

8.2.1 Latching of the Ferrite Antennas

For both types of ferrite antennas investigated in this thesis, the applied magnetic bias was continuously supplied, using either a permanent magnetic or a d-c coil. To reduce power consumption, one method would be to apply pulsed magnetic fields in order to latch the ferrite into its remnant-magnetization state. Some preliminary work using cylindrical disk FRAs (with diameters of 16 mm and varying heights from 2 mm to 5 mm) fed as in Figure 4.7 was carried out to determine the feasibility of latching. Latching was investigated by applying a magnetic field with a permanent magnet for a short duration (a few seconds), and then removing the magnet. The amount of latching was found to be related to the disk height. Remnant magnetization occurred for the 4 mm and 5 mm disks, while none was

observed for thinner disks. This remnant magnetization manifested itself as a latching of the resonant frequency. For the 5 mm disk, when an x-directed bias (as defined in Figure 4.7) was applied and removed, the frequency shifted up by 1.2% and remained latched. The unbiased resonant frequency could be restored by applying a y-directed magnetic field. With the 4 mm high disk, only a 0.9% shift in latched resonant frequency was observed, while no latching was observed for the thinner disks. Additional research could be carried out to further explore the possibility of the latching of the resonant frequency of FRAs. The amount of latching should be increased by using ferrites with higher saturation magnetizations and high remnant ratios. Exploring other FRA geometries or feed mechanisms might also lead to increases in latching performance.

8.2.2 Modeling of FRAs

The dielectric waveguide model of the FRAs (presented in Chapters 3 and 4) gives fairly reasonable estimates for the resonant frequency and the bandwidth of the resonators, usually within 5% to 10%. More rigorous solutions could be investigated using numerical techniques such as the transmission-line matrix method (TLM) or the finite elements method (FEM). The TLM method has been used to model the unbiased aperture-fed rectangular DRA as shown in Figure 4.1 [106]. The resonant frequency was estimated within 1%, while the bandwidth estimate was within a few percent. Problems occur, however, when the FRA is biased. Although a technique is available in both TLM and FEM to account for anisotropic (ferrite) material, the user must know the tensor permeability at each cell. A reasonable estimate of the tensor elements can be made if the ferrite material is uniformly saturated, however it becomes a difficult challenge to determine

these tensor elements when the ferrite is unsaturated or not uniformly saturated. A technique to determine the tensor permeability at each point within an arbitrary-shaped ferrite which is immersed in an externally applied (and perhaps non-uniform) magnetic field would contribute significantly to the development of an analysis tool for predicting the effects of magnetic bias on FRAs.

8.2.3 Higher-Order Modes of FRAs

This thesis has focused on the lowest order modes of both the rectangular and cylindrical FRAs, which were found to radiate as short dipoles. Future research could investigate the higher-order modes of FRAs. Higher-order modes, such as the magnetic quadrupole mode [107] would excite conical radiation patterns, whose beam peaks would be pushed down towards the horizon, which would be useful for mobile satellite communications. Of significant interest would be a magnetically switchable linear-to-circular polarized cylindrical disk FRA operating at higher-order modes to produce conical patterns.

8.2.4 Finite Ground Plane Modeling

Additional work could be carried out to improve the GTD model used to predict the effects of finite ground planes on the radiation patterns of the FRAs. The model could be extended to include higher order diffraction effects to model diffraction from more complex structures such as the coaxial launcher block used extensively in this thesis work (Figure 5.5). Also one could extend the model to include the effects of circular ground planes with small radii of curvature (which cannot be approximated with the straight edge diffraction coefficient).

8.2.5 Ferrite Rods

Further research on ferrite rods could include the investigation of alternate feeding mechanisms better suited for integration with existing planar structures. Also of potentially significant interest would be an investigation into the generation of circular-polarized radiation patterns. These patterns could be generated by optimizing the cross-sectional dimensions of the rectangular rod, or by using rods of circular cross-section.

