ELECTROMAGNETICALLY TRANSPARENT FEED NETWORKS FOR ANTENNA ARRAYS

DISSERTATION

Presented in Partial Fulfillment of the Requirements for the Degree Doctor of Philosophy in the Graduate School of The Ohio State University

By

Eugene Yi-Chien Lee, B.S., M.S.

* * * * *

The Ohio State University

2008

Dissertation Committee:

Roberto G. Rojas, Adviser
Eric K. Walton
Prabhakar H. Pathak
Randy Moses

Approved by

Adviser
Graduate Program in Electrical Engineering
© Copyright by

Eugene Yi-Chien Lee

2008
ABSTRACT

Multiple antennas underneath a single radome are attractive for platforms with limited space. Blockage scenarios are created when multiple antennas operate under a single radome. To avoid blockage, antennas that are electromagnetically transparent at frequencies outside of their respective radiating bandwidth are required. Although a number of structures in an antenna can contribute to the blockage, this study focuses on the feed networks.

Electromagnetic simulations are used to examine the scattering behavior of the feed networks obstructing a radiating source. Corporate feed network examples are investigated and were found to have poor transparency below the design frequency ($T_{\theta,\phi} < -1$ dB for $f < 1.3 f_0$). Feed networks designs are presented with the aim of electromagnetic transparency. Some designs provide unequal phase to a circularly polarized antenna array and rely on a far field phase matching method. This method rotates each array element according to the prior calculated phase output. The pseudo-random orientation of the array elements results in an improved antenna array radiating axial ratio ($AR = -2.18$ to -1). Right hand circularly polarized (RHCP) radiation is preserved when blocked by these feed networks.

Feed networks have been designed using optimization methods and classical design techniques. A feed network incorporating several of these techniques has been built and measured for L and S band antenna arrays in a multi-band antenna system. The
L band array blocks the S band array line of sight under certain pointing angles. An X band antenna is positioned underneath the L and S band antennas. The L band array attenuates the peak gain of the S band array but preserves the beamwidth. This attenuation is limited to the main beam and results in the sidelobe levels increasing with respect to the main beam under blockage. These antenna arrays yield RHCP radiation with an axial ratio better than -1.3 over the operating bandwidth and half power beamwidth, regardless of blockage conditions. The L and S band antennas provide approximately 1 dB of attenuation at X band while preserving the RHCP polarization.
This is dedicated to the memory of my father,

Dr. Yanien Lee, Ph.D.
ACKNOWLEDGMENTS

I would like to thank Bruce Montgomery, Gary Bruce, and the rest of Syntonics, LLC for giving me the opportunity to work on the multi-band antenna project. I would like to thank my research supervisor Eric Walton, my advisor Roberto Rojas, and the rest of my dissertation committee. I would like to acknowledge the help of Dana Kohlgraf and Ryan Pavlovicz who worked on the multi-band antenna. Ed Newman and Frank Paynter both provided valuable assistance with the Electromagnetic Surface Patch code.

I want to thank my friends and family for the encouragement over the years. In particular, I would like to acknowledge my parents and sisters for their support through this endeavor.
VITA

May 31, 1978 .......................... Born - Silver Spring, MD, USA

2000 .............................. B.S. in Electrical Engineering
University of Rochester

2002 .............................. M.S. in Electrical Engineering
The Ohio State University

2000-2002 .......................... Graduate Research Associate,
The ElectrocScience Laboratory at The
Ohio State University.

2002-2004 .......................... System Engineer,
Raytheon.

2004-present ........................ Antenna Engineer
Syntomics, LLC

2004-present ........................ Graduate Student
The Ohio State University

PUBLICATIONS

Research Publications


E. Lee, "Integration and testing of a transmission line system for an electromagnetically transparent antenna array," in 28th Annual Symp. of the Antenna Measurement Techniques Association, Austin, TX, Nov. 2006.


