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Techniques for Pattern Control of a Dielectric Rod Antenna Suitable for use in Mobile Communications

By

Gavin James Cox MEng AMIEE

A Doctoral Thesis

Submitted in partial fulfilment of the requirements for the award of Doctor of Philosophy by Loughborough University

September 2002

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To the ones I love
Abstract

This thesis describes the development of antennas suitable for mobile communication systems based on a dielectric rod antenna fed from circular waveguide. Pattern control of the antenna is implemented using a combination of Frequency Selective Surface (FSS) elements and metallic endcaps placed on the antenna. Both linear and circular polarised feeds have been made for these antennas to ensure they are suitable for a wide range of applications. The suitability of the dominant and next, higher order, waveguide mode were investigated and conclusions drawn as to their suitability for this type of antenna. The antennas were extensively modelled using a commercial TLM based solver and the results of these simulations were compared to the comprehensive set of antenna pattern measurements and S-parameter measurements obtained for the prototype antennas.

Keywords
Dielectric Rod Antenna
Frequency Selective Surface (FSS)
Leaky Wave Antenna
Mobile Communications
Transmission Line Matrix (TLM) Modelling
Waveguide Antenna
Publications arising directly from this research

Conferences


Acknowledgements

I would like to thank my supervisors Rob Seager and Yiannis Vardaxoglou for their help and support. Rob, I have always felt that I've been able to take any problem to you, even if it was not work related.

I must thank James Eade for inspiring me to consider a PhD. If it were not for his guidance through out my final year project, I would not find myself in the position I am now – Thanks James, and yes I should have started writing sooner!

The people I have worked with in the department have always been helpful and more than willing to give advice when I have asked for it. This is especially true of the other researchers in the Wireless Communications group and CMCR. Alex, Alford, Chin, George, Nick, Alistair, Mohan and Patrick you have been a great help when help has been needed.

Finally, but not least, to Debbe and my family. You have been there throughout. Without your love, help, support and encouragement I would not have got this far. The gratitude and thanks that I owe you is, I hope, reflected in this thesis.
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Abbreviations and Symbols

1G  1\textsuperscript{st} Generation
2G  2\textsuperscript{nd} Generation
3G  3\textsuperscript{rd} Generation

a Radius

ABC Absorbing Boundary Condition

\( B \) Magnetic Flux Density (vector quantity if denoted \( \vec{B} \) and direction, or function of, if denoted with a subscript)

CEM Computational Electromagnetics

CP Circular Polarisation

dB Decibel (referenced to 1 milliwatt if dBm)

D Periodicity or Electric Flux Density (vector quantity if denoted \( \vec{D} \) and direction, or function of, if denoted with a subscript)

DR Dielectric Resonator

e Natural logarithm

\( E \) Electric field (vector quantity if denoted \( \vec{E} \) and direction, or function of, if denoted with a subscript)

\( f \) Frequency (cut off frequency if denoted \( f_c \))

\( F_R \) Resonant Frequency

\( F_T \) Passband Frequency

FD Finite Difference or Frequency Domain

FDIE Frequency Domain Integral Equation

FDDE Frequency Domain Differential Equation

FE Finite Element

FSG Frequency Selective Guide

FSS Frequency Selective Surface

G Shunt Admittance

GHz Giga-Hertz

GPS Global Positioning System

GSM Global System for Mobile communication

GTD Geometric Theory of Diffraction
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Definition</th>
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<tr>
<td>(H)</td>
<td>Magnetic field (vector quantity if denoted (\overline{H}) and direction, or function of, if denoted with a subscript)</td>
</tr>
<tr>
<td>i</td>
<td>Current</td>
</tr>
<tr>
<td>(j)</td>
<td>(\sqrt{-1}) or Current Density</td>
</tr>
<tr>
<td>(J_n)</td>
<td>Bessel Function of the first kind</td>
</tr>
<tr>
<td>(J'_n)</td>
<td>Derivative of (J_n)</td>
</tr>
<tr>
<td>(k)</td>
<td>Wave number (= \omega \sqrt{\mu / \varepsilon} )</td>
</tr>
<tr>
<td>(k_c)</td>
<td>Cutoff wave number</td>
</tr>
<tr>
<td>L</td>
<td>Inductance (or length of FSS element)</td>
</tr>
<tr>
<td>LWA</td>
<td>Leaky Wave Antenna</td>
</tr>
<tr>
<td>m</td>
<td>Integer number or metre</td>
</tr>
<tr>
<td>mm</td>
<td>Millimetre</td>
</tr>
<tr>
<td>(\mu m)</td>
<td>Micrometre</td>
</tr>
<tr>
<td>Mbits</td>
<td>Mega bits</td>
</tr>
<tr>
<td>MOM</td>
<td>Method of Moments</td>
</tr>
<tr>
<td>n</td>
<td>Integer number</td>
</tr>
<tr>
<td>(p_{mn})</td>
<td>(m^{th}) route of (J_n)</td>
</tr>
<tr>
<td>(p'_{mn})</td>
<td>(m^{th}) route of (J'_n)</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
</tr>
<tr>
<td>PEC</td>
<td>Perfect Electric Conductor</td>
</tr>
<tr>
<td>PML</td>
<td>Perfectly Matched Layer</td>
</tr>
<tr>
<td>PTFE</td>
<td>Polytetrafluoroethylene</td>
</tr>
<tr>
<td>R</td>
<td>Series Resistance</td>
</tr>
<tr>
<td>R_s</td>
<td>Surface Resistance</td>
</tr>
<tr>
<td>s</td>
<td>Seconds</td>
</tr>
<tr>
<td>(S_{11})</td>
<td>Forward Reflection component of a Scattering matrix</td>
</tr>
<tr>
<td>(S_{21})</td>
<td>Forward Transmission component of a Scattering matrix</td>
</tr>
<tr>
<td>SCN</td>
<td>Symmetrical Condensed Node</td>
</tr>
<tr>
<td>SMA</td>
<td>Subminiature version A</td>
</tr>
<tr>
<td>t</td>
<td>Time</td>
</tr>
<tr>
<td>(\tan \delta)</td>
<td>Loss Tangent</td>
</tr>
<tr>
<td>TD</td>
<td>Time Domain</td>
</tr>
<tr>
<td>TDDE</td>
<td>Time Domain Differential Equation</td>
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</table>
TDIE  Time Domain Integral Equation  
TE   Transverse Electric  
TEM  Transverse Electromagnetic  
TLM  Transmission Line Modelling  
TM   Transverse Magnetic  
UMTS Universal Mobile Telecommunications system  
V    Volt  
W    Width of a FSS element  
Z    Impedance  

°  Degree  
∇  Vector operator  
α  Attenuation constant (conduction loss if denoted αc or dielectric loss if denoted αd)  
β  Propagation Constant  
Δ  'Change of'  
∂  Partial Derivative  
ε  Permittivity (relative if denoted εr or free space if denoted ε0)  
ϕ  Angle of rotation from the x-axis or radiation direction from a FSG  
γ  Complex propagation constant  
η  Impedance of free space  
λ  Wavelength (Guide wavelength if denoted λg or resonant wavelength if denoted λr)  
μ  Permeability (relative if denoted μr or free space if denoted μ0)  
π  3.147...  
ρ  Distance from waveguide centre  
σ  Conductivity  
Ω  Ohm  
ω  Angular Frequency ( = 2πf )
1 Introduction

1.1 Overview

The use of mobile and satellite communications and the services that are offered on the handset receivers has increased hugely in recent years. It has been reported in the press that as many as one in three people now owns a mobile telephone using the current Global System for Mobile communications (GSM) technology, also known as the 2nd generation (2G) of mobile telephony. This figure is predicted to rise to one in two in only a few years time. While a lot of the technologies being incorporated into mobile telephones utilise advances in the power of embedded devices the antenna, which could be argued is the most important feature of the handset, has to keep pace with these advances. Increasing demands have been placed on the specifications of the antenna such as frequencies of operation, bandwidth and specialised radiation patterns since the 1st generation (1G) of devices in the 1980s. Antennas for the next generation of handsets can have up to 3 bands of operation and, as multimedia services become available, the ability to cope with data rates of up to 2Mbits/s are required. A Universal Mobile Telecommunications System (UMTS) task force has suggested that within 10 years of the introduction of UMTS technology, known as the 3rd generation (3G), there could be 60 million users worldwide. This projected demand creates a large potential market for new technologies and, in particular, new antenna designs. A summary of the main services available at mobile frequencies is shown in Figure 1-1, and any special considerations are noted. Jones et al. and Richardson give a good introduction to and overview of the development of mobile communications and the challenges that are faced.
Different types of application have different requirements in terms of radiated antenna pattern shape. A satellite antenna, e.g. a Global Positioning System (GPS) receiver, requires a cardioid pattern as the optimum shape for maximum power to be received from orbiting satellites. The pattern most suited to a mobile telephone handset antenna is typical of a dipole e.g. radiating uniformly in all directions in the horizontal plane. The challenge is to incorporate these two types of antenna requirement into a single antenna, not just operating at the correct frequencies, but also generating the optimum antenna patterns for the respective technologies.

As handsets become smaller so the need for compact antenna designs increases. New antenna materials, which exhibit high dielectric constants, are available and can be used to decrease the physical size of the antenna using scaling techniques. As these pose more problems to feed than the mono-pole antennas traditional used, due to the difficulty of coupling energy directly into them, alternative feeding technologies are required. Use is made of existing feeding systems to achieve a solution to this engineering problem, especially where a circularly polarised field is required.

Much work has been done on Frequency Selective Surfaces (FSSs) and the leaky wave action they possess has been studied. This work aims to utilise characteristics that FSSs possess and combine them with an existing antenna technology to produce an antenna capable of working with mobile and satellite technologies. In simplistic
form this will be achieved by filling a circular waveguide with a dielectric filling extending from it as a basis for an antenna, illustrated in Figure 1-2, and investigating the beam shaping effect caused by FSS surface configurations. With suitable FSS design it is hoped that multiband antennas can be realised. At mobile frequencies it is hoped that the FSS’s leaky wave action can be used to broaden the radiated beam whilst at satellite frequencies the beam could radiate in a more vertical direction. The use of an endcap on the antenna can further enhance the beam shape, by allowing more radiation to be ‘pushed’ sidewards, and different types of endcap are investigated. These types of structure are fully explored in this thesis.

Figure 1-2 A generic diagram of the proposed antenna design

In terms of operation, the final waveguide has been designed to work at a centre frequency of 1.8GHz to place it within the centre of the mobile communications frequency spectrum. This allows GSM 1800 frequency bands to be covered and also the low frequency satellite part of the UMTS spectrum. Obviously, other services could be accommodated easily by changing the size of the waveguide, or by using a different dielectric material for the rod part of the antenna which would vary the frequency of operation from the one chosen for this study. Beam shaping could be achieved at any frequency above the cut off frequency of the waveguide by choosing the FSS operating frequency accordingly.

1.1.1 Relative Frequencies, their applications and CP Characteristics

Care has to be taken with the choice of frequencies used within the waveguide. The first two modes present are the $TE_{11}$ and $TM_{01}$. Due to the difficulties in obtaining
circular polarisation of the TM\textsubscript{01} mode, due to the nature of the field structure, any antenna designed for satellite use would have to work in the mono-mode region e.g. using the TE\textsubscript{11} mode. There is nothing to prevent other, linearly polarised, signals being received or transmitted using the waveguide working at frequencies where higher modes are present e.g. in the TM\textsubscript{01} region. The nature of the fields that cause CP problems, e.g. in the TM\textsubscript{01} mode, is discussed in further detail in 2.1.1.

1.2 Thesis Layout and Overview

Antennas based on dielectric rods have been successfully used in the majority of handset antennas since the 2\textsuperscript{nd} generation of phones. The remaining antennas in current use are usually based around the patch antenna having moved away from early monopole designs. The aim of this work is to examine the dielectric rod antenna and study possible methods of beam shaping that allow its use in technologies that require different antenna patterns. The thesis is laid out in the following manner:

Chapter 2 discusses the necessary theory required to understand the concepts used for this type of antenna. The main topics of discussion are waveguide theory, Transmission Line Modelling (TLM), Frequency Selective Surfaces (FSSs) and dielectric Rod antennas.

Chapter 3 discusses the possible feeding techniques that are suited to this type of antenna. Coaxial cable, microstrip slot coupling and coaxial cable driven microstrip slots are investigated. The ease with which these can be used to generate a circularly polarised field is considered.

Chapter 4 details the development of a prototype design complete with frequency selective surface elements with a working frequency of 9GHz. Elements consisting of dipoles, crossed-dipoles and square loops were investigated to see which would be most suitable for control of the rod's radiation pattern. The procedure used to measure the antenna is described.

Chapter 5 details the work carried out on the antenna designed to work at 1.8GHz. The design procedure is described, based on the work presented in the preceding
chapters, and a full set of measurements for the final FSS configuration used is presented.

Chapter 6 concludes the work, summarising the results and salient points for the feeds and antennas investigated. Emphasis is placed on the leaky wave action that the FSS exhibits. Results are concluded for individual sections and general conclusions applicable to all of the work undertaken are also presented. Suggestions for future work are listed.

References

1 UMTS Task Force Report, February 1996.


2 Theory of Guided Waves and FSS/FSG Design

In 1897 Lord Rayleigh\(^1\) proved mathematically that the idea of propagation inside hollow metal pipes was possible though it was not until 1936 that experimental results were published. Waveguides have since been used in numerous applications at higher frequencies, typically above 1GHz, because of their low loss and high power handling capabilities.

2.1 Waveguide Theory

Circular waveguides can not support the TEM mode that multiple conductors, such as coaxial cable are able to do since, in a cross-section of a guide, the area is enclosed by a conducting wall. This means that the electrostatic field within this region must be zero. Instead they support Transverse Electric (TE) and Transverse Magnetic (TM) modes. The geometry used to describe a waveguide of radius \(a\) is shown in Figure 2-1. Any point inside the waveguide can be described as a distance, \(\rho\), rotated from the x-axis by an angle, \(\phi\).

![Figure 2-1 Overview of the circular waveguide geometry](image)

2.1.1 Mode Structure

Expanding Maxwell’s equations of (2.1) and (2.2) in cylindrical co-ordinates and solving accordingly leads to equations (2.3) to (2.8) for TE modes and (2.9) to (2.14) for TM modes. The full derivation is shown in Appendix A Derivation of Circular Waveguide Equations

\[
\nabla \times \bar{E} = -j\omega \mu \bar{H} \tag{2.1}
\]
For TE modes:

\[\nabla \times \mathbf{H} = j\omega \mathbf{E}\]

(2.2)

For TE modes:

\[E_z = 0\]

(2.3)

\[H_z = (A \sin n\phi + B \cos n\phi)J_n(k_c \rho)e^{-j\beta z}\]

(2.4)

\[E_\rho = -\frac{j\omega \mu}{k_c^2 \rho} (A \cos n\phi - B \sin n\phi)J_n(k_c \rho)e^{-j\beta z}\]

(2.5)

\[E_\phi = \frac{j\omega \mu}{k_c} (A \sin n\phi + B \cos n\phi)J'_n(k_c \rho)e^{-j\beta z}\]

(2.6)

\[H_\rho = -\frac{j\beta}{k_c} (A \sin n\phi + B \cos n\phi)J'_n(k_c \rho)e^{-j\beta z}\]

(2.7)

\[H_\phi = -\frac{j\beta n}{k_c^2 \rho} (A \cos n\phi - B \sin n\phi)J_n(k_c \rho)e^{-j\beta z}\]

(2.8)

And for TM modes:

\[E_z = (A \sin n\phi + B \cos n\phi)J_n(k_c \rho)e^{-j\beta z}\]

(2.9)

\[H_z = 0\]

(2.10)

\[E_\rho = -\frac{j\beta}{k_c} (A \sin n\phi - B \cos n\phi)J'_n(k_c \rho)e^{-j\beta z}\]

(2.11)

\[E_\phi = -\frac{j\beta n}{k_c^2 \rho} (A \cos n\phi - B \sin n\phi)J_n(k_c \rho)e^{-j\beta z}\]

(2.12)

\[H_\rho = \frac{j\omega \epsilon_n}{k_c^2 \rho} (A \cos n\phi - B \sin n\phi)J'_n(k_c \rho)e^{-j\beta z}\]

(2.13)

\[H_\phi = -\frac{j\omega \epsilon}{k_c} (A \sin n\phi + B \cos n\phi)J_n(k_c \rho)e^{-j\beta z}\]

(2.14)

Where \(J_n(k_c \rho)\) and \(J'_n(k_c \rho)\) are Bessel functions of the first kind and the derivative of \(J_n\) with respect to its argument, respectively.

These equations lead to the field configurations within a circular waveguide and a comprehensive set of these is illustrated by Balanis. Of most importance to this work are the dominant mode, TE_{11}, and the next order mode TM_{01}. The field configurations for these two modes are illustrated below in Figure 2-2.
Figure 2-2 Field configurations of the TE\textsubscript{11} and TM\textsubscript{01} circular waveguide modes

2.1.1.1 Waveguide Cut off Frequency

Referring to Appendix A Derivation of Circular Waveguide Equations, the term $k_e$ has been shown to have only one value for a given mode and waveguide diameter. This leads to the value of $\beta$ having a range of values described in section 2.1.1.1.1 and 2.1.1.1.2.