8.2.6 Arrays

In Chapter 7, several linear arrays of DRAs and ferrite rods were investigated. There are many avenues for additional research for these arrays. Planar arrays of DRAs could be investigated, especially based on the microstrip-fed architecture, which lends itself well to an extension into a planar array. Another possibility would be the investigation of designing a Yagi-Uda type linear array, using a slot-fed (driven) DRA element with parasitic director and reflector DRA elements. This would again prove significantly useful for low-angle coverage applications, since the Yagi-Uda array would nominally produce an end-fire pattern. A thorough study of mutual coupling between DRA elements could also be undertaken as part of the array work.

Finally, there are several areas of future work to be pursued to further the development of ferrite rod arrays. Of significant interest would be alternate feeding mechanisms to excite the rods, and the development of planar array architectures.

Appendix A

A.1 Field Equations for Rectangular DRA $TE_{11\delta}^z$ Mode [43]

(See Figure 3.1)

$$H_x = [\delta G H_0 / (ad)] \sin(\pi x/a) \cos(\pi y/b) \sin(\pi \delta z/d)$$

$$H_y = [\delta H_0 G / (bd)] \cos(\pi x/a) \sin(\pi y/b) \sin(\pi \delta z/d)$$

$$H_z = H_0 \cos(\pi x/a) \cos(\pi y/b) \cos(\pi \delta z/d)$$

$$E_x = [j\omega\mu_0 H_0 G / (b\pi)] \cos(\pi x/a) \sin(\pi y/b) \cos(\pi \delta z/d)$$

$$E_y = [-j\omega\mu_0 G H_0 / (a\pi)] \sin(\pi x/a) \cos(\pi y/b) \cos(\pi \delta z/d)$$

$$E_z = 0$$

where: $G = a^2 + b^2$ and

$e^{j\omega t}$ time dependence is assumed.

A.2 Field Equations for Cylindrical Disk DRA [43]

(See Figure 3.2)

$TE_{01\delta}^z$ Mode

$$H_z = H_0 J_0(\beta\rho) \cos(\delta\pi z/2h)$$

$$H_\rho = [-\delta\pi H_0 / (\beta L)] J_0'(\beta\rho) \sin(\delta\pi z/2h)$$

$$E_\phi = (j\omega\mu_0 H_0 / \beta) J_0'(\beta\rho) \cos(\delta\pi z/2h)$$

$$H_\phi = E_\rho = E_z = 0$$

where: β satisfies $J_0(\beta r) = 0$, r = radius of disk

and $e^{j\omega t}$ time dependence is assumed.

$TM_{01\delta}^z$ Mode (Note $TM_{01\delta}$ is similar to $TE_{10\delta}$ with H replaced by E)

$$E_z = E_0 J_0(\beta\rho) \cos(\delta\pi z/2h)$$

$$H_\phi = -j\beta / (\omega\mu_0) E_0 J_0'(\beta\rho) \cos(\delta\pi z/2h)$$

$TE_{11\delta}^z$ Mode

$$E_z = 0$$

$$E_\rho = j\omega\mu_0 / (\rho\beta^2) H_0 J_1(\beta\rho) \sin\phi \sin(\delta\pi z/2h)$$

$$E_\phi = j\omega\mu_0 / \beta^2 H_0 J_1'(\beta\rho) \cos\phi \sin(\delta\pi z/2h)$$

$$H_z = H_0 J_1(\beta\rho) \cos\phi \sin(\delta\pi z/2h)$$

$$H_\phi = \delta\pi / (\beta 2h) H_0 J_1'(\beta\rho) \cos\phi \cos(\delta\pi z/2h)$$

$$H_\rho = \pi\delta / (\beta 2h) H_0 J_1'(\beta\rho) \cos\phi \cos(\delta\pi z/2h)$$

A3. Field Equations for Rectangular Dielectric Waveguide (See Figure 3.4)

Omitting the dependence $\exp(j(\omega t - k_z z))$, the main tranverse field components of the E_{pq}^y modes are E_y and H_x and are given below [39].