FIELDS OF STUDY

Major Field: Electrical Engineering

Studies in Electromagnetics and Antennas: Roberto Rojas-Teran
# TABLE OF CONTENTS

Abstract ................................................................. ii  
Dedication ................................................................. iv  
Acknowledgments ......................................................... v  
Vita ............................................................... vi  
List of Tables ......................................................... xi  
List of Figures ......................................................... xii  

Chapters:  

1. Introduction ......................................................... 1  
   1.1 Problem Description .............................................. 1  
   1.2 Problem Domain .................................................. 3  
      1.2.1 FSS ......................................................... 3  
      1.2.2 Radome Techniques ........................................ 4  
      1.2.3 Feed networks .............................................. 4  
   1.3 Dissertation Outline ........................................... 5  

2. Design Criteria and Evaluation ................................. 7  
   2.1 Numerical simulation ........................................... 7  
   2.2 Figures of Merit ................................................ 8  
      2.2.1 Simulation Test Setup .................................... 9  
      2.2.2 Transmissivity ........................................... 11  
      2.2.3 Polarization ............................................... 19  
   2.3 Transparency techniques ..................................... 20
3. Corporate feed networks ........................................... 22
   3.1 PEC Ground ......................................................... 22
   3.2 Parallel-Strip Circuits .......................................... 29
   3.3 Corporate Feed as a Frequency Selective Surface ................. 36
   3.4 Summary .......................................................... 44

4. Feed Network Geometry ............................................ 45
   4.1 Vertical orientation ................................................. 46
   4.2 Feed Network Layout ............................................... 48
      4.2.1 Unequal phase compensation ...................................... 49
      4.2.2 Series Feed Network ............................................ 54
   4.3 Rotational Symmetry ................................................ 60
   4.4 Final Comments ..................................................... 62

5. Feed Network Designs .............................................. 66
   5.1 Corporate FSS Optimization ....................................... 66
      5.1.1 Algorithm ...................................................... 67
      5.1.2 Variables ........................................................ 68
      5.1.3 Cost function ................................................... 70
      5.1.4 Results .......................................................... 77
   5.2 Non-Optimized Feed Network Design .............................. 80
      5.2.1 Feed Network Transparency Techniques ......................... 84
      5.2.2 Non-optimized Feed Network Simulations ....................... 87
   5.3 Comments ........................................................... 91

6. Tri-band Antenna Example ......................................... 92
   6.1 Multi-band Antenna Design ....................................... 92
      6.1.1 Array and FSS Design ......................................... 95
      6.1.2 Feed Network Design ......................................... 100
      6.1.3 Antenna Measurements ....................................... 104
   6.2 Summary .......................................................... 110

7. Summary and Conclusions .......................................... 115
   7.1 Results .......................................................... 115
   7.2 Further work ...................................................... 117