2.1.1.1.1 TE Case

For TE modes $\beta$ can have the range of values determined by (2.15), where $k$ is a constant, defined as $\omega \sqrt{\mu \varepsilon}$, and is known as the wave number or propagation constant of the medium.

$$\beta = \sqrt{k^2 - k_e^2} = \sqrt{k^2 - \left(\frac{p'_{nm}}{a}\right)^2} \text{ when } k > k_e = \frac{p'_{nm}}{a}$$

$$\beta = 0 \text{ when } k = k_e = \frac{p'_{nm}}{a}$$

$$\beta = -j \sqrt{k_e^2 - k^2} = -j \sqrt{\left(\frac{p'_{nm}}{a}\right)^2 - k^2} \text{ when } k < k_e = \frac{p'_{nm}}{a}$$

(2.15)

The cut off frequency is when $\beta = 0$ so, from (2.15):
Where \( p'_{nm} \) is the \( m^{th} \) root of \( J_n(x) \) such that \( J_n(p'_{nm}) = 0 \) and \( p'_{11} = 1.8412 \)

This leads to (2.15) being rewritten as:

\[
\beta = \sqrt{k^2 - k_c^2} = \beta \sqrt{1 - \left(\frac{k_c}{k}\right)^2} = \beta \sqrt{1 - \left(\frac{p'_{nm}}{ka}\right)^2} = \beta \sqrt{1 - \left(\frac{f_c}{f}\right)^2}
\]
when \( f > f_c \)

\[
\beta = 0 \quad \text{when } f = f_c
\]

\[
\beta = -j\sqrt{k_c^2 - k^2} = -jk\sqrt{\left(\frac{k_c}{k}\right)^2 - 1} = -jk\sqrt{\left(\frac{p'_{nm}}{ka}\right)^2 - 1} = -jk\sqrt{\left(\frac{f_c}{f}\right)^2 - 1}
\]
when \( f < f_c \)

The guide wavelength, \( \lambda_g \), is defined as:

\[
\lambda_g = \frac{2\pi}{\beta}
\]

Rewriting (2.17) the guide wavelength can be shown as:

\[
\lambda_g = \frac{\frac{2\pi}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}}}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}} \quad \text{when } f > f_c
\]

\[
\lambda_g = \infty \quad \text{when } f = f_c
\]

Guide wavelength for a given mode does not exist below the cut off frequency, as signified by the \( j \) term present in (2.19). The wave is, therefore, evanescent in nature e.g. decaying exponentially, and unlikely to propagate.

2.1.1.1.2 TM Case

For TM modes \( \beta \) can have the range of values determined by (2.20).

\[
\beta = \sqrt{k^2 - k_c^2} = \sqrt{k^2 - \left(\frac{p_{nm}}{a}\right)^2} \quad \text{when } k > k_c = \frac{p_{nm}}{a}
\]

\[
\beta = 0 \quad \text{when } k = k_c = \frac{p_{nm}}{a}
\]

\[
\beta = -j\sqrt{k_c^2 - k^2} = -jk\sqrt{\left(\frac{p_{nm}}{a}\right)^2 - k^2} \quad \text{when } k < k_c = \frac{p_{nm}}{a}
\]

The cut off frequency is when \( \beta = 0 \) so, from (2.20):

\[
f_c = \frac{p_{nm}}{2\pi a \sqrt{\mu \varepsilon}}
\]

Where \( p_{nm} \) is the \( m^{th} \) root of \( J_n(x) \) such that \( J_n(p_{nm}) = 0 \) and \( p_{11} = 2.4049 \)
This leads to (2.20) being rewritten as:

\[ \beta = \sqrt{k^2 - k_c^2} = \beta \sqrt{1 - \left(\frac{k_c}{k}\right)^2} = \beta \sqrt{1 - \left(\frac{f_c}{f}\right)^2} \quad \text{when } f > f_c \]

\[ \beta = 0 \quad \text{when } f = f_c \]

\[ \beta = -j \sqrt{k_r^2 - k_c^2} = -jk \sqrt{\left(\frac{k_c}{k}\right)^2 - 1} = -jk \sqrt{\left(\frac{f_c}{f}\right)^2 - 1} \quad \text{when } f < f_c \]

The guide wavelength, \( \lambda_g \), is defined as:

\[ \lambda_g = \frac{2\pi}{\beta} \quad (2.23) \]

Rewriting (2.22) the guide wavelength can be shown as:

\[ \lambda_g = \frac{2\pi}{\beta \sqrt{1 - \left(\frac{f_c}{f}\right)^2}} = \frac{\lambda}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}} \quad \text{when } f > f_c \]

\[ \lambda_g = \infty \quad \text{when } f = f_c \quad (2.24) \]

Guide wavelength for a given mode does not exist below the cut off frequency, as signified by the \( j \) term present in (2.24). The wave is, therefore, evanescent in nature, e.g. decaying exponentially, and so non-propagating.

### 2.1.1.2 Waveguide Impedance

The propagation of a wave inside a waveguide will be subject to a frequency dependant value of impedance. The value of this impedance can be stated in several ways, one being:

\[ Z = \frac{E_y}{H_z} \quad (2.25) \]

This is equal to:

\[ Z_{\tau\flat} = \frac{\omega \mu}{\beta} = \frac{\eta k}{\beta} \quad (2.26) \]

Where \( \eta = \sqrt{\frac{\mu}{\varepsilon}} \)

Which, when rewritten taking into account the possible values of \( \beta \) becomes:
Equations in (2.27) allow the following conclusions to be drawn about the nature of the waveguide impedance for TE modes. Above the cut off frequency the impedance is real and greater than the intrinsic impedance of the material inside the waveguide. At cut off the impedance becomes infinite. Below cut off the impedance is imaginary and inductive so the waveguide is able to store energy.

For the TM case the impedance can be expressed as:

\[ Z_{TM} = \frac{\beta}{\omega \varepsilon} = \frac{\eta \beta}{k} \quad (2.28) \]

Where \( \eta = \sqrt{\frac{\mu}{\varepsilon}} \)

Which, when rewritten taking into account the possible values of \( \beta \) becomes:

\[ Z_{TM} = \frac{\beta}{\omega \varepsilon} \sqrt{1 - \left(\frac{f_c}{f}\right)^2} = \frac{\mu}{\varepsilon} \sqrt{1 - \left(\frac{f_c}{f}\right)^2} = \eta \sqrt{1 - \left(\frac{f_c}{f}\right)^2} \quad \text{when } f > f_c \]

\[ Z_{TM} = 0 \quad \text{when } f = f_c \quad (2.29) \]

Equations in (2.29) allow the following conclusions to be drawn about the nature of the waveguide impedance for TM modes. Above the cut off frequency the impedance is real and less than the intrinsic impedance of the material inside the waveguide. At cut off the impedance is zero. Below cut off the impedance is imaginary and capacitive so the waveguide is able to store energy.
2.1.1.3 Bandwidth

Bandwidth of the antenna is an important factor, since the applications that this antenna is intended for have strict bandwidth criteria. It has been shown in sections 2.1.1.1.1 and 2.1.1.1.2 that the values that satisfy Bessel's equations are 1.8412 and 2.4049 respectively for the TE and TM case. Dividing the result of the TM case by the result of the TE case leads to a figure of 1.306. This figure is, given the lower cut off frequency, the frequency at which mono-mode operation will cease and the higher mode start to propagate. Expressed as a percentage it can be shown that the percentage bandwidth available within the mono-mode region of operation is:

\[
\% \text{ Bandwidth} = \frac{\Delta f}{f_{\text{centre}}} \times 100 = 26.5\% 
\]  

(2.30)

Where \( \Delta f \) is the difference between the TM01 and TE11 cut off frequencies and \( f_{\text{centre}} \) is the centre frequency within the mono-mode region.

This value for the maximum available bandwidth, given the current specifications\(^3\), is suitable for the intended application. Circular waveguide only has one degree of freedom, \( a \), to vary in the design of the waveguide and, as such, design options are more limited than with its rectangular equivalent. It is likely that the effect of the resonators will further reduce the available bandwidth.

2.1.1.4 Mode Orientation

Given the nature of the field in the TE_{11} field plot shown in Figure 2-2 it is clear that knowing the field orientation is going to be important when an FSS sheet is placed around the guide. This is to ensure that any symmetry present can be preserved in terms of the FSS placement in relation to the maximum E-field location inside the waveguide. This can be achieved by aligning the FSS in relation to the maximum E-field orientation. This allows consistency of FSS placement to be achieved and could, if desired, act as a point of reference for measurements. Field orientation when an electric dipole is used is well understood\(^4\) and the position of the maximum electric field strength is known. Field orientation across a slot used as a feeding mechanism for waveguide is also well documented in the literature. Due to the variety of ways in which a feed can be realised, the topic is discussed in detail in Chapter 3.
2.1.1.5 Conduction Loss

The calculation of the waveguide characteristics has been done assuming that conduction loss, $\alpha$, is equal to zero. The metallic parts of the waveguide will have conduction losses associated with the material used, which due to its advantages of cost, availability and ease of manufacture is in this case Brass, with a conductivity, $\sigma$, of $2.564\times10^7 S/m$. Collin\textsuperscript{5} states the loss equations for TE and TM modes as (2.31) and (2.32) respectively.

\[ \alpha_{c(TE_0)} = \frac{R_c}{a\eta} \sqrt{1 - \left( \frac{f_c}{f} \right)^2} \left[ \left( \frac{f_c}{f} \right)^2 + \frac{n^2}{(p')^2} \right] \text{Np/m} \]  

(2.31)

\[ \alpha_{c(TM_0)} = \frac{R_c}{a\eta} \frac{1}{\sqrt{1 - \left( \frac{f_c}{f} \right)^2}} \text{Np/m} \]  

(2.32)

Where $R_c = \frac{\omega\mu_0}{\sqrt{2\sigma}}$

Two wave guides of different diameters have been used and these are described in Chapter 4 and Chapter 5 and these have the following conductor losses associated with the waveguide part of the antenna shown in Table 2-1.

<table>
<thead>
<tr>
<th></th>
<th>a=7.5mm Waveguide</th>
<th>a=38mm Waveguide</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal working Frequency</td>
<td>9GHz</td>
<td>1.8GHz</td>
</tr>
<tr>
<td>$\beta$ (rad/m)</td>
<td>104.4</td>
<td>22.3</td>
</tr>
<tr>
<td>Conduction loss (Np/m)</td>
<td>0.06</td>
<td>$4.89\times10^{-3}$</td>
</tr>
</tbody>
</table>

Table 2-1 Conduction losses associated with the waveguides

The losses associated with conduction are small and will not have a significant effect on the antenna performance. They can be safely disregarded for the purposes of this project.

2.1.2 The Effect of Dielectric Loading

The presence of a dielectric material in the waveguide, used for supporting the FSS, has an interesting and useful effect on a standard, air loaded, waveguide which, due to its importance, will be discussed in the following sections. As can be seen in the
discussion presented in 2.1 the equations that govern waveguide behaviour, such as (2.15), (2.16) and (2.26) for the TE case, have an \( \varepsilon \) associated with them, such that:

\[
\varepsilon = \varepsilon_0 \varepsilon_r
\]  

(2.33)

Where \( \varepsilon_r \) is the relative permittivity of the filling material. The same is true for the TM case if the corresponding equations (2.20), (2.21) and (2.28) are used. If the filling material is anything other than air then the \( \varepsilon_r \) value will be greater than one. It is for this condition that the following sections relate.

### 2.1.2.1 Cut off Frequency

If a dielectric with relative dielectric constant \( \varepsilon_r \) is placed inside a waveguide with a cut frequency of \( x \text{GHz} \) the cut off frequency will remain the same, provided that the diameter of the waveguide is scaled by \( \sqrt{\varepsilon_r} \) as shown by equation (2.16) for the TE case.

This result is useful in as much as lower frequencies can propagate inside dielectrically loaded waveguides than would be possible if no dielectric filling were present. This leads to noticeable size reductions if higher dielectric constants are used. Materials with a relative permittivity of 36, such as Zirconium Tin Titanate, have been successfully used as antenna pucks\(^5\), in this case with a helical metalisation pattern present on the dielectric material’s surface. Ultimately, it is hoped that materials such as this could eventually be used for this proposed type of antenna, due to the size and weight savings that can be achieved.

### 2.1.2.2 Impedance

Given that \( \eta \) decreases, as given by equation (2.26), for any dielectric present the impedance of the waveguide will decrease by a factor of \( \sqrt{\varepsilon_r} \). This, again, is an advantage since the lower waveguide impedance will be easier to impedance match to the generally lower impedance feeding structures such as coaxial cable and microstrip.

### 2.1.2.3 Dielectric Loss

The dielectric filling used in the waveguide will not be ideal. To calculate the dielectric loss, \( \alpha_d \), that can be expected from the material equation (2.34) can be used, derived in Appendix B Derivation of Waveguide Losses:
\[ \alpha_d = \frac{k^2 \tan \delta}{2\beta} \text{ Np/m} \quad (2.34) \]

Two waveguides of different diameters have been used and these are described in Chapter 4 and Chapter 5 and these have the following dielectric losses associated with the waveguide part of the antenna shown in Table 2-2.

<table>
<thead>
<tr>
<th>Nominal working Frequency</th>
<th>( a=7.5\text{mm} ) Waveguide</th>
<th>( a=38\text{mm} ) Waveguide</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \beta ) (rad/m)</td>
<td>9GHz</td>
<td>1.8GHz</td>
</tr>
<tr>
<td>104.4</td>
<td>22.3</td>
<td></td>
</tr>
<tr>
<td>Dielectric loss (Np/m)</td>
<td>( 4.9 \times 10^{-4} )</td>
<td>( 9.2 \times 10^{-5} )</td>
</tr>
</tbody>
</table>

Table 2-2 Dielectric losses associated with the waveguides

The \( \tan \delta \) value quoted for the PTFE material is \( 3 \times 10^{-6} \).

Due to the low loss characteristics that the PTFE possesses the losses associated with the dielectric are small and will not have a significant effect on the antenna performance. They can be safely disregarded for the purposes of this project.

2.1.2.4 Bandwidth

In section 2.1.2.1 it was shown that the cut off frequency reduced by a factor of \( \sqrt{\varepsilon_r} \). The reduction applies to both the upper and lower cut off frequencies. In terms of the percentage bandwidth calculation given in section 2.1.1.3 it is clear that this will not change the resulting value.

2.2 Overview of Electromagnetic modelling techniques

Numerous techniques exist for modelling of electromagnetic structures. A choice has to be made about which technique is best suited to a particular problem and the availability of the technique usually through access to suitable computer code.

A TLM based solver has been employed for this work. The aim of the section is to provide an overview of the different techniques that are generally available and then to show that a TLM method is suited to the geometries considered. The TLM method is then explored and details such as meshing, boundary conditions, terminations, excitation and object geometry representation are considered in more detail.
Computational electromagnetics (CEM) can be broken down into two distinct branches: Time Domain, such as the finite difference/element technique and Frequency Domain techniques such as the Method of Moments (MOM) technique proposed by Harrington\(^7\). Some methods, such as the finite element method, can be formulated in the time or the frequency domain. In the book edited by Rao\(^8\) the four main principle models of CEM are described by Miller as follows:

1. Time domain differential equation (TDDE) models which have had their use increased due to advances in larger faster computers.
2. Time domain integral equation (TDIE) models. The technique has been known for 30 years and has seen increased attention in the last decade and is suited for a wide variety of applications.
3. Frequency domain integral equation (FDIE) models are the most widely studied of the techniques and were the first to be implemented.
4. Frequency domain differential equation (FDDE) models have seen increased use in recent years and have been traditionally applied to low frequency applications.

When considering a problem Umashankar and Tafl\oe\(^9\) suggest the following should be considered:

- Range of structure size, in wavelengths
- Generic structural geometry (shape) features
- Structure material composition

For each consideration listed above a table of suitability has been produced and compares the merits of four techniques consisting of the following:

- High-frequency diffraction theory, including Geometric Theory of Diffraction (GTD) and its variants
- The frequency domain integral equation formulation, with solution via the method of moments (MOM)
- Iterative approaches for the MOM employing preconditioned conjugate gradient techniques
The time-domain differential equation formulation, with solution via either the finite difference time domain or the finite element time domain methods.

Range of Structure size in wavelengths

<table>
<thead>
<tr>
<th>Modelling Approach</th>
<th>&lt;λ/10</th>
<th>→λ</th>
<th>→10λ</th>
<th>→100λ</th>
<th>&gt;100λ</th>
</tr>
</thead>
<tbody>
<tr>
<td>High Frequency Techniques</td>
<td>-</td>
<td>-</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
</tr>
<tr>
<td>Method of Moments</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>MOM/Iterative</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
<td>-</td>
</tr>
<tr>
<td>FD-TD / FE-TD</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
</tr>
</tbody>
</table>

- = Not applicable

? = Limited by computational resource
✓ = Applicable

Table 2-3 Range of Structure size in wavelengths

Generic structural geometry (shape) features

<table>
<thead>
<tr>
<th>Modelling Approach</th>
<th>Close to spherical</th>
<th>With spherical edges</th>
<th>With corner reflectors</th>
<th>With arbitrary unloaded cavities</th>
<th>With arbitrary loaded cavities</th>
</tr>
</thead>
<tbody>
<tr>
<td>High Frequency Techniques</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
<td>-</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Method of Moments</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MOM/Iterative</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>FD-TD / FE-TD</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

- = Not applicable

? = Unclear if applicable
✓ = Applicable

Table 2-4 Generic structural geometry (shape) features
Structure material composition

<table>
<thead>
<tr>
<th>Modelling Approach</th>
<th>Perfect Electrical conductors</th>
<th>Perfect magnetic conductors</th>
<th>Homogeneous dielectrics</th>
<th>Inhomogeneous dielectrics</th>
<th>Lossy media</th>
<th>Anisotropic media</th>
</tr>
</thead>
<tbody>
<tr>
<td>High-Frequency Techniques</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>Method of Moments</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
</tr>
<tr>
<td>MOM/Iterative</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>?</td>
<td>?</td>
</tr>
<tr>
<td>FD-TD / FE-TD</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
<td>✓</td>
</tr>
</tbody>
</table>

- = Not applicable
? = Unclear if applicable
✓ = Applicable

Table 2-5 Structure material composition

The problem to be solved is in the 10 to 100λ range, and has a homogeneous dielectric present. These criteria suggest that a time domain approach would be better suited to solving the proposed geometry.

Rao provides the following advantages:

- **Computational efficiency**
  For certain problems fewer arithmetic operations are required when calculations are performed in the time domain. If the model requires the early time peak response of the object to an impulsive field then the TD model is a more efficient approach than the FD model which requires frequency samples across a broad bandwidth followed by a Fourier transform to obtain the desired result. This work requires a broadband response so a TD solution is better choice because it provides a transient response that is only limited in bandwidth by the frequency content of the excitation and the time space sampling used in developing the model.