Region 1: ($|x| \leq a, y < b$)

$$E_y = E_0 \cos(k_x x) \cos(k_y y)$$

Regions 2 and 4: ($|x| > a, y < b$)

$$E_y = E_0 \cos(k_x a) \cos(k_y y) \exp[-k_{xo}(|x| - a)]$$

Regions 3 and 5: ($|x| < a, y > b$)

$$E_y = E_0 \cos(k_x x) \cos(k_y b) \exp[-k_{yo}(y - b)]$$

$$H_x = -E_y / \eta$$

where:

$$\eta = (\mu_0 / \epsilon_0)^{0.5} (k_0 / k_z);$$

k_x, k_y, k_{xo}, k_{yo} are the propagation constants in the x and y direction in the dielectric and air, respectively; and

k_z is the propagation constant in the z direction.

Appendix B

B.1 Diffraction Coefficient Equations

The diffraction coefficients for hard and soft polarization are listed below, with variables defined in Figure B.1 [83]

Soft (Horizontal) Polarization: Dirichlet Boundary Condition ($E_z=0$ on the faces of the wedge)

$$D_s(\rho', \beta^-, n) = D^i(\rho', \beta^-, n) - D^r(\rho', \beta^+, n)$$

Hard (Vertical) Polarization: Neumann Boundary Condition ($H_z=0$ on the faces of the wedge)

$$D_h(\rho', \beta^-, n) = D^i(\rho', \beta^-, n) + D^r(\rho', \beta^+, n)$$

$$D^i(\rho', \beta^-, n) = -\frac{e^{-j\pi/4}}{2n\sqrt{2\pi k}} \cdot \left\{ \begin{array}{l} C^+(\beta^-, n)F[k\rho' g^+(\beta^-)] + \\ C^-(\beta^-, n)F[k\rho' g^-(\beta^-)] \end{array} \right\}$$

$$D^r(\rho', \beta^+, n) = -\frac{e^{-j\pi/4}}{2n\sqrt{2\pi k}} \cdot \left\{ \begin{array}{l} C^+(\beta^+, n)F[k\rho' g^+(\beta^+)] + \\ C^-(\beta^+, n)F[k\rho' g^-(\beta^+)] \end{array} \right\}$$

where:

$$C^+(\beta, n) = \cot\left(\frac{\pi + \beta}{2n}\right) \quad C^-(\beta, n) = \cot\left(\frac{\pi - \beta}{2n}\right)$$

$$F(x) = 2j\sqrt{x}e^{jx} \int_x^\infty e^{-j\tau^2} d\tau$$

$$\beta^+ = \phi + \phi' \quad \beta^- = \phi - \phi'$$

$$g^+ = 1 + \cos[(\phi \pm \phi') - 2n\pi N^+] \quad g^- = 1 + \cos[(\phi \pm \phi') - 2n\pi N^-]$$

$$2n\pi N^+ - (\phi \pm \phi') = \pi \quad 2n\pi N^- - (\phi \pm \phi') = -\pi$$

Figure B.1 Diffraction from a straight wedge, from [83]

B.2 Computer Listings for GTD

```
/*-----
```

GTD Computation

Calculates the GTD pattern of a magnetic dipole on the surface of a rectangular ground plane.

- calls cf subroutine which calculates diffraction coefficients
- cf in turn calls fresnel which calculates the required fresnel integral

Inputs:

C1, C2: Values of incident $0.5 \cdot E_{go}$ field (at +/- 90 degrees)
 Frequency: frequency of operation (GHz)
 D1, D2: distances from edge 1 and edge 2 (meters)
 Ego[]: vector containing the GO field

Outputs:

Et[]: vector containing the total field ($E_{go} + E_{d1} + E_{d2}$)

```
-----*/
```

```
#include "math.h"
#include "complex.h"
#include "fresnel.h"
```

```
void compute_gtd(double C1, double C2, float Frequency, float D1,
                 float D2, float Ego[], double Et[])
```

```
{
    double D_Pi;
    float Fn; // Fn = 2 for a plane diffraction
    float Wv, K;
    double Theta, Angle;
    double C3, C4; //C1, C2 are the values of the GO field at the horizon
    complex Ed1[360], Ed2[360];
    complex Di, J, Arg;
    float Fmax;
    int i;

    D_Pi = acos(-1.0);
    Wv = 0.3 / Frequency; // wavelength (m)
    K = 2 * D_Pi / Wv;
```

```

J = complex(0.0, 1.0);
Fn = 2;           // angle for a plane wedge
D1 = D1 / Wv;
D2 = D2 / Wv;

// Compute Diffracted Fields
C3 = -1 / (sqrt(D1) * 2.0 * Fn * D_Pi);
C4 = -1 / (sqrt(D2) * 2.0 * Fn * D_Pi);
for (i = -180; i <= 180; i++)
{
    Theta = i * D_Pi / 180;

// Diffracted Fields at edge 1
    if (i < 90) Angle = Theta + D_Pi / 2.0;
    else       Angle = Theta - 3.0 * D_Pi / 2.0;
    Di = cf(D1, Angle, Fn);
    Arg = 2 * D_Pi * (D1 - D1 * sin(Theta)) + D_Pi / 4.0;
    Ed1[i + 180] = C1 * C3 * Di * exp(-J * Arg);

// Diffracted Fields at edge 2
    if ((i >= -90)) Angle = D_Pi / 2.0 - Theta;
    else           Angle = 5.0 * D_Pi / 2.0 - Theta;
    Di = cf(D2, Angle, Fn);
    Arg = 2 * D_Pi * (D2 + D2 * sin(Theta)) + D_Pi / 4.0;
    Ed2[i + 180] = C2 * C4 * Di * exp(-J * Arg);
}

// Total Field
Fmax = 0;
for (i = 0; i < 360; i++) Et[i] = sqrt(norm(Ego[i] + Ed1[i] + Ed2[i]));
for (i = 0; i < 360; i++) if (Fmax < Et[i]) Fmax = Et[i];
for (i = 0; i < 360; i++) Et[i] = Et[i] / Fmax;
D1 = D1 * Wv;
D2 = D2 * Wv;
}

```



```

/*-----
function: cf

Purpose: this function calculates the diffraction coefficients
         for a straight wedge
Based on FORTRAN subroutine found in [90]
Inputs:
    R:      distance from source to diffraction edge
    Angle:   observation angle, measured from edge
    Fn:      wedge number (FN = 2 for a plane)

Outputs:
    the function returns the complex diffraction coefficient
-----*/

```

```

complex cf(double R, double Angle, double Fn)
{
    int N;
    double Rcf, Ucf, Rag, D_Pi, Ts, LowLimit, A, Comp, D_N;
    complex Answer, Top, Exp, Upi, Unpi, J, Fresnel;

    D_Pi = acos(-1.0);
    J = complex(0.0, 1.0);
    Top = J * 2.0 * sqrt(2.0 * D_Pi * R);
    N = floor((D_Pi + Angle) / (2.0 * Fn * D_Pi) + 0.5);
    D_N = (float) N;
    A = 1.0 + cos(Angle - 2.0 * Fn * D_Pi * D_N);
    LowLimit = sqrt(2.0 * D_Pi * R * A);
    Exp = exp(J * 2.0 * D_Pi * R * A);
    Fresnel = fresnel(LowLimit);
    Rag = (D_Pi + Angle) / (2.0 * Fn);
    Ts = fabs(sin(Rag)); // Fortran SNGL function Double -> Single
    if (Ts > pow(10.0, -5.0))
    {
        Comp = sqrt(A) * cos(Rag) / sin(Rag);
    }
    else
    {
        Comp = -1.0 * sqrt(2.0) * Fn * sin(Angle / 2.0 - Fn * D_Pi * D_N);
        if (cos(Angle / 2.0 - Fn * D_Pi * D_N) < 0.0)
        {
            Comp = -Comp;
        }
    }
}

```

```

    }
}
Uppi = Top * Exp * Comp * Fresnel;
N = floor((-D_Pi + Angle) / (2.0 * Fn * D_Pi) + 0.5);
D_N = (float) N;
A = 1.0 + cos(Angle - 2.0 * Fn * D_Pi * D_N);
LowLimit = sqrt(2.0 * D_Pi * R * A);
Exp = exp(J * 2.0 * D_Pi * R * A);
Fresnel = fresnel(LowLimit);
Rag = (D_Pi - Angle) / (2.0 * Fn);
Ts = fabs(sin(Rag));
if (Ts > pow(10.0, -5.0))
{
    Comp = sqrt(A) * cos(Rag) / sin(Rag);
}
else
{
    Comp = sqrt(2.0) * Fn * sin(Angle / 2.0 - Fn * D_Pi * N);
    if (cos(Angle / 2.0 - Fn * D_Pi * N) <= 0.0)
    {
        Comp = -Comp;
    }
}

Unpi = Top * Exp * Comp * Fresnel;
Rcf = real(Uppi + Unpi);
Ucf = imag(Uppi + Unpi);
Answer = complex(Rcf, Ucf);
return (Answer);
}