Appendices:
   ix
# LIST OF TABLES

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.1 Antenna Parameters with and without Phase Compensation</td>
<td>52</td>
</tr>
<tr>
<td>5.1 Dipole Array Parameters</td>
<td>69</td>
</tr>
<tr>
<td>5.2 Optimization Results</td>
<td>77</td>
</tr>
<tr>
<td>6.1 Coupling (dB) for L band quarter panel</td>
<td>102</td>
</tr>
<tr>
<td>6.2 Coupling (dB) for S band quarter panel</td>
<td>102</td>
</tr>
<tr>
<td>A.1 Coupling ratios for a 6 element main line</td>
<td>120</td>
</tr>
<tr>
<td>A.2 Coupling ratios for a 5 element branch line</td>
<td>121</td>
</tr>
<tr>
<td>A.3 Coupling ratios for a 30 element Polygonal Array</td>
<td>123</td>
</tr>
</tbody>
</table>
# List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>Multiple antennas under a single radome</td>
<td>2</td>
</tr>
<tr>
<td>1.2</td>
<td>Transparent Antenna Layers</td>
<td>2</td>
</tr>
<tr>
<td>2.1</td>
<td>Test geometry</td>
<td>9</td>
</tr>
<tr>
<td>2.2</td>
<td>Transmission coefficient at $f_0$ for a crossed dipole behind a PEC sheet</td>
<td>14</td>
</tr>
<tr>
<td>2.3</td>
<td>Corporate feed for a 16x16 element array</td>
<td>15</td>
</tr>
<tr>
<td>2.4</td>
<td>Transmission coefficient at different $R$</td>
<td>17</td>
</tr>
<tr>
<td>2.5</td>
<td>Crossed dipole ($R = 13.8\lambda_0$) obstructed radiation pattern, x-z plane</td>
<td>18</td>
</tr>
<tr>
<td>3.1</td>
<td>Microstrip transmission line</td>
<td>23</td>
</tr>
<tr>
<td>3.2</td>
<td>Microstrip feed network with finite PEC ground plane</td>
<td>25</td>
</tr>
<tr>
<td>3.3</td>
<td>Transmission coefficient for corporate feed with finite ground plane</td>
<td>26</td>
</tr>
<tr>
<td>3.4</td>
<td>Axial ratio with finite ground plane corporate feed</td>
<td>29</td>
</tr>
<tr>
<td>3.5</td>
<td>Parallel Strip Concept</td>
<td>31</td>
</tr>
<tr>
<td>3.6</td>
<td>Transmission coefficient for parallel-strip corporate feed</td>
<td>33</td>
</tr>
<tr>
<td>3.7</td>
<td>Axial Ratio for parallel-strip corporate feed</td>
<td>34</td>
</tr>
<tr>
<td>3.8</td>
<td>2x2 array corporate feed networks</td>
<td>35</td>
</tr>
</tbody>
</table>
3.9 Corporate feed network with leads $m \geq \frac{\lambda}{4}$ ................. 36
3.10 Transmission coefficient for corporate feed leads $m \geq \frac{\lambda}{4}$ .... 37
3.11 Axial ratio for corporate feed network with leads $m \geq \frac{\lambda}{4}$ .... 38
3.12 Transmission coefficient for an array of dipoles ................. 40
3.13 Corporate Feed with Supplemental dipoles ....................... 41
3.14 Transmission Coefficient for corporate feed with supplemental dipoles 42
3.15 Axial ratio for RHCP crossed dipole underneath corporate feed with supplemental dipoles ....................... 43
4.1 Feed Networks with different orientation ....................... 46
4.2 Transmission Coefficient for horizontally oriented and vertically oriented feed networks ....................... 47
4.3 Axial ratio for RHCP crossed dipole blocked by horizontally oriented and vertically oriented feed networks ....................... 48
4.4 Phase compensation technique ....................... 51
4.5 RHCP antenna array fed by different feed networks ....................... 52
4.6 Radiation pattern for phase compensated array ....................... 53
4.7 Tuned couplers in series ....................... 54
4.8 Series in parallel feed network ....................... 57
4.9 Series feed network joint ....................... 58
4.10 Quadrature Hybrid Coupler ....................... 58
4.11 Transmission Coefficient vs. frequency for Series Feed Network .... 60
4.12 Axial ratio with series in parallel feed network ....................... 61
4.13 Rotationally Symmetric Feed Network for a 16x16 element array . . . 63
4.14 Transmission Coefficient vs. frequency for Rotationally Symmetric Series Feed Network . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 64
4.15 Axial ratio vs. frequency for rotationally symmetric feed networks . . 65
5.1 Optimization Parameters for Corporate Feed FSS geometry . . . . . . 71
5.2 Objective function map with respect to \( l_{dip} \) . . . . . . . . . . . . . 75
5.3 Objective function map with respect to \( d_x \) and \( d_y \) and \( l_{dip} = 0.5\lambda_0 \) . . 76
5.4 Optimization objective function . . . . . . . . . . . . . . . . . . . . . 78
5.5 Radiating characteristics at \( f_0 \) for crossed dipole array over optimized corporate feed/FSS . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 79
5.6 Transmission characteristics for crossed dipole underneath optimized corporate feed/FSS . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 81
5.7 Design guidelines for a feed network for an electromagnetically transparent antenna array . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 82
5.8 Non-optimized feed network design . . . . . . . . . . . . . . . . . . . . 87
5.9 Transmission coefficient for non-optimized transparent feed network . 88
5.10 Transmission coefficient for FSS with and without feed network present . 89
5.11 Axial ratio for non-optimized feed network design against corporate feed designs . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 90
6.1 Multi-band Antenna . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 93
6.2 Multi-band Antenna . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 94
6.3 Dual Rhombic Loop . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 95
6.4 Single element RHCP gain . . . . . . . . . . . . . . . . . . . . . . . . . . . . 96
6.5 FSS layer ................................................................. 97
6.6 Simulated and Measured Transmission characteristics of L band FSS 98
6.7 Measured Transmission characteristics of L band and S band FSS . . 99
6.8 Feed network quarter panel for L band array .............................. 101
6.9 Single quadrant gain for horizontal polarization .......................... 103
6.10 S band array radiation pattern, free space ............................... 105
6.11 S band array radiation pattern, 25% blockage ......................... 106
6.12 S band array radiation pattern, 50% blockage ......................... 107
6.13 S band array radiation pattern, 75% blockage ......................... 108
6.14 S band array radiation pattern, 100% blockage ....................... 109
6.15 S band array axial ratio ................................................. 111
6.16 Normalized RHCP boresight gain at X band with L and S band array 
    blockage .................................................................. 112
6.17 X band axial ratio with L and S band array blockage ................. 113
CHAPTER 1