- **Problem requirements**
TD solutions can model non-linear media, components and time-varying media and components in a straightforward manner which a FD cannot.

For a fuller explanation of the efficiency, memory requirements and similar parameters between various schemes Miller\textsuperscript{10} provides a good source of information. A comparison of time domain methods is given by Celuch-Marcysiak and Gwarek\textsuperscript{11}.

Of all the time domain schemes it is the Finite Difference Time Domain (FDTD) method and the TLM method that are the most popular. The two techniques have been shown to be equivalent in certain circumstances by Johns\textsuperscript{12}, noting the subsequent comments\textsuperscript{13}, Simons and Bridges\textsuperscript{14} and Chen \textit{et al.}\textsuperscript{15}. The formulation of the two is different with the FDTD method based on the Yee cell\textsuperscript{16} and the TLM method on the Symmetrical Condensed Node (SCN)\textsuperscript{17}, the point at which the transmission lines intersect. The Yee cell and the SCN are shown in Figure 2-3.

![Figure 2-3 The Yee cell (shown on the left) and the Symmetrical Condensed Node from Christopoulos\textsuperscript{22}](image)

The SCN of Figure 2-3 is further broken down into transmission line clusters and theses are shown in Figure 2-4.
The major difference between TLM and FDTD is that FDTD generates field components in separate halves of the time step, in other words it calculates E followed by H, followed by E etc. As can be seen from Figure 2-3 the field components are displaced in space at the edges of the cell. In TLM the E and H fields are calculated at the centre of the cell at the same time step and are co-located. This difference, and how it affects the modelling of a problem, has long been debated. The SCN does require more memory than the Yee cell formulation. The stability of the TLM method tends to be better than the FDTD method for rectangular faced cells where the side lengths differ, and this may mean that models containing a high permittivity material can be represented more efficiently. The advantage in using rectangular faced cells is that a problem can be discretised to give the best meshing in a particular direction without the need additional computer resource in the others.

It has been suggested by Simons et al \(^{18}\) that TLM could model the areas around sharp edges more accurately than FDTD and it was concluded that more cells were required in the FDTD solution to get the same accuracy as the TLM method.

### 2.2.1 Transmission Line Modelling Overview

The Transmission Line Modelling method or, more commonly, the TLM method utilises the fact that Maxwell's equations can have equivalent circuit representations, first derived by Kron \(^{19}\) in 1944. Johns and Beurle \(^{20}\) developed the method in 1971.
These ideas were further developed and a commercial electromagnetic solver, Micro-Stripes, was launched. There is a good, brief summary of the development of the TLM method and Micro-Stripes in the manual supplied with version 3 by Flomerics EM21, formerly Kimberley Communications Consultants Ltd and an in depth study of the TLM method is given by Cristopoulos22. Regardless of the TLM software package chosen, there are TLM modelling issues common to all such as meshing, boundary conditions and signal propagation etc. which are detailed in the following sections.

2.2.1.1 Meshing

Three possible types of meshing exist for the discretised transmission lines regular, variable and multigrid. Care should be taken to synchronise the pulses across the mesh, as discussed in section 2.2.1.4, so the variable and multigrid mesh need additional mathematical treatment to ensure this. Examples in two-dimensions of the three mesh types are shown below in Figure 2-5.

![Figure 2-5 Regular, variable and multigrid meshes](image)

2.2.1.2 Radiation boundary conditions

To reduce a problem to a manageable size there must be a bound placed on the problem to represent free space. There are several types of radiation boundary condition available in the TLM method. These include:

- The Absorbing Boundary Condition (ABC). This boundary is equivalent to terminating the transmission line with a load of characteristic impedance to approximate an infinite length and hence the free space nature of the problem. One consideration to take into account is if the angle of incidence at the boundary is not perpendicular. If this occurs the correct impedance can be achieved by adjusting the reflection coefficient.

- Perfectly Matched Layers (PMLs). PMLs where originally introduced by Berenger23 for use in the FDTD methods. The PML was first implemented in the
TLM method by using an interface between the FDTD and TLM network\textsuperscript{24,25}. However, it has been shown that the use of a non-uniform TLM-FDTD mesh provides more inaccurate absorbing conditions than obtained by Bérenger with the FDTD method\textsuperscript{26}. Work by Pena and Ney\textsuperscript{27} and others\textsuperscript{28,29,30,31} is continuing the development of PMLs in the TLM method.

The PML boundary itself is constructed by splitting the six electromagnetic field components into 12 sub-components and introducing anisotropic electric and magnetic conductivities\textsuperscript{32}. The nature of these sub-components is such that they cannot exist in reality\textsuperscript{33}.

2.2.1.3 Terminations

The following types of termination are achieved by terminating the transmission line in the appropriate manner.

- A plane of Symmetry or open circuit. This is modelled using the appropriate open circuit on the transmission line.
- A Perfect Electric Conductor (PEC) and conducting boundary are represented by a short circuit such that the reflected signal is identical to the incident signal, but with the sign changed.

2.2.1.4 Signal Propagation on a TLM mesh

Once a model has been constructed and discretised into a mesh consisting of the representative mesh of transmission lines energy has to be applied onto it. It is easiest to consider a two-dimensional case for this. If a 1V pulse is incident upon a mesh the resulting reflection and transmission coefficients, calculated using transmission line theory, are shown in Figure 2-6.

\begin{figure}[h]
\centering
\includegraphics[width=0.5\textwidth]{reflection_transmission_coefficients.png}
\caption{Reflection and transmission coefficients resulting from a voltage pulse from Christopoulos\textsuperscript{22}}
\end{figure}
Figure 2-7 shows how this process is extended and how the scattering event propagates across the mesh.

Figure 2-7 Impulse placed on the mesh and the subsequent scattering events from Christopoulos\textsuperscript{22}

2.2.1.5 Material Representation

Material and object representation are dealt with by varying the electrical properties of the transmission lines. As suggested by Kron\textsuperscript{19} equivalence exists between the circuit and field representation of electrical problems. If the transmission line of Figure 2-8 is studied the equivalences can be derived.
R and G are the series resistance and shunt admittance and L and C are the series inductance and shunt capacitance per unit length, $\Delta x$. From Kirchhoff's voltage and current laws the following can be obtained assuming $\Delta x \to 0$:

$$\Delta X \frac{\partial v}{\partial x} = -L \frac{\partial i}{\partial t} - iR \quad (2.35)$$

$$\Delta X \frac{\partial i}{\partial x} = -Gv - C \frac{\partial v}{\partial t} \quad (2.36)$$

Differentiating (2.35) with respect to $t$ and (2.36) with respect to $x$ to eliminate voltage leads, after combining, to an equation for current and, using a similar procedure, but eliminating current this time, an equation for voltage. These are shown in (2.37) and (2.38) respectively.

$$\frac{\partial^2 i}{\partial x^2} = \frac{GR}{(\Delta x)^2} i + \frac{1}{(\Delta x)^2} (GL + RC) \frac{\partial i}{\partial t} + \frac{LC}{(\Delta x)^2} \frac{\partial^2 i}{\partial t^2} \quad (2.37)$$

$$\frac{\partial^2 v}{\partial x^2} = \frac{GR}{(\Delta x)^2} v + \frac{1}{(\Delta x)^2} (GL + RC) \frac{\partial v}{\partial t} + \frac{LC}{(\Delta x)^2} \frac{\partial^2 v}{\partial t^2} \quad (2.38)$$

Rewriting (2.1) and (2.2) in their alternative form gives:

$$\nabla \times \vec{E} = -\frac{\partial \vec{B}}{\partial t} \quad (2.39)$$

$$\nabla \times \vec{H} = j + \frac{\partial \vec{D}}{\partial t} \quad (2.40)$$

These may be expanded and, for a one-dimensional problem with variation only in the forward $x$-direction, the only non-zero field components will be $E_y$ and $B_z$. This reduces the full expansion to the following two equations:

$$\frac{\partial E_y}{\partial x} = -\frac{\partial B_z}{\partial t} \quad (2.41)$$
\[-\frac{\partial H_z}{\partial x} = j_y \cdot \frac{\partial D_y}{\partial t}\]

Multiplying \((2.42)\) by \(\mu\) and applying Ohm's Law, \(j_y = \sigma E_y\), where \(\sigma\) is the electrical conductivity of the medium gives:

\[-\frac{\partial B_z}{\partial x} = \mu \sigma E_y + \mu \varepsilon \frac{\partial E_y}{\partial t}\]  

\((2.43)\)

Taking the derivative of \((2.43)\) with respect to \(t\), differentiating \((2.41)\) with respect to \(x\) and eliminating \(B_z\) gives:

\[\frac{\partial^2 E_y}{\partial x^2} = \mu \varepsilon \frac{\partial^2 E_y}{\partial t^2} + \mu \sigma \frac{\partial E_y}{\partial t}\]

\((2.44)\)

or, alternatively in terms of the current density:

\[\frac{\partial^2 j_y}{\partial x^2} = \mu \varepsilon \frac{\partial^2 j_y}{\partial t^2} + \mu \sigma \frac{\partial j_y}{\partial t}\]

\((2.45)\)

An identical equation is obtained for \(B_z\). These equations describe one-dimensional wave propagation in a lossy medium.

For the case when \(R = 0\) in equation \((2.37)\) in can be re-written as:

\[\frac{\partial^2 i}{\partial x^2} = \frac{LC}{(\Delta x)^2} \frac{\partial^2 i}{\partial t^2} + \frac{GL}{(\Delta x)^2} \frac{\partial i}{\partial t}\]

\((2.46)\)

Comparing equations \((2.45)\) and \((2.46)\) the following equivalences can be drawn out:

\[i \leftrightarrow j_y, \quad \frac{C}{\Delta x} \leftrightarrow \varepsilon, \quad \frac{L}{\Delta x} \leftrightarrow \mu, \quad \frac{G}{\Delta x} \leftrightarrow \sigma\]

\((2.47)\)

\(E_y\) can be associated with the voltage on the line. Equation \((2.47)\) shows that as parameters vary along the \(x\)-direction changes in permittivity, permeability or electrical conductivity may be modelled by changing the capacitance, inductance and conductance of the line. However, the same propagation time must be maintained throughout the transmission line network so that synchronism is ensured. This approach acts as the starting point for evaluation of the SCN, although additional techniques need to be utilised to deal with modelling issues such as dispersive media.
2.2.2 A brief overview of the Micro-Stripes Program

During the course of this work development of the Micro-Stripes package has continued and version 3 to 5.6 have been used. Backward compatibility has been maintained and earlier designs can be run on later versions of the software with no loss in accuracy. Where possible additional features added in later versions have been avoided, such as Perfectly Matched Layers, PMLs, unless there is a compelling case in terms of accuracy to do so, to ensure that results are comparable. This, it is hoped, will minimise any variations in the results due to modelling discrepancies between the software versions.

In summary a model can be defined using either a text based description language, or a graphical design package. The model is built up from primitive shapes, such as cubes and cylinders, and logical functions are used to join or subtract them from each other so that the model geometry can be built up. Material attributes, such as dielectric constant and conductivity etc., can be attached to the shapes defined in the model. Excitation points are defined and these, in the types of components used for this work, take the form of E and H field patterns to excite waveguide modes.

Depending on the required results different outputs can be selected. For scattering parameter results it is usual to have to place points to sample the forward and reverse going waves which are then used to calculate $S_{11}$ and $S_{21}$ etc. Care needs to be exercised in the positioning of these points to ensure that the correct results are obtained. Far-field plots can be obtained by enclosing the device with a surface on which the near-field strength is measured and this can be post-processed to calculate the far-field radiation pattern. Surface currents can be obtained in a similar way, as can fields inside a device such as a waveguide.

To ensure accuracy the model has cell edges defined and the number of cells required between these points has to be specified. Areas where a high resolution is needed, or complex field interactions are occurring, require more cells than those areas where there is expected to be little change in the fields or those areas of less interest. Obviously the highest accuracy will be obtained for a mesh with a very small cell size. When modelling curved object, such as circular waveguide, a certain amount of
irregularity will appear in the surface caused by the surface being represented in square blocks. This is corrugated surface is commonly called the staircasing effect. This staircasing creates a longer current path around the surface affecting impedance calculations. Micro-Stripes automatically corrects the impedance of staircased surfaces. The effect staircasing has can be reduced by using a smaller meshing size. This becomes impractical if the model is of any reasonable physical size in both terms of memory required and computational time. This is due to the ‘time step’ decreasing so it will take longer to calculate the solution for a given problem.

If areas of fine mesh extend to areas where less resolution is required they can be ‘lumped’ together to increase their size. There is, of course, a drawback in as much as there is a maximum size the cells can be and the largest cell length in any one direction defines this. Usually the largest cell size needs to smaller than 1/10th of the free space wavelength at the model’s ‘maximum model frequency’. In regions where dielectric material is present a smaller mesh is required than in areas of free space, due to the variation of fields over shorter distances. Within Micro-Stripes there exists an auto meshing facility that will, if required, take these factors into account.

Once the model has been created and the excitation and outputs defined it is discretised. The mesh can be a regular or graded e.g. one that has cells that are rectangular in nature. This is done automatically by breaking the model down into a 3-dimensional grid that forms the interconnected transmission lines that, based on the solving of the associated equations, gives the electromagnetic results. Field plots are viewed directly using a viewer supplied with the software. Scattering parameter results produce a time domain file. This has to be filtered, have the in going and out going waves resolved and then a scattering matrix is created.

2.3 Frequency Selective Surfaces and Frequency Selective Guides

Vardaxoglou describes FSSs as “printed arrays of conducting elements (or apertures in a conducting plane) and can be shaped, both electronically and physically, in order to be accommodated in a specific application”. Different geometries for use as FSSs exist including Dipoles, Tripoles, Crossed dipoles, Jerusalem crosses, Square loops and circular loops. Three of these geometries are shown below in Figure 2-9.
Figure 2-9 Popular geometries for use as FSSs

A typical response for this type of element, when formed into an array and illustrated in Figure 2-10, is shown in Figure 2-11 with a clearly defined stop band at the FSS resonant frequency, $F_R$, and good transmission through the FSS at pass band frequency, $F_T$. The photo reverse of the elements shown in Figure 2-9 will result in a single pass band response. The wavelength, $\lambda_R$, at which resonance will occur, is approximately equal to $2L$ for the dipole array and $4L$ for the square loop array. Works by Munk$^{35}$ and Wu$^{36}$ is a good source of additional theory and information.
Figure 2-11 Typical FSS array response

Computer programs, based on the equations Vardaxoglou describes, have been written at Loughborough University for the design of FSSs\textsuperscript{37}. The verification of these programs has been well documented in the literature\textsuperscript{34}. The program to predict the response of a square loop conducting element array requires an element to be described in terms of the conducting element length and spacing in both dimensions on a planar sheet. The program is designed to compute the response of a double square loop. The smaller of the two conducting loops, which fits inside the first, is still present, but its dimensions can be reduced in size so its effect becomes negligible.

Traditionally when this type of structure is used for a radome or horn application the FSS is designed on a planar sheet. Once these structures are curved the planar periodicity may not be preserved and Makino et al.\textsuperscript{38} have developed a technique that allows the arrangement of elements such that the desired performance can be obtained when the structure is curved.

FSSs have been rolled into cylindrical structures and used as guiding structures\textsuperscript{39,40} called Frequency Selective Guides (FSGs). Work by Loukos\textsuperscript{41} has previously specified the element width to be $1/7$th of the element length when defining the modes and fields within a FSG. Eade\textsuperscript{42} looked at the transition of solid waveguide to FSGs. The propagation within these structures and their transition to waveguide has been studied by Eade et al.\textsuperscript{43} and the convention used to define the width of the element, based on its length, continued. As 3-D modelling of the elements on a dielectric has been undertaken in this study the definition of element width as $1/7$th of the element length need not be continued and gives more design options for the FSS. A good
approximation used for the design of FSGs\textsuperscript{44} suggests that an incidence angle of 60° to the FSS is a good starting approximation for obtaining the correct element size and spacing of the FSG. This is based upon the fact that propagation would not occur inside a solid circular waveguide if the propagating wave were incident at 0° or 90° e.g. when $f \leq f_c$.

One design constraint that must be placed upon the FSS specification is that there must be an integer number of elements around the rod. The simplest way to ensure that this is achieved is to divide the required number of elements into the rod’s circumference and use this figure as the periodicity of the FSS. This ensures that the rod has uniform coverage and that there are no gaps or excessive overlaps present when the FSS is rolled around the rod. Now that the periodicity has been defined, the element length can be varied to achieve the desired frequency response. If a design cannot be realised then an alternative number of elements should be tried. Designs that have an element length similar in dimension to the periodicity should, generally, be avoided, as suggested by Cox\textsuperscript{44}, since the close coupling of the elements can lead to errors in the FSS frequency response prediction.

2.3.1 FSS Manufacture

As has been previously stated FSSs are periodic arrays of metallic elements. It would be impractical to have these constructed and attached to the rod individually so the technique used to manufacture FSS, as used successfully at Loughborough University, has been employed. The process used at Loughborough to manufacture them is as follows:

- Copper sheet on a Mylar backing is readily available and this, in the past, has proved suitable for this type of application. The copper is 18μm thick and is attached to a Mylar backing with either a 0.03mm or 0.04mm thickness. The loss tangent of this material, tanδ, is ≈ 0.03. When compared to the thickness of the rods used the effect of the dielectric backing is negligible. The elements can be constructed using the same design and manufacture process commonly used for PCB manufacture.

- Once the intended design has satisfied the design criteria the dimensions are entered into a CAD package and copied to form an array of elements with exactly the same size and periodicity on the sheet. This design is then plotted on an
acetate sheet and a photo-negative image created on another sheet which is needed for the next step.