```

```
/*-----
function: fresnel
```

Purpose: this function calculates the Fresnel integral required by
function cf

Based on "Computation of Fresnel Integrals", J. Boersma, J.Math.
Comp., vol. 14, p.380, 1960

Inputs:

Lowlimit: lower limit of Fresnel integration

Outputs:

the function returns the complex Fresnel integral

```
----- */
```

```
complex fresnel(double LowLimit)
```

```
{
    complex Answer;
    double A[12], B[12], C[12], D[12];
    double X, Y, Fr, Fi, Co, So, D_Pi;
    int i;
```

```
// Initialize Constants
```

```
D_Pi = acos(-1.0);
```

A[1] = 1.595769140;	B[1] = -0.000000033;
A[2] = -0.000001702;	B[2] = 4.255387524;
A[3] = -6.808568854;	B[3] = -0.000092810;
A[4] = -0.000576361;	B[4] = -7.780020400;
A[5] = 6.920691902;	B[5] = -0.009520895;
A[6] = -0.016898657;	B[6] = 5.075161298;
A[7] = -3.050485660;	B[7] = -0.138341947;
A[8] = -0.075752419;	B[8] = -1.363729124;
A[9] = 0.850663781;	B[9] = -0.403349276;
A[10] = -0.025639041;	B[10] = 0.702222016;
A[11] = -0.150230960;	B[11] = -0.216195929;
A[12] = 0.034404779;	B[12] = 0.019547031;

C[1] = 0.0;	D[1] = 0.199471140;
C[2] = -0.024933975;	D[2] = 0.000000023;
C[3] = 0.000003936;	D[3] = -0.009351341;
C[4] = 0.005770956;	D[4] = 0.000023006;
C[5] = 0.000689892;	D[5] = 0.004851466;
C[6] = -0.009497136;	D[6] = 0.001903218;
C[7] = 0.011948809;	D[7] = -0.017122914;
C[8] = -0.006748873;	D[8] = 0.029064067;
C[9] = 0.000246420;	D[9] = -0.027928955;
C[10] = 0.002102967;	D[10] = 0.016497308;
C[11] = -0.001217930;	D[11] = -0.005598515;
C[12] = 0.000233939;	D[12] = 0.000838386;

// Calculate Fresnel Integral from LowLimit to Infinity

```

if (LowLimit <= 0.0)
{
    Co = So = -0.0;
}
else
{
    X = LowLimit * LowLimit;
    Fr = Fi = 0.0;
    if (X < 4.0)
    {
        Y = X / 4.0;
        for (i = 12; i >= 2; i--)
        {
            Fr = (Fr + A[i]) * Y;
            Fi = (Fi + B[i]) * Y;
        }
        Fr += A[1];
        Fi += B[1];
        Co = (Fr * cos(X) + Fi * sin(X)) * sqrt(Y);
        So = (Fr * sin(X) - Fi * cos(X)) * sqrt(Y);
    }
}

```

```

else
{
    Y = 4.0 / X;
    for (i = 12; i >= 2; i--)
    {
        Fr = (Fr + C[i]) * Y;
        Fi = (Fi + D[i]) * Y;
    }
    Fr += C[1];
    Fi += D[1];
    Co = 0.5 + (Fr * cos(X) + Fi * sin(X)) * sqrt(Y);
    So = 0.5 + (Fr * sin(X) - Fi * cos(X)) * sqrt(Y);
}
Co = sqrt(D_Pi / 2.0) * (0.5 - Co);
So = sqrt(D_Pi / 2.0) * (So - 0.5);
Answer = complex(Co, So);

return (Answer);
}

```

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