INTRODUCTION

1.1 Problem Description

Multiple antennas operating under a single radome offer an advantage on platforms with stringent space requirements. However, multiple antennas within a single radome can also create an obstructed line of sight for an antenna. The outer antennas create blockage for the inner antennas as seen in Figure 1.1. Partial blockage can also occur with independently pointing antennas. A phase shift for part of the incident or radiated field may occur in these situations. Blockage scenarios require an antenna that is electromagnetically transparent at frequencies outside the antenna’s radiating bandwidth.

It has been shown that a planar array ground plane can be replaced with a frequency selective surface (FSS) [1]. In addition to the FSS, an electromagnetically transparent antenna must have an array of radiators and feed network that are transparent at the underlying antenna’s operational frequencies, see Figure 1.2.

A corporate feed is a typical method used to feed an antenna array. Unfortunately, this type of feed network may cause blockage. A feed network design must
Figure 1.1: Multiple antennas under a single radome

Figure 1.2: Transparent Antenna Layers
take electromagnetic transparency into consideration or risk blockage. This dissertation explores various feed network design techniques that consider electromagnetic transparency as a design goal.

1.2 Problem Domain

The problem is to design an antenna array feed network that provides the desired signal distribution to the antenna array elements while maintaining the transparency of the antenna array. This is a new area of research and limited work has been done in this area. Research exists in other areas with similar goals. Radomes and FSS are designed to have a unity transmission coefficient within a given frequency range and zero transmission coefficient over another frequency range. Antenna array feed networks are designed to be capable of providing precise power distribution to an antenna array.

1.2.1 FSS

A FSS can be designed to have a unity or close to unity transmission coefficient with respect to both frequency and angle of incidence [2]. FSS designs and techniques can be adapted to feed networks. Feed networks will be shown that demonstrate frequency dependent reflective behavior.

Erdemli previously used a FSS in the near field of an array in order to improve the far field gain [1]. Stacked FSS layers are designed to be reflective at unique frequencies and are positioned at different distances below the planar array to create a broadband reflector for the broadband planar array [1].

Kent examined an antenna array of non-planar rectangular loops embedded in a stratified dielectric medium [3]. This phased array must behave like a radome
structure, i.e. be electromagnetically transparent. This phased array design must also consider typical phased array design issues, e.g. scan angle. Kent derives the scattered fields of the antenna array by considering the rectangular loops as a doubly infinite infinite array and using the plane wave spectrum technique.