- The photo-negative mask is placed on the surface of the copper sheet, which has a photo-resist layer, and then exposed to ultra-violet light. A Ferric Chloride solution is used to selectively remove the unwanted copper leaving the desired element on the dielectric substrate. The time required to do this is brief and the achieved accuracy can be within ±6μm of the mask dimensions. The final designs are however not accurate to the original dimensions specified and it was found that this is due to the photo-plotter having an accuracy of ±0.1mm, even when pens with of nibs of 0.1mm are used. The ink would spread at the corners of elements as the pen dwells at these places before changing direction. Efforts were made to try and reduce this error using techniques such as backing the pen off from the corners. Even with care slight ‘dog boning’, the name given to this defect due to the shape of the element produced, does still occur. Visual checks are made on the finished photo plots and, if the dog boning is excessive, they are replotted, or areas where high accuracy had been achieved on the finished FSS sheet are used.

- The finished FSS sheet is trimmed and strips cut corresponding to the number of elements required around the rod. A small overlap is usually left to facilitate adhesion of the FSG to itself using an adhesive that allows for repositioning. If the elements are to be placed on a dielectric rod the overlap is not so critical.

The generic diagram for this type of structure manufacture is shown in Figure 2-12. An overlap is not shown, but would normally be no larger than the width between the metallic elements.
Storage of FSGs has always been problematic to ensure they maintain their circular uniformity after being carefully rolled around a tight radius to encourage curvature of the metallic elements on the more flexible dielectric substrate. Since the elements are to be placed on a solid rod curvature can be achieved by the use of adhesive and the circular profile of the rod alone. Work carried out by Cox\textsuperscript{44} has suggested improvements to the storage of FSGs.

2.4 Dielectric Rod Waveguides and Antennas

Dielectric rod antennas are quite common and have been well documented in the literature. Kajfez and Guillon\textsuperscript{45} have thoroughly documented the behaviour of this type of device. In simplistic terms they can be thought of as an optical fibre with the rod represented the core and the surrounding air the cladding. Optical Fibres are extensively documented and Okoshi\textsuperscript{46} is a good starting point if this analogy is pursued.

Dielectric rods have been successfully used as antennas in the past. Dombek\textsuperscript{47} reports on experimental results obtained from Dielectric rod antennas. Work by Kishk and Shafai\textsuperscript{48} confirmed Dombek's results numerically. They have also presented results for dielectric rods with other types of end, such as tapered and bulbous and reported the gains associated with these. The rod description and excitations described by Kishk and Shafai, based on Dombek's work, have been implemented in Micro-Stripes. The results Micro-Stripes predicted are given in Figure 2-13 for each $L/\lambda_0$ value for the $-3$ and $-10$dB beam widths and have been overlaid with the previously
published results. These results show that Micro-Stripes is a valuable tool to accurately predict this type of antenna performance.

![Figure 2-13 Micro-Stripes prediction of Dombek's results](image)

2.5 Overview of the Leaky Wave Action of a FSG and Leaky Wave Antennas

Away from the resonance the FSGs will exhibit some form of partial leakage of power. This leakage of power is termed the leaky wave action of the FSG. In essence the leaky wave action causes power to be radiated from the FSG at an angle away from boresight of the antenna. This angle can be scanned with frequency and power directed away from the boresight and at a predictable angle. Eade\textsuperscript{43} used equation (2.48) to calculate the propagation constant for FSGs and, since the applications proposed here is similar, it should again be of use.

$$\phi = \sin^{-1}\left(\frac{\beta}{k_0}\right)$$  \hspace{1cm} (2.48)

Antennas have previously made use of this type of radiation leakage, though usually on a rectangular waveguide\textsuperscript{49} or dielectric slab\textsuperscript{50}. The use of a FSS wall in a conventional metal waveguide\textsuperscript{51} has also been employed. This type of antenna is generally termed a Leaky Wave Antenna or LWA. Guglielmi and Boccalone\textsuperscript{52} give a good design guide for this type of structure. A circular LWA has been presented by EL-Muslimany \textit{et al.}\textsuperscript{53} employing circular metallic strips around a dielectric rod antenna and using a circularly symmetric TE wave. This design is similar in nature to
the one Ramo, Whinnery and Van Duzer present for mode suppression, this one allowing the good propagation of $TE_{0m}$ modes.

Attempts have been made to form the beams from this type of antenna by other means. Chen et al propose a method that uses the phasing of two feeds to steer the beam using a sum and difference technique. Steering of beams electronically has been reported by, amongst others, Horn et al. and Huang et al. using p-i-n diodes. These methods utilise additional circuitry and, as such, provide a more complex solution to beam steering than the method being proposed here.

References


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3 Feeding Techniques

3.1 Feeding Overview

Mobile telephone antennas work with linear polarisation and GPS antennas require the ability to receive a circularly polarised signal. There are various considerations concerned with the feed such as complexity, bandwidth and how applicable they are to a final engineering solution of the antenna and all these need to be considered when designing the feeding structure.

Excitation of rectangular waveguide using coaxial cable is well documented in the literature. Circular waveguide excitation with coaxial cable is less well documented. An electric field probe can be used to excite the \( \text{TE}_{11} \) mode inside a circular waveguide. A length of coaxial cable placed along the centre line of a waveguide, i.e. passed through the sidewall into the waveguide, is sufficient to induce the required fields. Maximum power input to the waveguide will occur if the 50\( \Omega \) coaxial feed is matched to the waveguide impedance. It is usual to approximate this match to be at a distance of \( \lambda_g/4 \) from the back wall of the waveguide to remove the effect of the rear wall. Deshpande and Das\(^1\) propose a way to evaluate the input impedance of a coaxial probe taking into account the necessary factors such as probe and waveguide dimensions, distance from the back wall and frequency of operation. This type of feed is suitable for \( \text{TE}_{11} \) excitation. This paper was evaluated using Mathcad, noting the authors' subsequent corrections\(^2\), using the dimensions relevant to the designs proposed within this work.

If the probe is placed through the end wall of the circular waveguide and bent to form a loop terminating on the sidewall of the waveguide the \( \text{TE}_{11} \) mode can, again, be excited. Deshpande and Das\(^3\) also investigate this type of end launcher. Since the waveguides used in this study will have a solid dielectric filling this type of launcher would prove extremely hard to manufacture and, as such, is given no further consideration.

Microstrip lines can be used to excite slots, which can be used as feed mechanisms for waveguides\(^4\). In the simplest form this consists of a microstrip line running at 90\(^\circ\) to
the long edge of a thin slot etched in the ground plane which forms the back wall of the waveguide. Slot coupling is most commonly used as a feeding technique for microstrip patch antennas as used extensively by Pozar\textsuperscript{5} but can be used to excite waveguides as described by Izadian and Izadian\textsuperscript{6}, Das et al.\textsuperscript{7} and Omar and Dib\textsuperscript{8}.

The slot is generally narrow and is usually designed to be \( \lambda/2 \) in length. The slot will have a impedance value associated with it that, for impedance matching purposes, will be required for the design of a suitable feeding network. Kraus\textsuperscript{9} calculates this impedance using the technique of complementary dipoles. His analysis requires the complementary dipole to possess a large length-diameter ratio.

Fan et al.\textsuperscript{10} proposed a method that used microstrip directly as a feeding mechanism. This type of feed was evaluated and considered impractical if circular polarisation was required due to the complexity of the circuit board construction. Hajian et al.\textsuperscript{11} proposes a novel design for the transition, but it seems over complex if it were to be mass-produced.

For simulation purposes, where the feeding structure need not be considered, specific, ideal TE modes can be excited inside circular waveguides without the excitation of the evanescent modes that can be associated with the other types of feed techniques listed above. This can be done in two ways: a specific mode can be excited within Micro-Stripes\textsuperscript{12}, the simulation software, by using a default setting, or a thin line of charge can be placed manually across the diameter of the waveguide. Rules govern the placement of mode excitations so care has to be taken that the excited mode is, indeed, the one desired. The regions defined as metallic in Micro-Stripes can have impedance and conductance associated with them. If these properties are used and set to the characteristic impedance of the waveguide, e.g. for the back wall of the waveguide, they can stop unwanted reflections and improve the match by, in effect, making the waveguide appear infinite in length. If coaxial launchers are used for the launching of \( \text{TE}_{11} \) in simulations the back wall of the waveguide is given metallic properties, such as conductance, and will reflect incident waves, as would be expected in a real system. This allows, by careful positioning of the coaxial cable, the optimisation of the return loss.
The next higher order mode for circular waveguide is the TM₀₁ mode. The software cannot excite TM modes using a default set-up as it can with a TE₁₁ mode. This mode has been considered for some of the designs presented and has been implemented using a coaxial end launcher through the back wall of the waveguide. The field patterns generated were checked along the length of the waveguide to confirm that they did produce the to the expected TM₀₁ mode.

3.1.1 Circular polarisation considerations

To obtain circular polarisation (CP) Balanis¹³ lists the criteria as:

a) The field must have two orthogonal linear components, and
b) The two components must have the same magnitude, and
c) The two components must have a time-phase difference of odd multiples of 90°.

Feeds for two different frequencies are needed. One for the initial study at 9GHz and one for the frequency of intended use at 1.8GHz. The 1.8GHz feed will require the ability to generate a circularly polarised field.

3.2 Coaxial Feed

Kishk and Shafai¹⁴ describe a dielectric rod antenna fed by waveguide. This paper was used as a basis for the feeding mechanism. A feed was modelled in Micro-Stripes and a cut away diagram is shown in Figure 3-1. Different pin insertion lengths were tried to obtain the best return loss. For each insertion distance a different rod with a section removed for correct fitting of the coaxial cable was required. To improve the match further the amount of dielectric about the inner conductor was varied to create an air gap between the connector and the dielectric rod. The three conditions tried were 0mm (e.g. tight fitting), 0.1mm removed and all the surrounding dielectric removed. The SMA coaxial connector passed through a brass spacer so that the insertion distance could be varied. By using spacers of different thickness the pin insertion could be varied from 1/4, 3/8, 1/2 and 3/4 of the distance across the diameter.
Waveguide and retaining bolts
Pin
Air gap
Spacer (to obtain correct insertion)
SMA connector and retaining bolts

Figure 3-1 Cut away through a Micro-Stripes model of a Dielectric Rod Antenna

The model shown in Figure 3-1 has the dielectric rod stopping at the end of the waveguide and a pin insertion of $\frac{3}{8}$ of the diameter of the waveguide.

3.2.1 Measured Results

The device described in section 3.2 was measured using a Marconi 6210 reflection analyser. Results from both a long and short rod are presented below in Figure 3-2 using a pin insertion of $\frac{3}{8}$ of the diameter.

Figure 3-2 Comparison of long and short pin fed dielectric waveguide

The $TE_{11}$ mono-mode region is between 8.3 and 10.6GHz for the guide dimensions and permittivity. As can be seen the longer rod, represented by the circular data labels in Figure 3-2, shows a good reflection coefficient at 9.7GHz. In an engineering...
application the long rod is the only practicable option, since the FSS elements have no means of support on the short rod. This measurement was used as a benchmark to compare the modelled results against. As Micro-Stripes had not been used in the group to model such detailed structures confidence had to be in place before further, lengthy, simulations were run and relied upon. This allowed the model to be fine tuned to fit as closely as possible to the measurements. This allowed experimentation within the model to establish the accuracy required that gave an acceptable comparison against he measured results. Once this had been established Micro-Stripes was relied upon to accurately predict the results and only then were the devices manufactured.

3.2.2 Micro-Stripes Model Results

Due to a frequency shift causing the troughs to appear at higher frequencies all the way across the measured frequency band the model was run again keeping the pin insertion constant and varying the relative permittivity of the dielectric. The results for the short dielectric rod are presented in Figure 3-3.

![Figure 3-3 Measured and modelled results of the short dielectric rod antenna](image)

It can be seen that the frequencies match for the 1/2 diameter insertion modelled and measured results. The manufacturer’s data sheet quotes the value of the relative dielectric constant of PTFE, as 2.1. These results were obtained by using a value of 2.0 in the model. A lower permittivity fill would indicate that the modelled diameter is smaller than originally thought. This could be due to the stair casing in the model,
due to the mesh discretization, causing the waveguide diameter to appear a different size to the intended one. The difference in amplitudes could possibly be explained by the fact that manufacturing tolerances meant that the spacer used did not quite insert the tip of the pin the exact distance expected and modelled. A small amount of variation in the insertion distance is seen to have a considerable effect and if the spacer is not precise, but too thin, small air gaps will be formed around the tip of the coaxial cable. This could explain the differences seen in the measured and modelled results shown in Figure 3-3. Overall, the frequency of operation match each other and the differences in amplitude are due to matching issues. The modelled results do give confidence in the ability of Micro-Stripes to model this type of structure.

3.3 Microstrip Slot Coupled Feed

Work carried out by K. Zorzos designed a feed and single slot to produce mono-mode TE$_{11}$ in a waveguide by means of slot coupling. While there is nothing wrong with a pin fed waveguide it is not easily realised if the technology is to be used inside a mobile telephone handset. Ideally to excite TE$_{11}$ the pin should pass through a sidewall of the waveguide but in a handset it is most convenient to pass the probe through the end wall.

Using microstrip, a feed can conveniently be taken from the circuit board to the antenna. A decision had been taken that, if slot coupling could be shown to work, it would be preferred to the pin feed.

3.3.1 Modelling Results

Calculations taking into account the possible dielectric loading effects of the MC3D substrate, $\varepsilon_r=3.53$, on the slot suggested a slot length of less than $\lambda/2$ would be the correct length for resonant coupling at 9GHz. Based on this premise a slot length of between 9 and 10mm should be resonant. Micro-Stripes simulations confirmed that there was, indeed, a resonance at the correct frequency with these dimensions.

3.3.2 Measured Results

A single slot was routed onto the ground plane of several MC3D substrates. Four slots were constructed in total varying by 0.25mm between 9.25mm and 10mm. The 9.5mm slot was found to give the best performance in terms of reflection coefficient
at 9GHz. A line 1mm wide and 89mm long fed it. This gave the line approximately 100Ω impedance. The measurement system is 50Ω and the impedance of the loaded waveguide, at 9GHz, is known to be 683Ω, so the combination impedance from the microstrip line and slot are at intermediate values. These intermediate steps in impedance should act as a crude transformer between the initial and final values of impedance. The ground plane formed the back wall of the waveguide, which was held securely to it by Allen bolts. The measured reflection coefficient is shown below in Figure 3-4.

Figure 3-4 Single slot fed waveguide reflection coefficient

As can be seen the slot exhibits the best reflection coefficient very close to 9GHz and at 9GHz the reflection coefficient is -25dB. If the response in Figure 3-4 is compared to the pin feed it can be seen that the response is lower, -5dB, across the region where transmission is occurring. In both cases the available bandwidth is small, approximately 7.5% at -10dB, compared with the possible 22-23% bandwidth available in circular waveguide.

3.3.3 Circular Polarisation Considerations

As mentioned previously in 3.1 Global Positioning Systems (GPS) antennas need to receive circularly polarised fields. Antennas for this application need to be able to support a spinning electric field. This approach requires two identical feeds in terms of impedance, but one offset by 90° from the other as described in 3.1.1. Separate slots and feed lines can be used as shown by Murakami et al. for a dual frequency
patch and it is easy to see how, with phase delay, this design could be used to generate CP. After the dimensions of the waveguide have been reduced by dielectric loading it is possible that the slots would radiate outside of the waveguide and so this design may not be practical. The spinning field can be achieved using a cross-slot in the substrate ground plane as used in the design of CP Dielectric Resonator Antennas by Huang et al.\(^1\). Huang uses a single feed and slightly different slot lengths to control the two near-degenerate resonant modes to create CP. A combination of these feeds is proposed such that the slots will be placed centrally under the waveguide forming a cross but fed separately from a single source. One feed has to have an additional line length added, equivalent to a 90° phase shift, whilst the impedances have to be matched between the single input line and, when split, the two lines. This design allows the feeds to be offset from the centre of each slot, a technique that can be used to impedance match the feed and slot\(^9\).

### 3.3.3.1 Modelled results

The splitting of the feed can be realised entirely using microstrip. A possible feed design to achieve the necessary CP for the antenna is shown below in Figure 3-5

![Figure 3-5 Proposed CP feed using microstrip](image)

(which also partly shows a field output region).

**Figure 3-5 Proposed CP feed using microstrip**

This design gave the following simulated return loss response shown in Figure 3-6. A bandwidth of 9.1% at -10dB has been suggested from the modelling with this feed, so it has slightly more bandwidth than the single slot used previously in 3.3.2.
From the field output animation it was clear the E-field of the TE_{11} mode did rotate within the waveguide, along its length. There was no way the uniformity of this rotation can be gauged directly from Micro-Stripes. The feed design was analysed by requesting a field output in the waveguide and post-processing it to look at the rotation of the electric field vectors along the length of the waveguide. The ‘core’ of cells extracted each had the same dimensions in the x, y and z directions. A Mathcad sheet was written to plot the magnitude and direction of a field point in the centre of the guide. The shape traced out by the electric field vector should be circular as it increments one row of cells at a time along the waveguide. Figure 3-7 shows the plot obtained for the feed shown in Figure 3-5. This type of plot allowed comparisons to be made between the different feeds. Although not perfectly circular this was the best result obtained from the feeds simulated.

Figure 3-6 Simulated Return Loss of the CP feed
It has been our experience that once a design is manufactured it often has to be experimentally tuned to give a desired reflection coefficient due to manufacturing tolerances and the like. If this design were manufactured in the standard way using Printed Circuit Board (PCB) technology the fine-tuning would not be an option. The lines could be placed using copper tape, backed with adhesive, to allow for adjustment. Although used successfully for the simple feed for the 9GHz design this method is imprecise in terms of getting the critical dimensions of the design to the required accuracy. Where the design required a bend two pieces of tape would have to be adhered one on top of another continue the microstrip line. The tape has more loss associated with it than a permanently etched/routed line. When the device was constructed it was found that the measurements did not correspond well to the simulated response and it was felt that this was due to the manufacture using the tape. As this prototype would need to give results that justified PCB manufacture it was felt unjustifiable to continue with this approach to generate CP.

3.3.4 Coaxial Cable Driven Slot Coupling

Another option for the feeding lines is to use coaxial cable to drive the slots as described by Kraus. Again, an extra length of coaxial cable can be used to add the required phase shift but this design requires a splitter for the cable if it is to be driven from a single source.
Models of this type of design were simulated and gave satisfactory results, but suggested they would be quite narrow band. Adjustment is possible in the design by re-soldering the cable inner and outer to the ground plane. If CP is not required both slots can still be fed simultaneously with the only effect being that the linear electric field will be rotated by a constant 45°. It was decided that coaxial cable would be a better option than trying to use microstrip lines. Phase shift, although possible with extra cable length, would be added by a dedicated component that could be adjusted to compensate for any difference in coaxial cable length. Reflection coefficient can be varied by changing the position of the coaxial cable in relation to the slot, due to the different impedance presented. This is easily achieved by re-soldering. A general overview is shown in Figure 3-8 and a more detailed view of the coaxial cable in Figure 3-9.

Figure 3-8 General view of the slot arrangement
Cell size in Micro-Stripes was sufficiently small to allow the features to have enough detail to accurately represent the physical structure. This is shown in Figure 3-10.

The expected $S_{11}$ result is shown in 3.3.4.2.

**3.3.4.1 CP considerations of the coaxial cable driven slot**

The axial ratio of an antenna is a measure of the electric field polarisation, and hence can show how well the feeding structure is generating CP. Balanis\(^{13}\) states the necessary equations to calculate the axial ratio, but these do not give the correct result.
for certain phase and amplitude conditions. To overcome this a Mathcad sheet was written to calculate, based upon the polarisation ellipse, the axial ratio by calculating the major and minor axis values based on simple trigonometry. This approach allowed the axial ratio to be calculated and optimised from the design information available from the manufacturer's data sheets of devices needed to achieve the CP requirements e.g. splitters and phase shifters.

Good CP is expected as the axial ratio one would expect from this feed has been calculated to be 1.071, or 0.593dB. A plot similar to the one shown in Figure 3-7 is given in Figure 3-11 and, as can be clearly seen, a much better CP is achieved.

![Mathcad plot of the CP generated by the coaxial driven slots based on the relative magnitude and phase of E_x and E_y components to 1V/m](image)

Figure 3-11 Mathcad plot of the CP generated by the coaxial driven slots based on the relative magnitude and phase of E_x and E_y components to 1V/m

It is hoped that this approach will reduce the amount of rear going radiation since, unlike microstrip line, coaxial cable usually does not radiate along its entire length.

### 3.3.4.2 Measured Results

It can be seen from Figure 3-9 that the end of the coaxial cable has been terminated with a metal plug that has a matched load of 50Ω in the simulation and the TEM field has been created internally within the cable. In reality this is not how the device would be fed, or constructed. The coaxial cable is taken to the edge of the substrate and a SMA connector is added to allow attachment of the cable to the splitter, phase
shifter and the network analyser depending on what the desired outcome is. The outer conductor is scored using a scalpel until it is possible to slide it from the inner using a pair of pliers. The dielectric is trimmed and then the SMA slipped on to the cable and soldered in place to form a permanent attachment. Whilst it is hoped that the SMA connections to both cables will be identical differences will occur and this will further degrade the measured axial ratio.

Due to the necessary attachment of the inner of the coaxial cable to the substrate metallisation layer, solder is required. This will, of course, change the profile of the connection from that of the one simulated. By careful application of solder the amount of interference caused by this can be minimised.

By using the adjustment on the phase shifter to add or subtract phase an ideal shift of 90° can be obtained, even if the cables are not of the same exact length. The device is capable of a shift of 60°/GHz, up to a maximum working frequency of 2.3GHz. This allows for the possibility of a phase shift at frequencies where the TM01 mode will propagate though, due to the nature of the TM01 E and H field patterns, CP is not possible and the upper frequency limit is restricted by the splitter. There is, of course, no reason why a single coaxial feed can not be excited, and the other matched, to allow higher frequencies of the TM01 mode to propagate though the presence of the second slot could give rise to unwanted results.

The S11 of the feeding structure was measured with a single piece of coaxial cable soldered in place and this measurement was compared to the simulation prediction, detailed in 3.3.4, and the comparison is shown in Figure 3-12.
Figure 3-12 Measured and Simulated results of the coaxial driven slot

Good agreement can be seen in the return loss across the whole of the measured frequency range. The measured bandwidth is less than predicted as is the depth of the return loss, but this is typical of this type of comparison since the simulated result should be treated as an ideal.

If each coaxial cable is left unmatched and the $S_{11}$ of the other measured the results should be identical, if the feeds are identical. Figure 3-13 shows that the comparison between the two is good. The differences are most likely due to the irregularities in the soldering at the ends of each coaxial cable.
Figure 3-13 Comparisons between feeds and the effect of the splitter

The splitter will have an effect on the measurements and, for this reason, a full characterisation of the feeds with the splitter present was undertaken. The steps undertaken are detailed below:

- Coaxial cable feed 1 matched and J3 of the splitter matched. The $S_{11}$ of coaxial cable feed 2 from J2 obtained.
- Coaxial cable feed 1 matched and J2 of the splitter matched. The $S_{11}$ of coaxial cable feed 2 from J3 obtained.
- Coaxial cable feed 2 matched and J2 of the splitter matched. The $S_{11}$ of coaxial cable feed 2 from J3 obtained.
- Coaxial cable feed 2 matched and J3 of the splitter matched. The $S_{11}$ of coaxial cable feed 2 from J2 obtained.

It was found that there was no noticeable difference in measurement between steps 1 and 2 and steps 3 and 4. There were, however, differences between steps 1 and 4, as would be expected having noted the differences in the soldering of the cables. This result is shown in Figure 3-14. Due to the components used this measurement is now only valid up to 2GHz.
Figure 3-14 Effect of the splitter on the coaxial feeds

The $S_{11}$ of the antenna and feed network with both feeds in operation is shown in Figure 3-15. In this configuration the field pattern will be rotated by 45°, within the waveguide, in relation to the centre line of one of the feeding slots.

Figure 3-15 Return Loss of the antenna with the splitter present

3.3.4.3 Calibration of the phase shifter

Since phase is critical to the CP measurements the calibration of the device must include adjusting the phase between the two ports of the phase shifter to give a 90° difference at 1.8GHz. The steps taken are detailed below and the general set-up is shown in Figure 3-16.
The HP8753D network analyser is set-up and calibrated to perform an $S_{21}$ measurement.

- Port 3 is matched and the phase of the $S_{21}$ measurement from port 1 to port 2 is obtained.
- Port 2 is matched and the phase of the $S_{21}$ measurement from port 1 to port 3 is obtained.
- The phase is adjusted to give a difference of 90° between the phase at port 2 and the phase at port 3. For example if port 2 is $-10^\circ$ then port 3 should read $80^\circ$.
- Step 2 is repeated to ensure that the phase adjustment has not altered the initial phase reading.

The $S_{11}$ changes as phase adjustments are made. The graph shown in Figure 3-17 gives the measured $S_{11}$ when a 90° shift is present between the end of the cable at port 2 and the end of the cable at port 3.
Figure 3-17 $S_{11}$ when there is a 90° phase shift between ports

References


12 Micro-Stripes V3 to V5.6 User and Reference Manuals, Flomerics Electromagnetic Division, TLM House, PO Box 5097, Nottingham, NG16 3BF.


4 Prototype Design of the new antenna at 9GHz

FSS elements will resonate when a wave, at the right frequency, is incident upon them. The FSS is used to control the antenna pattern by utilising the leaky wave behaviour exhibited by a non-resonant array, as described in section 2.5. It will be shown that this property can be utilised in antenna design to control the beam patterns produced by the dielectric rod. In a provisional study for his MSc. Zorzos studied, through measurement, the suitability of different FSS resonators placed on a dielectric rod antenna fed by a linearly polarised microstrip slot.

The antenna consisted of a waveguide 15mm in diameter filled with PTFE which, when the dielectric was taken into account, gave a cut off frequency equivalent to a waveguide of 10.6mm in diameter. The waveguide was 50mm long and the PTFE rod protruded for a further 50mm from the opening of the waveguide. The study looked at the following FSS resonator shapes dipole, crossed-dipole and square loops, which were designed to resonate at 9GHz. This frequency was chosen based on the reflection coefficient plot shown in Figure 3-4. Measurements to characterise the antenna fully were performed and the pertinent results, mostly relating to the antenna when only half coverage of elements are in place, are presented in this Chapter.

4.1 Measurement Set up

FSS elements were placed on the antenna at 0 and 90° to the electric field and the general set-up is shown in Figure 4-1. The elements are etched on to a copper sheet with a dielectric backing, as described in section 2.3.1, and the effect this might have had was studied by comparing measurements with and without a specially etched piece of Mylar sheet that, through etching, was devoid of all copper. It was found that the additional dielectric layer of Mylar under the elements did not have any appreciable effect on either the antenna patterns or the $S_{11}$ results, so all measurements only have the Mylar backing where the elements are present. This is particularly relevant if only half coverage of elements is used i.e. the part of the PTFE rod that does not have metallic coverage of elements will also not have any Mylar backing. The position of the elements was varied longitudinally along the guide. It was found that the elements seemed to have most effect when their backing was butted against the end of the waveguide, probably due to the fact that the distance was...
then equal to the distance between elements and so optimised. All the measurements presented in this chapter have the elements butted against the waveguide. None of the antenna measurements performed used an endcap. To characterise the antenna fully the E and H plane was swept with the feed rotated 90°. A 360° sweep was taken around the circumference. Due to the measurement capabilities and the antenna support position this was done in three, separate, stages of 140°, to include an overlap, and then the separate plots joined to create a full 360° plot.

![Diagram](image)

**Figure 4-1 Measurement set up for the FSS elements placed on the slot fed dielectric rod antenna (Rear elevation)**

### 4.2 Dipole Measured Results

Two dipole designs were manufactured for 9GHz changing some of the design parameters, but still maintaining the resonant frequency. The measurements presented relate to the design with the following characteristics that are summarised in Table 4-1 and the results presented relate to half coverage of elements only.

<table>
<thead>
<tr>
<th>Design</th>
<th>Length of Element</th>
<th>Width of Element</th>
<th>Periodicity (D_x)</th>
<th>Periodicity (D_y)</th>
</tr>
</thead>
<tbody>
<tr>
<td>'12'</td>
<td>12.9mm</td>
<td>0.3mm</td>
<td>15.0mm</td>
<td>15.0mm</td>
</tr>
</tbody>
</table>

**Table 4-1 Summary of Dipole FSS Design**

### 4.2.1 Radial Measurements

The plots shown in Figure 4-2 and Figure 4-3 are the 360° sweeps for the rod with the elements at 0 and 90° orientations.
Figure 4-2 360-degree plot of Dipoles with elements at 0 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.

Figure 4-3 360-degree plot of Dipoles with elements at 90 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.

The 360° plots show little difference between the plain dielectric rod and the one with the addition of elements. A slight difference does occur at approximately 45° as the null present in the dielectric rod has been cancelled out.

4.2.2 Cut Measurements

The plots below show the patterns obtained when the measurements, described in section 4.1, were performed.
Azimuthal cut of design '12' with elements at 0 degrees

Figure 4-4 Co and Cross Polar plot of a Dipole array on a dielectric rod antenna

From Figure 4-4 and Figure 4-5 which compare the effect of placing the FSS elements at 0 and 90° respectively to the feed it can be seen that most difference is observed when the elements are at 0° to the feed. This is particularly notable in the Co-polar measurement, where the pattern at -50° is different. The elements seem also to pull up the cross-polar values when compared to the plain dielectric rod.
Figure 4-6 Co and Cross Polar plot of a Dipole array on a dielectric rod antenna

Figure 4-7 Co and Cross Polar plot of a Dipole array on a dielectric rod antenna

Again, Figure 4-6 and Figure 4-7 show that most effect is seen when the elements are at 0° to the feed. These cuts are at 90° to the first set of cut measurements and it is again the co-polar plot that shows the differences with deeper troughs at ±60°. The cross-polar level has changed little.

4.3 Crossed-Dipole Measured Results

Two crossed-dipole designs were manufactured for 9GHz. The measurements presented relate to designs with characteristics that are summarised in Table 4-2 and the results presented relate to half coverage of elements only.
<table>
<thead>
<tr>
<th>Design</th>
<th>Length of Element</th>
<th>Width of Element</th>
<th>Periodicity (D_x)</th>
<th>Periodicity (D_y)</th>
</tr>
</thead>
<tbody>
<tr>
<td>'15'</td>
<td>10.9mm</td>
<td>0.3mm</td>
<td>12.0mm</td>
<td>12.0mm</td>
</tr>
<tr>
<td>'17'</td>
<td>10.9mm</td>
<td>0.5mm</td>
<td>12.0mm</td>
<td>12.0mm</td>
</tr>
</tbody>
</table>

Table 4-2 Summary of Crossed-Dipole FSS Design

4.3.1 Radial Measurements

The plots shown in Figure 4-8, Figure 4-9, Figure 4-10 and Figure 4-11 are the 360° sweeps with the elements at 0° and 90° orientations for designs ‘15’ and ‘17’ respectively.

![360plot of Design '15' Cross-Dipoles at 0 Degrees](image)

Figure 4-8 360-degree plot of Crossed-Dipoles with elements at 0 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.
Figure 4-9 360-degree plot of Crossed-Dipoles with elements at 90 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.

Figure 4-10 360-degree plot of Crossed-Dipoles with elements at 0 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.
Figure 4-11 360-degree plot of Crossed-Dipoles with elements at 90 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.

It can be seen that if the elements are placed at 90° to the maximum TE_{11} field line there is little effect. If the elements are placed at 0° the antenna patterns have reduced power at 0 and 180° so they appear ‘flatter’ at the top and bottom of the plots.

4.3.2 Cut Measurements

Using the same method as previously described cut measurements were taken of the Crossed-Dipoles. Plots for the designs detailed in Table 4-2 are shown below in Figure 4-12, Figure 4-13, Figure 4-14 and Figure 4-15 for the azimuth series of measurements.
Azimuthal cut of design 'IS' with elements at 0 degrees

Figure 4-12 Co and Cross Polar plot of a Cross Dipole array on a dielectric rod antenna

Azimuthal cut of design 'IS' with elements at 90 degrees

Figure 4-13 Co and Cross Polar plot of a Crossed-Dipole array on a dielectric rod antenna
Figure 4-14 Co and Cross Polar plot of a Crossed-Dipole array on a dielectric rod antenna

When the elements are at 0° there appears to be an improvement in the directivity of the antenna by 5dBm at 0°. When the elements are placed at 90° a large improvement, approximately 10dBm, is seen in the cross-polar component between ±10°.

The elevation series of measurements are shown below in Figure 4-16, Figure 4-17, Figure 4-18, and Figure 4-19.
Figure 4-16 Co and Cross Polar plot of a Crossed-Dipole array on a dielectric rod antenna

Figure 4-17 Co and Cross Polar plot of a Crossed-Dipole array on a dielectric rod antenna
Figure 4-18 Co and Cross Polar plot of a Crossed-Dipole array on a dielectric rod antenna

Figure 4-19 Co and Cross Polar plot of a Crossed-Dipole array on a dielectric rod antenna

Comparing design '15' and '17' with the plain rod antenna it can be seen that having the elements at $0^\circ$ improves the gain of the antenna, but the cross-polar level rises. Having the elements at $90^\circ$ appears to worsen the antenna performance, with a drop in gain noticeable, and no discernible improvement in the cross-polar level to compensate for this.
4.4 Square-Loop Measurement Results

Ten square-loop designs were simulated and four were taken forward into manufacture. The results of two of these are presented below. A summary of their characteristics is shown below in Table 4-3 and the results presented relate to half coverage of elements only.

<table>
<thead>
<tr>
<th>Design</th>
<th>Length of Element</th>
<th>Width of Element</th>
<th>Periodicity (Dx)</th>
<th>Periodicity (Dy)</th>
</tr>
</thead>
<tbody>
<tr>
<td>‘8’</td>
<td>9.0mm</td>
<td>0.4mm</td>
<td>13.1mm</td>
<td>13.1mm</td>
</tr>
<tr>
<td>‘10’</td>
<td>9.0mm</td>
<td>0.6mm</td>
<td>12.7mm</td>
<td>12.7mm</td>
</tr>
</tbody>
</table>

Table 4-3 Summary of Square-Loop FSS Design

4.4.1 Radial Measurements

The plots shown in Figure 4-20, Figure 4-21, Figure 4-22 and Figure 4-23 are the 360° sweeps for the rod with elements at 0 and 90° orientations.

Figure 4-20 360-degree plot of Square-Loops with elements at 0 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.
Figure 4-21 360-degree plot of Square-Loops with elements at 90 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.

Figure 4-22 360-degree plot of Square-Loops with elements at 0 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.
Figure 4-23 360-degree plot of Square-Loops with elements at 90 degrees to the feed. Plan view from Figure 4-1. Radial axis denotes power in dBm.

The results for the 90° case are very similar to the dipole results. When the elements are at 0° there is a noticeable cancelling out of both nulls at 45° and 315° in Figure 4-20 and Figure 4-22. Figure 4-22 shows strong compression of the antenna pattern at 0 and 180°. If, at resonance, the FSS were acting as a metal sheet, as would be expected, then it could be argued that more power would be reflected from them and the antenna pattern would distort resulting in patterns seen in Figure 4-20 and Figure 4-22. A similar result, using Dipoles and an end cap, has been obtained using Micro-Stripes TLM modelling by Epa². This result would seem to support his results.

4.4.2 Cut Measurements

Using the same method as the dipole measurement cuts were taken of the square-loops. Plots of the designs detailed in Table 4-3 are shown below. Figure 4-24 to Figure 4-27 are for the azimuth series of measurements and Figure 4-28 to Figure 4-31 are for the elevation series.
Figure 4-24 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna

Figure 4-25 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna
Figure 4-26 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna
Figure 4-28 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna

Figure 4-29 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna
Figure 4-30 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna

Figure 4-31 Co and Cross Polar plot of a Square Loop array on a dielectric rod antenna

Few, if any conclusions can be drawn from Figure 4-24 and Figure 4-25. Figure 4-26 shows a shifting towards the left-hand side of the graph. It is not clear what caused this. It could be surface wave suppression. Figure 4-28 shows an interesting trend with the cross-polar measurement. It is higher than the dielectric rod for most angles, but at 4° drops to a very low value. At this angle the antenna appears to be suitable for systems with linear polarisation requirements and would work exceptionally well.
4.5 Half element coverage of Square-Loop elements.