Each non-planar rectangular loop uses an embedded source. The feed network is excluded from Kent’s analysis. This dissertation considers the transmission and reflection characteristics of a feed network for a finite antenna array, only including the antenna array radiating elements when necessary.

1.2.2 Radome Techniques

Radomes are used to enclose antennas for protection from the elements or for structural integrity [4], [5]. Metal dipoles are sometimes added to improve upon the structural integrity of the radome [6].

Radomes must be electromagnetically transparent but also must be structurally strong. Commonly used techniques to accomplish one goal may reduce performance for another goal. FSS and sandwiched layers of dielectric have been used in order to improve the transmission coefficient [2], [7]. The FSS embedded within the radome have been analyzed with the plane wave spectrum (PWS) technique [2], [8].

1.2.3 Feed networks

The feed network design and measurement results shown in Chapter 6 have been presented previously [9]–[14]. A multiband antenna system that operates with 3 independently pointing planar arrays with the same rotational center was considered in [9], [10]. The outer two planar arrays need to be transparent to prevent blockage. This transparency requirement resulted in new techniques used in designing the
feed networks for the two outer planar arrays [11], [12]. Measurements show that the integrated feed networks provide a controlled signal distribution and improved transmissivity [13], [14].

Previous work has compared the radar cross section (RCS) of a series feed network to a corporate feed network [15], [16]. Feed networks are generally not included in the analysis of transparent antenna arrays characteristics [3].

1.3 Dissertation Outline

Chapter 2 introduces the figures of merit, design tools, and environmental parameters that will be used throughout this work. A corporate feed example is designed and simulated using a method of moments (MOM) numerical technique. This corporate feed model is used to demonstrate the simulation geometry and results presented in the rest of the dissertation.

Corporate feeds are a common type of antenna array feed network. Chapter 3 investigates the transmission coefficient of different corporate feeds using different transmission line topologies. Microstrip and parallel-strip implementations are examined. Chapter 4 examines various design techniques to improve the transmission coefficient. A technique is introduced that achieves far field radiation without phase matched outputs from the feed network.

Chapter 5 examines the use of optimization for corporate feed design. A corporate feed network is combined with an array of dipoles using an optimization technique and a classical design technique. Another feed network design is shown that uses several of the previously shown transparency design techniques. Simulation results
compare the performance of the developed transparent feed network to commonly used corporate feed designs.

Chapter 6 examines the construction, integration, and measurement of a feed network for an electromagnetically transparent antenna array. This feed network is based on the design proposed in Chapter 5. This dissertation is concluded in Chapter 7, where the accomplished work is summarized and ideas for future work are discussed.
CHAPTER 2

DESIGN CRITERIA AND EVALUATION

A good design begins with a well defined problem that has quantifiable performance objectives. Criteria and tools are established prior to dealing with the designs in order to establish consistent evaluations for each feed network design technique.

2.1 Numerical simulation

The Electromagnetic Surface Patch Code 5.4 (ESP 5.4) was developed at The Ohio State University. Electromagnetic Surface Patch Code 5.4 (ESP 5.4) is used to model the scattering and radiating characteristics of the feed networks and multiband antenna arrays in Chapter 6. ESP 5.4 uses a method of moments (MOM) technique based on a piecewise sinusoidal solution for thin wires [17]. This technique is fairly accurate since the modeled feed networks are wire structures. ESP 5.4 is costly in terms of computational time since the size of the matrix equation to be solved increases as the wire structure being simulated grows larger. The simulations performed typically took 10 hours or more on an Intel Pentium 4 CPU 3.0 GHz with 1.0 GB of RAM. Some electrically large structures took over a week to simulate.
The ESP 5.4 geometries were generated using MathWorks MATLAB 7.2 (MATLAB) [18]. These geometries were visualized using ESP Workbench [19]. ESP Workbench was also used to generate many of the wire element graphics shown.