The antenna was simulated with and without partial coverage of elements present. The elements were placed in a 3x2 arrangement along the dielectric rod. This meant that the rod had a complete longitudinal coverage of elements and only a partial coverage around the circumference. The model used for simulations is shown in Figure 4-32 with partial coverage of elements. The excitation point and the far field output request are also visible.

![Figure 4-32](image)

**Figure 4-32 Complete modelled assembly of waveguide, rod and FSS elements**

The design for the square loop resonator with a periodicity of 12.7mm, based on design ‘10’, and detailed in 4.4 was modelled using a TLM modelling package, Micro-Stripes\(^3\). The measurements were carried out with a ground plane present, to allow the use of a microstrip feed, but this was not included in the simulations to improve the computational speed and reduce the time required to obtain results. Figure 4-33 shows the azimuthal plot for the antenna, Figure 4-34 shows the elevation plot and Figure 4-35 the cross-polar elevation cut. These are plotted against the corresponding measurement shown in Figure 4-26 and Figure 4-30.
Figure 4-33 Co-Polar Azimuth Cut for the 9GHz Design

Figure 4-34 Co-Polar Elevation Cut for the 9GHz Design
Figure 4-35 Cross Polar Elevation Cut for the 9GHz Design

Reasonable agreement can be seen with the measured results and this raises confidence that these devices can accurately be modelled using the software and that reliance can, once again, be placed upon the simulation results.

Figure 4-33 shows that there is a shift in beam of approximately 20° shift from the bore-sight direction. There is, as expected, no shift present in the Co-Polar elevation cut of Figure 4-34. If the elements are rotated so that they are parallel to the E-field the interaction between the two becomes insignificant and no beam shifting is observed.

To show that it is the elements that cause the beam shifting, frequencies other than 9GHz were modelled. Figure 4-36 shows the rod at 9GHz with no elements. The beam shifting effect of the elements of approximately 20° is clearly seen in Figure 4-37 but has returned to a vertical position at 10GHz in Figure 4-38. A model was run in which the elements were replaced by a metal sheet. The sheet altered the beam shape in a similar manner to the elements, but the angle of shift was constant with frequency. This clearly shows that the beam shifting effect is frequency dependant and caused by the presence of the FSS elements.
Figure 4-36 Beam shifting simulation of a plain rod at 9GHz

Figure 4-37 Beam shifting simulation of a plain rod with elements at 9GHz

Figure 4-38 Beam shifting simulation of a plain rod with elements at 10GHz
References


3 Micro-Stripes, Flomerics Electromagnetic Division, Nottingham, England.
5 Prototype Design of the new antenna at 1.8GHz

The 9GHz design has shown that a dielectric Rod antenna with suitable positioning of FSS can change the radiation pattern of the antenna. This chapter documents the work that has been undertaken, and presents new results when the waveguide has circularly polarised (CP) fields launched within it and documents the effects of two endcap configurations, described as flat and coned.

The work detailed in Chapter 4 looked at a rod with a relative permittivity of 2 and a diameter of 15mm. These dimensions are suitable for a working frequency of 9GHz. As already described, the aim of this work is for this antenna to work at mobile communication frequencies. These frequencies are typically between 1.8 and 2.5GHz, so the previous model has been scaled in size and the element size redesigned accordingly so that the centre frequency of the dielectrically loaded waveguide is now 1.8GHz.

5.1 Manufacture and Modelling detail

Again, all simulations have taken place using a TLM modelling package. For each polarisation of the feed there are six basic antenna combinations if an endcap, full element coverage and half element coverage are modelled.

5.1.1 Waveguide Detail

The waveguide radius that corresponds to propagation at a centre frequency of 1.8GHz is 53.74mm. With the addition of a dielectric material, which has previously been found to model accurately at \( \varepsilon_r=2 \), this radius scales to 38mm. The length of the rod was chosen to be 137.1mm, with the waveguide extending for 55.4mm along the length. This length of the rod matches closely the ratio of diameter to length of existing antennas used by the Centre for Mobile Communications Research at Loughborough University\(^1\). It allows two elements to fit longitudinally along the length of the rod observing the correct periodicity for the FSS along the rod.

5.1.2 FSS detail

The FSS was designed assuming it was on a planar sheet with a medium of \( \varepsilon_r=1 \) one side and a layer \( \varepsilon_r=2 \) extending for 76mm behind it. This assumption does not allow for an exact design because the FSS will be placed on a curved surface, but allows for
a good approximation that the FSS will work in the correct frequency region. This fact is desirable so that the centre frequency of operation of the FSS will be shifted away from the waveguide centre frequency. The design program was run with an angle of incidence of 0.5, 45 and 85° to examine the range of possible responses from the FSS. It was found that this variation in angle did not move the FSS resonant frequency by a significant amount. The transmission between 1.7GHz and 1.9GHz was below -18.5dB when the angle of incidence was at 45°. The final design had a periodicity of 39.8mm, an element length of 37.7mm in the x and y directions and the width of the element was 5mm. These dimensions correspond to little propagation through the FSS at 1.80Hz, e.g. it would be resonant if measurements were taken using a planar sheet. A frequency sweep was performed with the FSS in place on the dielectric rod to find the exact resonance of the FSS, obtained by noting the frequency of maximum boresight gain e.g. the region where the FSS is resonant and guiding energy. The result of this sweep is shown in Figure 5-1.

![Frequency Scan of Boresight Antenna Gain](image)

**Figure 5-1 Frequency scans of the FSS on the rod antenna**

Where reference is made to 'co and cross-polar' in Figure 5-1 it relates to the antenna and dipole positions used for the angular scan measurements described in 5.1.4.

Clearly, the FSS exhibits maximum gain at around 1.73GHz. It also has a significant effect on the cross-polar performance and the co-polar performance above 1.85GHz when compared to the plain rod antenna. The leaky wave region of the structure can be seen clearly by the drop in boresight gain of the co-polar measurement of the
antenna with elements between 1.8 and 1.85GHz. Drop in boresight gain may be due to a differing return loss between the rod with element and the rod without. However, a measurement of this type was performed for the 9GHz antenna and no appreciable difference was noted.

5.1.3 Modelling Detail

The modelling package has various ways that the model can be constructed and simulated. This section details some of the issues involved with the modelling so a greater insight can be gained. Depending on the symmetry used computer run time was up to 6 hours per design on a P3 800MHz computer with 512MB of RAM.

5.1.3.1 Cell size

Maximum cell size is automatically defined by the package based on the maximum model frequency and any dielectrics used so that an accurate frequency response can be calculated. The size of cells between elements was as little as 0.4mm in some areas within the model where it was felt that extra accuracy was required. Figure 5-2 shows the cell outline superimposed on top of an element. Two planes of symmetry have been used in this example, based on the nature of the E and H-fields within the waveguide, and exist on the left and bottom edge of the cells, resulting in only the area overlaid with the grid being modelled. Staircasing of the model's curved edges, described in section 2.2.1 will occur.

Figure 5-2 Typical cell detail around the elements
5.1.3.2 Boundary Conditions

Version 5.5 and later of Micro-stripes supports Perfectly Matched Layers (PMLs) for the first time. These are an alternative type of Absorbing Boundary Condition (ABC) to the ‘matched termination’ type used in earlier versions to define the model bounds. PMLs are usually more suited to antenna applications and they are extensively used in FDTD packages for radiating structures. Since previous work had been carried out successfully using the absorbing boundary conditions provided within earlier versions these have been used for this work also.

5.1.3.3 Feeding Technique

To speed up the simulation time the TE\textsubscript{11} mode was set up within the waveguide using the ‘Initial Mode’ feature within MicroStripes. It has been shown that other feed techniques, described in Chapter 3, are able to set up the desired modes correctly, but it was decided that the ‘Initial Mode’ approach would be acceptable for this set of results. As a slot has not been included within the model definition, the associated ground-plane is also omitted.

5.1.3.4 Matching

The rear wall of the waveguide was matched to the impedance of an incident wave at 1.8GHz in a dielectrically loaded guide. Although this match is only exact for a single frequency, this type of match is commonly used in this type of problem where the reverse going wave has to appear in phase with the forward going wave. In this case, it is achieved by absorbing the wave on the back wall and thus approximating an infinite length of waveguide. In practice this is true for most feeds that rely on fractions of a wavelength between the feed and end wall of the waveguide to provide the match. All other metals defined within the model used the standard values for their metallic properties.

5.1.3.5 Symmetry Issues

The TLM modeller allows advantage to be taken of any symmetry present in the model with electric and magnetic wall boundary conditions. The criteria for this to apply are that the electric or magnetic fields incident upon the plane of symmetry must be normal to it. Obviously, this requires the model to possess some form of symmetry, but also for the fields inside it to possess symmetry along the same plane.
These conditions are met within the waveguide, but can not be guaranteed outside of the waveguide after the fields have interacted with the elements. Models were run with and without symmetry planes and the results were identical. Based on this, all the models utilised symmetry, where appropriate, to reduce the modelling time required.

5.1.4 Measurement Detail

Due to the low frequency measurements needed for this antenna the use of horn antennas for the measurements become prohibitive due to their size. For this reason, a wide band dipole antenna, with a centre frequency of 1.8GHz, was used as the receiving antenna. The return loss of the dipole was measured and the response is show in Figure 5-3. The gain of the dipole had been previously measured as 3dB.

![Measured Response of the Dipole Antenna](image)

**Figure 5-3 Response of the receiving dipole antenna**

Due to the nature of this response, it can be expected that results measured at 1.7GHz will be 7dB worse, if compared directly without any post-processing, than that of the result measured at 1.8GHz. This variation will occur at the other frequencies that were measured, 1.9 and 2.0GHz, but to a greater extent. For this reason, the results have been normalised to the maximum value of the co or cross-polar component for each variation of the antenna to allow meaningful comparison of the measured antenna performance. This allows information about the relative amplitudes of the outputs to be gained. Outputs from Micro-stripes are normalised to the maximum of each frequency step requested. Without normalising the measurements a second time
direct comparison can not be made with the measurements, though it is only the relative amplitude of the results that change and this can easily be compensated for visually when looking at the plots.

Due to the nature of the feeding structure used the resultant E-field will not lie parallel to either of the feeds, but instead bisect them. The measurements were performed with the antenna orientated such that the resultant E-field was used as reference for the measurements. The notation used for the measured results is shown in Figure 5-4.

![Diagram of antenna orientations for 1.8GHz measurements](image)

**Figure 5-4 Antenna orientations for 1.8GHz measurements**

Figure 5-4a and Figure 5-4b show the antenna position for the azimuthal cuts for the cross and co-polar sweeps respectively. Figure 5-4c and Figure 5-4d show the antenna position for the elevation cuts for the co and cross-polar sweeps respectively. The dipole was swept in the horizontal plane about the antenna in the two orientations shown to measure both the co and cross-polar components of the antenna radiation pattern.
5.2 Plain rod results

The antenna was modelled in the form described in section 5.1. The rod can have no elements attached or can have full or partial coverage of FSS elements. The measurements consisted of obtaining sweeps when the frequency output was varied between 1.7 and 2.0GHz, the TE_{11} mono-mode region of the waveguide, in steps of 0.1GHz. Comprehensive antenna pattern measurements, as described in 5.1.4, were performed to verify the simulations.

5.2.1 Simulated Results

This section presents the pertinent results found in the simulations carried out on this antenna design. The antenna model was run, as described in 5.1.3, and the following results were obtained.

5.2.1.1 Linearly Polarised Results

Simulated outputs for the plain dielectric rod antenna are shown in Figure 5-5 and Figure 5-6. Due to the arbitrary nature of the axes chosen for modelling purposes ‘0 degrees from the E-Plane’ relates to the azimuthal co-polar measurements and ‘90 degrees to the E-plane’ relates to the elevation co-polar measurement. Using this, comparison between simulated and measured results of Figure 5-7 and Figure 5-9 can be made.

Figure 5-5 Co-Polar cut 0 degrees from E-plane (azimuthal co-polar)
As can be seen there is good agreement between the co-polar elevation plots, but the azimuthal plot of Figure 5-7 has some large nulls present which do not appear in the simulation, these are thought due to fringing effects from the ground plane which was absent from the simulations.

5.2.2 Measured Results

Following the simulations the antenna configurations were constructed and measured in the anechoic chamber facility at Loughborough University.

5.2.2.1 Linearly Polarised Results

The antenna was fed by splitting the feed from the sweeper unit resulting in each coaxial cable being fed in phase. Results for the plain rod are shown in Figure 5-7 to Figure 5-10, the rod with elements in Figure 5-15 to Figure 5-18 and the rod with half coverage of elements in Figure 5-19 to Figure 5-22. These results are presented in full to act as a reference for the subsequent results.
Figure 5-7 Azimuthal co-polar result for linearly polarised plain rod

Figure 5-8 Azimuthal cross-polar result for linearly polarised plain rod
Figure 5-9 Elevation co-polar result for linearly polarised plain rod

Figure 5-10 Elevation cross-polar result for linearly polarised plain rod

To allow the effects of the endcap and elements to be studied, the following plots were combined at the 1.8GHz frequency: Plain Rod, Rod with flat endcap, Rod with elements, Rod with endcap and elements, coned endcap and coned endcap and elements. These plots are show in Figure 5-11 to Figure 5-14 for the linearly polarised measurements.
Figure 5-11 Comparisons of different antenna parameters at 1.8GHz – Azimuthal co-polar

Figure 5-12 Comparisons of different antenna parameters at 1.8GHz – Azimuthal cross-polar
Figure 5-13 Comparisons of different antenna parameters at 1.8GHz – Elevation co-polar

Figure 5-14 Comparisons of different antenna parameters at 1.8GHz – Elevation cross-polar

Taking the results presented in Figure 5-11 they show that the plain rod antenna power maximum is away from 0°. This is unexpected and the other results do not show similar results. For this reason, this particular result has to be treated with some suspicion. The addition of a flat endcap appears to have little effect, with the result broadly following the plain rod result. The plot does however show that the elements do, indeed, have a broadening effect on the antenna pattern with a clear improvement being seen at angles of −20 to −80° and 40 to 70°. Generally the improvement is
about 5dB, but can, in places be as great as 12dB. Performance of the antenna with elements and a flat endcap is poor, when compared with the other antennas, coned endcap antenna excluded. The coned endcap results display a poor antenna pattern and do not appear to be suitable for the intended applications in the azimuthal co-polar plot.

The azimuthal cross polar result in Figure 5-12 shows that at boresight and positive angles the plain rod is generally lower than that for the co-polar case, though at some frequencies it is equal to, or even above, the co-polar value. At angles other than boresight the antenna with elements exhibits good separation between the co and cross-polar components showing that it has good linear polarisation. The antenna with a flat endcap has a much lower cross polar component indicating that it would also be acceptable if the device was working in a linear polarised application.

The co-polar elevation plot of Figure 5-13 shows a very clear distinction between the antennas with coned endcaps and those without. The coned results are poor. The other four results are similar in nature. The antenna with elements exhibits a slightly higher gain than the others, though this is cancelled out when the other three antennas have better performance at higher angles. The plain rod and flat endcap are very similar in the region of \( \pm 20^\circ \), but at angles around \( \pm 20-60^\circ \) the plain rod has higher gain but at extreme angles of greater than \( \pm 60^\circ \) the endcap antenna generally has a better performance.

The elevation cross-polar result of Figure 5-14 shows the antenna patterns to be closely spaced with less variation seen between the different designs. The antenna with elements has a higher value at larger angles and this offsets the poorer performance in the mid-range of angles. All four of the other antenna exhibit a similar performance but of note is that the coned endcap antenna performance shown in this plot is comparable in magnitude to the other traces and this has not been evident before. If axial ratios were calculated for the elevation co and cross-polar results they would be poor around the boresight of the antenna, but improve as the angle away from boresight increased.
As the antenna with elements has been shown to be of interest, the frequency sweep results for the linearly polarised antenna are presented in Figure 5-15 to Figure 5-18.

Figure 5-15 Azimuthal co-polar result for linearly polarised plain rod with elements

Figure 5-16 Azimuthal cross-polar result for linearly polarised plain rod with elements
Clearly the 1.7 and 1.8GHz frequencies are most suited to potential antenna applications based upon their patterns. These results do support the belief that the FSS is resonant between 1.7 and 1.8GHz.

FSS was placed around half the antenna circumference, as was done previously for the 9GHz antenna configuration and described in section 4.5. The results for this series of measurements are shown in Figure 5-19 to Figure 5-22. These results again
support the belief that the FSS is having the desired effect by showing that there is more power directed to the negative angles of the plots.

Figure 5-19 Azimuthal co-polar result for linearly polarised plain rod with 1/2 elements

Figure 5-20 Azimuthal cross-polar result for linearly polarised plain rod with 1/2 elements
Figure 5-21 Elevation co-polar result for linearly polarised plain rod with 1/2 elements

Figure 5-22 Elevation cross-polar result for linearly polarised plain rod with 1/2 elements

5.2.2.2 Circularly Polarised Results

The antenna was fed by inserting a phase shifter to add a 90° shift to one of the coaxial lines, in relation to the other. The results for this feeding mechanism for the plain rod antenna are shown below in Figure 5-23 to Figure 5-26, the rod with elements in Figure 5-31 to Figure 5-34 and the rod with half coverage of elements in
Figure 5-35 to Figure 5-38. The results were obtained using a dipole as a receiving antenna with the co and cross-polar components defined as described in section 5.1.4.

Figure 5-23 Azimuthal co-polar result for circularly polarised plain rod

Figure 5-24 Azimuthal cross-polar result for circularly polarised plain rod
There are subtle differences between the CP results and the linear results presented in section 5.2.2.1. The CP results for the azimuthal co-polar in Figure 5-23 do not show the offset of the main beam, seen in the linear polarisation measurement of Figure 5-7 at 1.8GHz. There is a noticeable difference in the powers of the 1.9 and 2.0GHz plots. The azimuthal cross-polar plot of Figure 5-24 has the spread of powers at boresight of only 5dB. Comparing the levels with the co-polar azimuthal there is approximately a 5dB variation between the co and cross-polar results. The co-polar elevation plot of Figure 5-25 shows that, when compared with Figure 5-26 the cross-
polar levels can be of the magnitude of 13dB different at boresight in the case of the 1.7GHz measurement.

Again, comparing the CP measurements at 1.8GHz, as done for the linear results, the four comparative plots are shown in Figure 5-27 to Figure 5-30.

Figure 5-27 Comparisons of different antenna parameters at 1.8GHz - Azimuthal co-polar CP

Figure 5-28 Comparisons of different antenna parameters at 1.8GHz - Azimuthal cross-polar CP
Elevation co-polar result of plain rod antenna comparison at 1.8GHz CP

Elevation cross-polar result of plain rod antenna comparison at 1.8GHz CP

Figure 5-29 Comparisons of different antenna parameters at 1.8GHz - Elevation co-polar CP

Figure 5-30 Comparisons of different antenna parameters at 1.8GHz - Elevation cross-polar CP

Taking into account the differences highlighted between the linear and CP plots direct comparison between the two is inappropriate and so comparison will be drawn, in this case, between the CP results only.

The plot of Figure 5-27 clearly shows that in the azimuthal plot that the coned endcap antenna radiates far less power than the other four antennas. The antenna with elements does give a broader pattern and this is particularly noticeable in the -40 to -
80° region when compared with the plain rod antenna. The antenna with an endcap has lobes at approximately ±60°, though these are below the power levels of both the plain rod and the rod with elements. Boresight gain for the plain rod, the antenna with elements and the flat endcap antenna is almost identical, being within 2dB of each other.

The plot of Figure 5-28 is notable in that the antenna with elements has the maximum boresight gain of the antennas presented. This, again, supports the belief that the FSS radiates in the 1.7-1.8GHz range. The supporting plot of the frequency scans of the CP antenna with elements are shown in Figure 5-31 to Figure 5-34 and support this conclusion. At angles greater than ±40° the patterns become similar in nature regardless of the design. The spread of antenna boresight gains is approximately 13dB and this corresponds to the value seen in the cross-polar elevation plot of Figure 5-30.

The co-polar elevation plot of Figure 5-29 shows clearly the lobes caused by the addition of the endcap at the 60° regions for both antennas with a flat endcap. Boresight gain is similar for three of the antennas. A noticeable result in this plot is the antenna with flat endcap and elements, although it has a lower boresight gain than the flat endcap antenna it does exhibit a desirable feature in the -40 to -50° and the 25 to 50° region. It appears that the addition of the elements, and hence the leaky wave action, has cancelled the nulls present in the antenna with just the flat endcap.

The cross-polar elevation plot of Figure 5-30 clearly shows that the addition of FSS elements has broadened the beam pattern. There is a slight offset in antenna maximum and it is probable that this has been caused by some antenna misalignment in the measurement. At boresight and at extreme angles the axial ratio of the antennas is fair. The addition of the phase shift should have caused the co and cross-polar antenna patterns to be similar. This result for the elevation plots is consistent with the azimuthal CP results.

Again, as for the linear results the plots for the antenna with elements and the antenna with half coverage of elements are presented in Figure 5-31 to Figure 5-34 for the antenna with elements and Figure 5-35 to Figure 5-38 for the half covered antenna.
Figure 5-31 Azimuthal co-polar result for circularly polarised plain rod with elements

Figure 5-32 Azimuthal cross-polar result for circularly polarised plain rod with elements
Figure 5-33 Elevation co-polar result for circularly polarised plain rod with elements

Figure 5-34 Elevation cross-polar result for circularly polarised plain rod with elements
Azimuthal co-polar result of antenna with 1/2 elements CP

Figure 5-35 Azimuthal co-polar result for circularly polarised plain rod with 1/2 elements

Azimuthal cross-polar result of antenna with 1/2 elements CP

Figure 5-36 Azimuthal cross-polar result for circularly polarised plain rod with 1/2 elements
5.3 Endcap-results

In an attempt to increase the suitability of beam shape for use in mobile telephone antennas, by radiating more power at wider angles, a brass endcap was considered. In its simplest the endcap consisted of a flat piece of brass sheet placed on the end of the dielectric rod. A more complex arrangement of a brass cone was tried, with varying
pitches, to see if that improved the antenna pattern when compared with the plain, flat endcap, or the plain rod.

Different cone pitches were simulated to study the effect they had, and to see if they improved the suitability of the radiation pattern. The area of PTFE that the coned endcap occupied was removed. The cone pitches investigated by simulation are shown in Figure 5-39. The FSS is left so this can be seen in relation to the tip of the cone also. The pitch that showed the most promising result when simulated was that shown as having a distance from the end of 116.15m. This is the cone pitch that has been measured.

![Figure 5-39 Overview of cone insertion](image)

Some results for the work done on antenna with endcaps antennas has been presented in the previous discussion of 5.2.2.1 and 5.2.2.2. Full sets of results for both types of endcap measured are presented in Appendix C Complete listing of 1.8GHz Results.

5.4 TM01 Results

The TM01 mode was excited in the TLM simulator by using a short coaxial probe through the end wall of the waveguide as described in Chapter 3. The E-field structure of the TM01 mode emerges from the centre of the waveguide in a radial fashion. This means that the E-field is always perpendicular on the surface of the FSS, no matter how the FSS is placed on the surface of the dielectric rod. By using the TM01 mode it is could be expected intuitively that a stronger leaky wave action will be observed and this effect will be better suited to applications where this sort of antenna pattern shape is required.
To work at the proposed UMTS satellite frequencies a CP field would be required. As discussed previously, a TM\textsubscript{01} field with CP cannot be excited easily within the waveguide, or for that matter in simulations. Linear excitation of the waveguide in the TM\textsubscript{01} region was simulated and the results looked acceptable for use in an antenna for satellite use, if a suitable feed mechanism were designed.

References

1 Centre For Mobile Communications Research, Department of Electronic and Electrical Engineering, Loughborough University, Leicestershire, England.
6 Conclusions

This work set out to develop techniques to manipulate antenna patterns of a dielectric rod antenna using a variety of techniques such as FSS elements and endcaps. The work has demonstrated that this is possible and that there are good reasons for this work to be pursued further as it improves the performance of antennas in common use in mobile communications.

Taking the main topics of discussion in Chapter 3, Chapter 4 and Chapter 5 conclusions will be drawn on each one individually in sections 6.1 to 6.3 respectively. General conclusions, applicable to all antennas, and ideas for further work will then be discussed in sections 6.4 and 6.5 respectively.

6.1 Feeding Techniques

Various techniques were investigated to feed the antenna. The feed design that gave the largest measured wide-band response, desirable for maximum coupling of power across the waveguide mono-mode region, was the coaxial cable fed antenna that had results presented in 3.2. This proved impractical for providing a CP fed antenna, and other, alternative feeds were investigated.

Slot coupling, utilising microstrip line, was identified as a possible solution as this worked successfully as a feed for the 9GHz antenna. Efforts were made to use some of the techniques used to feed Dielectric Resonator (DR) antennas, such as crossed slots, for the antenna detailed in Chapter 4 to generate CP fields. Creating the additional phase shift required for CP operation of the antenna was attempted with increased lengths of microstrip line on one of the feed branches. Modelled results of the return loss, shown in Figure 3-6, looked promising but on further investigation, including construction of the feed, the nature of the CP could not be verified and this led to the conclusion that this method needed better confirmation or other methods needed to be explored.

It was for this reason that a decision was taken to use commercially available splitters and phase shifters, as this would provide some degree of control of the phase offset and remove the need for complicated microstrip line networks. It had been shown
that the cross slot coupling was a viable way of proceeding, but the delivery of the CP field to the slot with the microstrip, and hence the waveguide, was the point at which the technique became unreliable. A further consideration was the amount of radiation from the back of the antenna and, in an effort to minimise this, coaxial cable was investigated as a possible feeding mechanism. The only radiation at the back of the antenna, other than fringing fields around the ground plane would be at the coaxial cable tip and the slot, not the whole length of the line which would be the case if microstrip were used. Due to their complexities these commercial components could not be easily modelled using the simulation software and the final return loss when theses devices were present could not be accurately predicted. Modelling of the slots did show that where possible the simulation software was capable of accurately modelling the antenna and feed and the plots shown in Figure 3-12 demonstrate good agreement between the simulated and measured results. When both feeds are measured, Figure 3-13, the response is seen to be similar in nature and the differences between feeds are small. It is not until the feed has the splitter added that a significant deterioration is seen in the frequency response, even when the other feed is matched, as shown in Figure 3-14. The final feed with the splitter only, Figure 3-15, shows that the feed frequency response is less than desirable at 1.85GHz, but acceptable elsewhere within the mono-mode region of waveguide operation. Once the phase shifter is added, Figure 3-17, the response changes again drastically and is good between 1.7 and 1.8GHz, poor at 1.85GHz and acceptable at 2GHz.

6.2 Prototype Design of the new antenna at 9GHz

The primary aim of the work in this chapter was to carry out a parametric study to determine the suitability of the various FSS element types on the performance of the dielectric rod antenna. Three of the most common types of element were tried, dipole, crossed-dipole and square loop. Several designs of each element were designed that were resonant at 9GHz and these were manufactured and measured. Not only were several designs of each element tried, but the positioning of the elements was also varied. Position of the elements around the rod, in relation to the electric field was investigated. The conclusions that can be drawn from the measurements presented in sections 4.2 to 4.4 are summarised below.
The dipole elements had little effect on the measured radiation patterns and very little difference is noted between the antenna with the dipole element and the one without. Cross-polar levels of the antenna with elements at 0° are slightly raised if Figure 4-4 and Figure 4-6 are studied.

Crossed-dipole elements appear to have more of an effect on the radiation pattern if the radial measurements of Figure 4-8 and Figure 4-11 are studied. There is more variation in pattern than when the dipoles were present on the rod antenna. When either design of elements is placed at 0° there is a noticeable improvement in boresight gain of the azimuthal and elevation cut patterns. This is desired since it shows that the elements are resonant at 9GHz and there is improved guiding of energy along the rod. Cross-polar performance does deteriorate, and the level rises, but it is still typically 20dB lower than the co-polar value.

Square-loop resonators appear to resonate slightly away from the desired frequency of operation, 9GHz. This is supported by the fact that the radial measurements of Figure 4-20 to Figure 4-23 have many of the ‘nulls’ present in the plain rod antenna filled in due to leaky wave action. This is also supported by the fact that the gains at the antenna boresight are at slightly lower levels when compared to the plain rod antenna. Figure 4-26 does show that leakage is occurring with the plot showing that the antenna with elements has a broader main beam than the plain rod antenna.

In summary, when the elements are placed at 0° there appears to be most interaction between them and the fields within the rod antenna.

Cross-polar performance does deteriorate, but not to such an extent that the antenna performance would be significantly affected, when the elements are placed at 0°. Cross-polar performance when the elements are at 90° is very similar to the plain rod antenna.

When elements were placed around half of the rod antenna the shift in main beam was quite dramatic. The results presented in the section 4.5 show excellent agreement with the measured results. It has been argued that the beam shift away from the elements is due to the elements being resonant and this was confirmed by representing
the elements as a metal sheet. At frequencies away from 9GHz there is a leaky wave action and this is clearly seen in the plot of Figure 4-38, when compared with Figure 4-36, as the broader antenna pattern.

It was on the basis of these results that the decision to proceed on future designs with square-looped resonators was made.

6.3 Prototype Design of the new antenna at 1.8GHz

The aim of the work carried out in this chapter was to establish the effect and suitability of an endcap on this particular type of antenna design and to generate CP fields within the guide to determine the antenna performance for use in satellite communications.

Care was taken to ensure that the results gained from the work in Chapter 4 were implemented, such as element position on the rod to align the elements with the resultant electric field to achieve maximum benefit from the elements.

6.3.1 Linearly polarised results

The plots of Figure 5-15 to Figure 5-18 certainly show that the antenna is capable of broadening the beam, if compared to the equivalent plain rod, though it is hard to say whether resonance of the FSS is at 1.7 or 1.8GHz. Limitations in the dipole receiver mean that direct comparison is difficult, due to the frequency dependant nature of the dipole response, but it does appear that at 1.9 and 2.0GHz the antenna pattern does not exhibit the main beams seen at the other frequencies. This would seem to indicate that there is more even distribution of power over a wide range of angles, due to the leaky wave action of the elements. Figure 5-17 illustrates this conclusion nicely as does Figure 5-15. Due to the nature of the resultant electric fields having a vertical and horizontal component the co and cross-polar plots of the 1.8GHz antenna will not show the marked difference seen in the co and cross-polar results of the 9GHz antenna.

The endcap does not have as much effect in the broadening of the beam as would be hoped for. The endcap, as suggested by the simulated results, introduces nulls at approximately ±50°. The antenna with elements tends not to suffer from these and it
can be argued that if a choice had to be made between using elements and an endcap to improve the antenna performance, elements would be preferred.

6.3.2 Circularly polarised results

The CP results support the fact that the FSS is working at the desired frequency of 1.8GHz having boresight gain values equal, or higher, than those of the plain rod antenna. The addition of the elements has widened the antenna pattern when the plain rod plot of Figure 5-23 is compared with that of Figure 5-31 for the antenna with elements.

To better categorise the nature of the CP a measurement facility capable of rotating the receiver antenna with the rotating E-field is required. No such facility was available, so it was not possible for a full analysis of the CP fields radiated from the antenna to be undertaken.

6.4 General conclusions

It has been shown in this work that this type of antenna could be used in the intended applications described in Chapter 1, if certain refinements, which are detailed below in section 6.5, are undertaken.

This work has achieved the following goals:

- The effect of FSS element type on this antenna has been investigated
- Two prototype antennas, working at 1.8GHz and 9GHz, have been constructed and tested
- Endcap configurations have been investigated
- Linear and CP fields have been excited within the waveguide and measurements performed on the antenna to characterise its performance.

6.5 Future work

This work has investigated several parameters pertinent to the performance of this antenna. As work has progressed problems that did not require immediate investigation have been encountered and some of these are detailed below along with ideas to progress this work further towards the eventual aim of having an antenna that could be used in modern mobile communication devices.
• Improvement in feeding the antenna

The antenna feeds that have been utilised are quite primitive in their design and construction. The use of coaxial cables is preferred due to their non-frequency dependant characteristics. Slots are frequency dependent and allow radiation to leak out behind the antenna. They also require a ground plane, which is not ideal if it has to extend past the rear wall of the waveguide. If coaxial cables could be arranged in such a way through the sidewall of the waveguide to generate CP a better performance could be expected.

• Miniaturisation

The antenna currently uses a material with an $\varepsilon_r \approx 2$. Further miniaturisation is possible with higher dielectric constant materials. This material would increase the dielectric losses and provide a great challenge as the accuracy of placing feeds and error on the element design and manufacture could become crucial factors in the antenna performance.

• Reduction in Specific Absorption Rate (SAR)

SAR is a measure of the power that is absorbed by the user of devices radiating electromagnetic energy. In recent years concern has been expressed about the levels of SAR people are exposed to and the possible health effects this may have. With clever element position and design it may be possible for this to be reduced. Resonant elements could be placed on the side of the antenna nearest the head to provide shielding to the head whilst elements of a slightly different dimension are placed on the opposite side of the antenna to aid the antenna performance utilising leaky wave action.

• CP field generation using the TM_{01} mode

The TM_{01} mode could be useful for certain parts of the UMTS spectrum. Ways to generate a CP field within this mode could be investigated to see if there is benefit in utilising this particular waveguide mode in conjunction with the TE_{11} mode to increase the frequency range that this type of antenna could operate at. This would
require several feeding structures for each mode or one carefully designed wide band feed.
Appendix A Derivation of Circular Waveguide Equations

Starting from Maxwell’s Equations the components of waveguide fields and properties are derived in circular co-ordinates as follows.

\[ \nabla \times \vec{E} = -j\omega\mu\vec{H} \]  
(A.1)

\[ \nabla \times \vec{H} = j\omega\varepsilon\vec{E} \]  
(A.2)

Assuming a time and z dependence of \( e^{(i\omega t-z)} \) in (A.1) a modified curl operation can be written for \( \vec{E} \) such that:

\[
\text{Curl} \vec{E} = \frac{1}{\rho} \begin{vmatrix}
I & \rho J & K \\
\frac{\partial}{\partial \rho} & \frac{\partial}{\partial \phi} & \pm \gamma \\
E_\rho & \rho E_\phi & E_z
\end{vmatrix} = -j\omega\mu H
\]  
(A.3)

This can be expanded in to the following terms:

\[ \frac{\partial}{\partial \phi} E_z + \rho E_\gamma = -j \rho \omega \mu H_\rho \]  
(A.4)

\[ -\frac{\partial}{\partial \rho} E_z - \gamma E_\rho = -j \omega \mu H_\phi \]  
(A.5)

\[ \frac{\partial}{\partial \rho} \rho E_\phi - \frac{\partial}{\partial \phi} E_\rho = -j \omega \rho \mu H_z \]  
(A.6)

And for (A.2)as:

\[ \frac{\partial}{\partial \phi} H_\phi + \rho H_\gamma = j \rho \omega \varepsilon E_\rho \]  
(A.7)

\[ -\frac{\partial}{\partial \rho} H_\phi - \gamma H_\rho = j \omega \varepsilon E_\phi \]  
(A.8)

\[ \frac{\partial}{\partial \rho} \rho H_\rho - \frac{\partial}{\partial \phi} H_\phi = j \omega \rho \varepsilon E_z \]  
(A.9)

Multiplication of (A.8) by \( \frac{1}{\gamma} \) and substitution of \( H_\rho \) into (A.4) leads to:

\[ \frac{\partial}{\partial \phi} E_z + \rho E_\phi = -j \rho \omega \mu \left( -j \omega \varepsilon \frac{\partial H_\rho}{\partial \rho} - \frac{1}{\gamma} \frac{\partial H_\phi}{\partial \phi} \right) \]  
(A.10)