Ansoft High Frequency Structure Simulation v10.4 (HFSS) was used to examine certain FSS structures and uses a finite element method (FEM) solution. HFSS 10.4 allows us to simulate an infinite FSS with an incident plane wave. The solution time is on the order of a few minutes.

The feed network S parameters are not evaluated for every technique due to limitations of the various simulation packages used. The multi-band antenna feed network S parameters were modeled in HP ADS during the design phase. The multi-band antenna feed network signal distribution is also examined during the implementation and measurement in Chapter 6.

2.2 Figures of Merit

Figures of merit are needed to make quantitative comparisons of the design techniques. Defining criteria in terms of quantitative results allows the use of optimizers in Chapter 5 as well as objective comparisons of each technique. The figures of merit are simulated using numerical computational methods. These simulated results are compared to measured results in Chapter 6.

The design goal is to develop a feed network that will have minimal effect on the far field radiation of an underlying antenna. The feed network must also distribute a signal to an antenna array. The feed network must be transparent over specific frequency bandwidths, similar to an FSS. The specific frequency band for transparency is tied to unique project requirements. Design techniques developed to make a feed
network transparent at a specific bandwidth will not be applicable to most situations. We must consider the techniques based on the transparent bandwidth size and cutoff frequencies. We also need to be concerned with attenuation and polarization of a signal when its path is obstructed by a feed network.

2.2.1 Simulation Test Setup

A pair of dipoles are fed $90^\circ$ out of phase at the design frequency to create a right hand circularly polarized (RHCP) plane wave, see Figure 2.1. The crossed dipoles are modeled as perfectly matched to the source and produce broadband gain.

![Figure 2.1: Test geometry](image)
The figures of merit used in this dissertation compare the free space radiation of crossed dipoles to the radiation obtained with a feed network obstruction. The distance, \( R \), between the device under test (DUT) and the crossed dipoles must be large enough that a caustic is not created. The crossed dipole is aligned with the \( x \) and \( y \) axes and is parallel to the DUT.

Angles near boresight are of primary concern. This reflects conditions in which the device under test completely obscuring the crossed dipoles. Partial blockage of the source will be minimal if the attenuation and phase shift for the total blockage case is also minimal. The figure of merit simulations were performed at \( \theta = 0^\circ \), \( \phi = 0^\circ \) unless otherwise specified. The result is the \( \hat{\phi} \) component is aligned with the \( \hat{y} \) component and the \( \hat{\theta} \) component is aligned with the \( \hat{x} \) component.

The feed networks are designed to feed a planar 16x16 array with 0.5\( \lambda_0 \) spacing. This spacing is chosen to eliminate grating lobes. The feed network may be coplanar to the array or located on a parallel layer. The antenna array is excluded in many of the simulations in order to isolate the behavior of the feed network from the other layers as well as expedite the simulations.

The feed network reflective bandwidth is an area of concern. A feed network with a large reflective bandwidth at the resonant frequency will not allow operation of nearby antennas with similar radiating frequencies. For the purposes of the analysis, the antenna array for the feed network radiates over a narrow bandwidth.

The figures of merit are simulated from \( 0.5f_0 \leq f \leq 2.0f_0 \), where \( f_0 \) is the resonant frequency of the antenna array fed by the feed network. A unity transmission coefficient is most difficult to obtain near the resonant frequency of the array. A unity
transmission coefficient is easier to achieve at higher frequencies \( f \geq 2.0f_0 \) due to the smaller wavelength [12].