Rearranging leads to:

\[ E_\phi (\gamma^2 - \omega^2 \mu \varepsilon) = j \omega \mu \frac{\partial H_\rho}{\partial \rho} - \frac{\gamma}{\rho} \frac{\partial E_z}{\partial \phi} \]  
(A.11)
Where \((\gamma^2 - \omega^2 \mu \varepsilon) = k_c^2\) and the term \(\omega^2 \mu \varepsilon\) is a constant, \(k^2\), known as the wave number or propagation constant of the medium. Assuming there are no losses in the material such that \(\gamma = \alpha + j \beta\) can be written as \(j \beta\) for forward going propagation (A.11) finally leads to:

\[
E_\phi = -j \left( \frac{\beta}{k_c^2} \left( \frac{\partial E_z}{\partial r} - \omega \mu \frac{\partial H_\theta}{\partial \phi} \right) \right)
\]

(A.12)

Similar expressions for \(E_\rho\), \(H_\rho\) and \(H_\phi\) can be found using the same process to give:

\[
E_\rho = -j \left( \frac{\beta}{k_c^2} \left( \frac{\partial E_z}{\partial r} + \omega \mu \frac{\partial H_\theta}{\partial \phi} \right) \right)
\]

(A.13)

\[
H_\rho = \frac{j}{k_c^2} \left( \omega \mu \frac{\partial E_z}{\partial r} - \beta \frac{\partial H_\theta}{\partial \phi} \right)
\]

(A.14)

\[
H_\phi = -j \left( \omega \mu \frac{\partial E_z}{\partial r} + \beta \frac{\partial H_\theta}{\partial \phi} \right)
\]

(A.15)

Since the total electric and magnetic intensities in the charge free regions between the conducting boundaries must also satisfy wave equation, which, for the TE case is where \(E_z = 0\) and \(H_z\) is a solution such that:

\[
\nabla^2 \vec{H} = -k_c^2 \vec{H}
\]

(A.16)

Which, when broken into parts is equal to:

\[
\nabla^2 \vec{H} = \nabla^2 E_z + \frac{\partial^2 H_\theta}{\partial z^2}
\]

(A.17)

Where \(\frac{\partial^2 \vec{H}}{\partial z^2} = \gamma^2 \vec{H} = -k_c^2 H_z\) such that when expanded in cylindrical co-ordinates it becomes:

\[
\frac{\partial^2 H_z}{\partial \rho^2} + \frac{1}{\rho} \frac{\partial H_z}{\partial \rho} + \frac{1}{\rho^2} \frac{\partial^2 H_z}{\partial \phi^2} = -k_c^2 H_z
\]

(A.18)

Equation (A.18) is a function of two variables and, if the method of separation of variables is used, can be written as \(H_\rho = R(\rho)F(\phi)\) which can then be substituted to obtain two ordinary differential equations. Here \(R\) is a function of \(\rho\) alone and \(F\) is a function of \(\phi\) alone and, after multiplying through by \(\rho^2\) (A.18), becomes:

\[
\frac{\rho^2 \frac{d^2 R}{d \rho^2} + \rho \frac{d R}{d \rho} + \rho^2 k_c^2}{F} = -\frac{1}{F} \frac{d^2 F}{d \phi^2}
\]

(A.19)
Of which each side is only dependent on $\rho$ or $\phi$. For this equation to hold each side must be equal to a constant, $\nu^2$, such that:

$$\frac{d^2 F}{d\phi^2} + \nu^2 F = 0$$  \hfill (A.20)

Or:

$$\rho^2 \frac{d^2 R}{d\rho^2} + \rho \frac{dR}{d\rho} + (\rho^2 k_e^2 - \nu^2) R = 0$$  \hfill (A.21)

The general solution of (A.20) is:

$$F_{\rho} = A\sin \nu \phi + B\cos \nu \phi$$  \hfill (A.22)

And since this function must be periodic, such that $\phi \pm 2m\pi$ holds, $\nu$ must be an integer $n$.

Equation (A.21) is in a form that satisfies Bessel's equation and the general solution of which is:

$$R_{\rho} = C J_n(k_e \rho) + D Y_n(k_e \rho)$$  \hfill (A.23)

At $\rho = 0$, $Y_n(k_e \rho)$ becomes infinite and so $D$ can equal zero.

Therefore:

$$H_{\rho} = \left( A\sin \nu \phi + B\cos \nu \phi \right) J_n(k_e \rho)$$  \hfill (A.24)

Constraints can also be placed such that the tangential E-field must be zero on the waveguide wall ($\rho = a$) and that $E_{\rho}$ must be zero for TE fields. Substituting (A.24) into (A.12) leads to:

$$E_{\rho}(\rho, \phi, z) = \frac{\epsilon_0 \mu_0}{k_e} \left( A\sin \nu \phi + B\cos \nu \phi \right) J_n'(k_e \rho) e^{-jk_0 z}$$  \hfill (A.25)

Where $J_n'(k_e \rho)$ is the derivative of $J_n$ with respect to its argument. For $E_{\rho}$ to vanish at the wall of the waveguide, ($\rho = a$), $J_n'(ka)$ must be equal to zero.

The roots of $J_n'(x)$ are defined as $p_{nm}'$, so that $J_n'(p_{nm}') = 0$, where $p_{nm}'$ is the $m^{th}$ root of $J_n'$. Therefore $k_e$ must have the value:

$$k_{con} = \frac{p_{nm}'}{a}$$  \hfill (A.26)

Values of $p_{nm}'$ are tabulated in a variety of places such as Balanis. 1
The TM mode derivation is similar and (A.24) can be rewritten replacing $H_z$ with $E_z$.

Boundary conditions at $\rho = a$ are such that:

$$E_z(\rho, \phi) = 0$$  \hspace{1cm} (A.27)

And therefore:

$$J_n(k_c \rho) = 0$$  \hspace{1cm} (A.28)

Leading to:

$$k_c = \frac{p_{nm}}{a}$$  \hspace{1cm} (A.29)

Where $p_{nm}$ is the $m^{th}$ root of $J_n(x)$ such that $J_n(p_{nm}) = 0$.

Values of $p_{nm}$ are tabulated in a variety of places such as Balanis¹.

References

Appendix B Derivation of Waveguide Losses

Referring to Maxwell's equations and including a term for the conduction current as well as the displacement current one can write:

\[ \nabla \times H = \sigma E + j \omega \varepsilon E = j \omega \left( \varepsilon - j \frac{\sigma}{\omega} \right) E \]  \hspace{1cm} (B.1)

Where the complex permittivity is defined as \( \varepsilon = \varepsilon' - j \varepsilon'' \).

The angle formed in the complex plane of the tangent of the real and imaginary parts of \( \varepsilon \) is defined as the loss tangent.

\[ \tan|\delta| = \frac{\sigma}{\omega \varepsilon} \]  \hspace{1cm} (B.2)

Ramo, Whinnery and Van Duzer\(^1\) show that the complex permittivity can be expressed as:

\[ \varepsilon = \varepsilon' - j \varepsilon'' \]  \hspace{1cm} (B.3)

Substituting this value of \( \varepsilon \) into (B.1), since it has already been defined leads to:

\[ \nabla \times H = j \omega \varepsilon E = j \omega (\varepsilon' - j \varepsilon'') E = \omega \varepsilon'' E + j \omega \varepsilon' E \]  \hspace{1cm} (B.4)

There is no need for the conduction current \( J = \sigma E \) since \( \omega \varepsilon'' E \) replaces it. Therefore \( \omega \varepsilon'' \) is equivalent to \( \sigma \) in a lossy dielectric. Also, \( \varepsilon' \) is equivalent to the real term \( \varepsilon \), which is equal to \( \varepsilon \omega \varepsilon \).

\[ \varepsilon = \varepsilon' \quad \sigma = \omega \varepsilon'' \]  \hspace{1cm} (B.5)

Therefore \( \varepsilon'' \) describes all loss mechanisms in a dielectric at a specified frequency. This allows (B.2) to be written as:

\[ \tan|\delta| = \frac{\varepsilon''}{\varepsilon'} \]  \hspace{1cm} (B.6)

It has been stated in Appendix A Derivation of Circular Waveguide Equations that:

\[ \gamma = \alpha + j \beta = \sqrt{k_e^2 - k^2} \]  \hspace{1cm} (B.7)

Rewriting this, taking into account the complex permittivity terms expressed in (B.3) leads to:

\[ \gamma = \sqrt{k_e^2 - \omega^2 \mu_0 \varepsilon_0 \varepsilon_r (1 - j \tan \delta)} \]  \hspace{1cm} (B.8)
Since most dielectric materials have a small $\tan \delta$ value ($\tan \delta \ll 1$) the expression can be simplified using the first couple of terms from Taylor’s expansion.

$$\sqrt{a^2 + x^2} = a + \frac{1}{2} \left( \frac{x^2}{a} \right) \quad \text{for} \ x \ll a \quad \text{(B.9)}$$

Rewriting (B.8) leads to:

$$\gamma = \sqrt{k_c^2 - k^2} + jk^2 \tan \delta \quad \text{(B.10)}$$

Applying Taylor’s expansion of (B.9) to (B.10) one can write:

$$\gamma = \sqrt{k_c^2 - k^2} + \frac{jk^2 \tan \delta}{2\sqrt{k_c^2 - k^2}} \quad \text{(B.11)}$$

Which is equal to:

$$\frac{k^2 \tan \delta}{2\beta} + j\beta \quad \text{(B.12)}$$

So, for material with low loss characteristics the phase constant is unchanged and the loss associated with the dielectric, $\alpha_d$, is given by the first term in (B.12). The units for $\alpha_d$ are Np/m.

References

Appendix C Complete listing of 1.8GHz Results

Results for the plain rod are shown in Figure C-1 to Figure C-4, the rod with elements in Figure C-5 to Figure C-8 and the rod with half coverage of elements in Figure C-9 to Figure C-12.

Figure C-1 Azimuthal co-polar result for linearly polarised plain rod

Figure C-2 Azimuthal cross-polar result for linearly polarised plain rod
Figure C-3 Elevation co-polar result for linearly polarised plain rod

Figure C-4 Elevation cross-polar result for linearly polarised plain rod
Figure C-5 Azimuthal co-polar result for linearly polarised plain rod with elements

Figure C-6 Azimuthal cross-polar result for linearly polarised plain rod with elements
Figure C-7 Elevation co-polar result for linearly polarised plain rod with elements

Figure C-8 Elevation cross-polar result for linearly polarised plain rod with elements
Figure C-9 Azimuthal co-polar result for linearly polarised plain rod with 1/2 elements

Figure C-10 Azimuthal cross-polar result for linearly polarised plain rod with 1/2 elements
Figure C-11 Elevation co-polar result for linearly polarised plain rod with 1/2 elements

Figure C-12 Elevation cross-polar result for linearly polarised plain rod with 1/2 elements

The results for the CP feeding mechanism for the plain rod antenna are shown below in Figure C-13 to Figure C-16, the rod with elements in Figure C-17 to Figure C-20 and the rod with half coverage of elements in Figure C-21 to Figure C-24.
Figure C-13 Azimuthal co-polar result for circularly polarised plain rod

Figure C-14 Azimuthal cross-polar result for circularly polarised plain rod
Figure C-15 Elevation co-polar result for circularly polarised plain rod

Elevation co-polar result of plain rod antenna CP

Figure C-16 Elevation cross-polar result for circularly polarised plain rod

Elevation cross polar result of plain rod antenna CP
Figure C-17 Azimuthal co-polar result for circularly polarised plain rod with elements

Figure C-18 Azimuthal cross-polar result for circularly polarised plain rod with elements
Figure C-19 Elevation co-polar result for circularly polarised plain rod with elements

Figure C-20 Elevation cross-polar result for circularly polarised plain rod with elements
Azimuthal co-polar result of antenna with 1/2 elements CP

Figure C-21 Azimuthal co-polar result for circularly polarised plain rod with 1/2 elements

Azimuthal cross-polar result of antenna with 1/2 elements CP

Figure C-22 Azimuthal cross-polar result for circularly polarised plain rod with 1/2 elements
Figure C-23 Elevation co-polar result for circularly polarised plain rod with 1/2 elements

Figure C-24 Elevation cross-polar result for circularly polarised plain rod with 1/2 elements

Results for the plain rod with a flat endcap are shown in Figure C-25 to Figure C-28, the rod with elements in Figure C-29 to Figure C-32 and the rod with half coverage of elements in Figure C-33 to Figure C-36.
Figure C-25 Azimuthal co-polar result for linearly polarised plain rod with a flat endcap

Figure C-26 Azimuthal cross-polar result for linearly polarised plain rod with a flat endcap
Figure C-27 Elevation co-polar result for linearly polarised plain rod with a flat endcap

Figure C-28 Elevation cross-polar result for linearly polarised plain rod with a flat endcap
Figure C-29 Azimuthal co-polar result for linearly polarised plain rod with a flat endcap and elements

Figure C-30 Azimuthal cross-polar result for linearly polarised plain rod with a flat endcap and elements
Figure C-31 Elevation co-polar result for linearly polarised plain rod with a flat endcap and elements

Figure C-32 Elevation cross-polar result for linearly polarised plain rod with a flat endcap and elements
Figure C-33 Azimuthal co-polar result for linearly polarised plain rod with a flat endcap and 1/2 elements

Figure C-34 Azimuthal cross-polar result for linearly polarised plain rod with a flat endcap and 1/2 elements
Figure C-35 Elevation co-polar result for linearly polarised plain rod with a flat endcap and 1/2 elements

Figure C-36 Elevation cross-polar result for linearly polarised plain rod with a flat endcap and 1/2 elements

The results for the CP feeding mechanism for the plain rod antenna with a flat endcap are shown below in Figure C-37 to Figure C-40, the rod with elements in Figure C-41 to Figure C-44 and the rod with half coverage of elements in Figure C-45 to Figure C-48.
Figure C-37 Azimuthal co-polar result for circularly polarised plain rod with a flat endcap

Figure C-38 Azimuthal cross-polar result for circularly polarised plain rod with a flat endcap
Figure C-39 Elevation co-polar result for circularly polarised plain rod with a flat endcap

Figure C-40 Elevation cross-polar result for circularly polarised plain rod with a flat endcap
Azimuthal co-polar result of antenna with flat endcap and elements CP

Figure C-41 Azimuthal co-polar result for circularly polarised plain rod with a flat endcap and elements

Azimuthal cross polar result of antenna with flat endcap and elements CP

Figure C-42 Azimuthal cross-polar result for circularly polarised plain rod with a flat endcap and elements
Figure C-43 Elevation co-polar result for circularly polarised plain rod with a flat endcap and elements

Figure C-44 Elevation cross-polar result for circularly polarised plain rod with a flat endcap and elements
Figure C-45 Azimuthal co-polar result for circularly polarised plain rod with a flat endcap and 1/2 elements

Figure C-46 Azimuthal cross-polar result for circularly polarised plain rod with a flat endcap and 1/2 elements
Elevation co-polar result of antenna with flat endcap and 1/2 elements CP

Figure C-47 Elevation co-polar result for circularly polarised plain rod with a flat endcap and 1/2 elements

Elevation cross-polar result of antenna with flat endcap and 1/2 elements CP

Figure C-48 Elevation cross-polar result for circularly polarised plain rod with a flat endcap and 1/2 elements

Results for the plain rod with a coned endcap are shown in Figure C-49 to Figure C-52, the rod with elements in Figure C-53 to Figure C-56 and the rod with half coverage of elements in Figure C-57 to Figure C-60.
Figure C-49 Azimuthal co-polar result for linearly polarised plain rod with a coned endcap

Figure C-50 Azimuthal cross-polar result for linearly polarised plain rod with a coned endcap
Figure C-51 Elevation co-polar result for linearly polarised plain rod with a coned endcap

Figure C-52 Elevation cross-polar result for linearly polarised plain rod with a coned endcap
Figure C-53 Azimuthal co-polar result for linearly polarised plain rod with a coned endcap and elements

Figure C-54 Azimuthal cross-polar result for linearly polarised plain rod with a coned endcap and elements
Figure C-55 Elevation co-polar result for linearly polarised plain rod with a coned endcap and elements

Figure C-56 Elevation cross-polar result for linearly polarised plain rod with a coned endcap and elements
Azimuthal co-polar result of antenna with coned endcap and 1/2 elements

Figure C-57 Azimuthal co-polar result for linearly polarised plain rod with a coned endcap and 1/2 elements

Azimuthal cross-polar result of antenna with coned endcap and 1/2 elements

Figure C-58 Azimuthal cross-polar result for linearly polarised plain rod with a coned endcap and 1/2 elements
Figure C-59 Elevation co-polar result for linearly polarised plain rod with a coned endcap and 1/2 elements

Figure C-60 Elevation cross-polar result for linearly polarised plain rod with a coned endcap and 1/2 elements

The results for the CP feeding mechanism for the plain rod antenna with a coned endcap are shown below in Figure C-61 to Figure C-64, the rod with elements in Figure C-65 to Figure C-68 and the rod with half coverage of elements in Figure C-69 to Figure C-72.
Figure C-61 Azimuthal co-polar result for circularly polarised plain rod with a coned endcap

Figure C-62 Azimuthal cross-polar result for circularly polarised plain rod with a coned endcap
Figure C-63 Elevation co-polar result for circularly polarised plain rod with a coned endcap

Figure C-64 Elevation cross-polar result for circularly polarised plain rod with a coned endcap
Figure C-65 Azimuthal co-polar result for circularly polarised plain rod with a coned endcap and elements

Figure C-66 Azimuthal cross-polar result for circularly polarised plain rod with a coned endcap and elements
Figure C-67 Elevation co-polar result for circularly polarised plain rod with a coned endcap and elements

Figure C-68 Elevation cross-polar result for circularly polarised plain rod with a coned endcap and elements
Figure C-69 Azimuthal co-polar result for circularly polarised plain rod with a coned endcap and 1/2 elements

Figure C-70 Azimuthal cross-polar result for circularly polarised plain rod with a coned endcap and 1/2 elements
Figure C-71 Elevation co-polar result for circularly polarised plain rod with a coned endcap and 1/2 elements

Figure C-72 Elevation cross-polar result for circularly polarised plain rod with a coned endcap and 1/2 elements