### 2.2.2 Transmissivity

The Fresnel transmission coefficient for a planar interface at normal incidence is a function of the transmitted and incident electric fields across a dielectric interface [20] and is defined as

\[
T = \frac{E_t}{E_i}
\]  

(2.1)

where \( E_t \) is the electrical field transmitted through the interface, and \( E_i \) is the electrical field incident upon the interface. The Fresnel transmission coefficient at oblique incidence yields

\[
\begin{bmatrix}
E_t^\parallel \\
E_t^\perp
\end{bmatrix} = [T] \begin{bmatrix}
E_i^\parallel \\
E_i^\perp
\end{bmatrix}
\]

(2.2)

\[
[T] = \begin{bmatrix}
\frac{E_t^\parallel}{E_i^\parallel} & \frac{E_t^\perp}{E_i^\perp} \\
\frac{E_t^\parallel}{E_i^\parallel} & \frac{E_t^\perp}{E_i^\perp}
\end{bmatrix}
\]

(2.3)

where \( \parallel \) and \( \perp \) refer to the parallel and perpendicular components respectively.

ESP 5.4 allows us to simulate the far field directivity of a radiating element. The far field radiation of a right hand circularly polarized (RHCP) crossed dipole is simulated. The free space directivity of the crossed dipole is compared to the directivity pattern of the crossed dipole obstructed by the device under test (DUT). The directivity is expressed as

\[
D = \frac{4\pi U(\theta, \phi)}{P_{rad}}
\]

(2.4)
where

\[ U = \frac{r^2}{2\eta} |\vec{E}(r, \theta, \phi)|^2 \]  \hspace{1cm} (2.5)

and \( P_{\text{rad}} \) is the total power radiated and \( \eta \) is the impedance of free space [21]. Comparing the obstructed and free space directivity yields

\[
D_{\text{obs}} = \frac{4\pi r^2}{2\eta P_{\text{rad}}} |\vec{E}_{\text{obs}}|^2 \\
D_{\text{free}} = \frac{4\pi r^2}{2\eta P_{\text{rad}}} |\vec{E}_{\text{free}}|^2 \\
D_{\text{obs}} / D_{\text{free}} = \frac{|\vec{E}_{\text{obs}}|^2}{|\vec{E}_{\text{free}}|^2} \\
\] \hspace{1cm} (2.6) \hspace{1cm} (2.7) \hspace{1cm} (2.8) \hspace{1cm} (2.9)

Instead of using the standard definition of the Fresnel transmission coefficient in terms of parallel and perpendicular components, transmission coefficient will be defined in terms of the far-field \( \theta \) and \( \phi \) components.

\[
|T_{ij}| = \left| \frac{\vec{E}_{\text{obs}}}{\vec{E}_{\text{free}}} \right| = \sqrt{\frac{D_{\text{obs}}}{D_{\text{free}}}} \\
\] \hspace{1cm} (2.10)

where \( i, j = \theta, \phi \).

The DUT transmission coefficient calculated in this manner is different from the traditionally used Fresnel transmission coefficient. The DUT transmission coefficient is a closer representation of what occurs when a feed network obscures an underlying antenna. The DUT transmission coefficient includes diffraction effects in addition to transmission through the interface. The diffraction effects can be significant depending on the nature of the DUT.

A perfect electrical conductor (PEC) plate was placed in front of a set of crossed dipoles (\( R = 2.4m = 13.6\lambda_0 \)). The PEC plate is the same size (\( L = 1.3m = 7.5\lambda_0 \)) as the feed networks examined in later chapters. A PEC plate represents the worst
case blockage scenario. The edge and corner diffraction effects will also be significant with a PEC plate due to the knife edge around the perimeter and the four corners. This can result in a significantly higher boresight gain.

Figure 2.2 shows the x-z plane radiation pattern of a crossed dipole behind a PEC plate. The $\hat{\theta}$ field is the hard polarization and the $\hat{\phi}$ field is the soft polarization. This is consistent with what is expected from knife edge diffraction [22]. The upper hemisphere radiation pattern ($-90^\circ \leq \theta \leq 90^\circ$) shows a null between $\pm 20^\circ$. The $\pm 20^\circ$ sub-angle corresponds to the angles subtended by the PEC plate. Any radiation within $\pm 20^\circ$ must be a result of edge and corner diffraction from the PEC plate. The main lobe at boresight is thus attributed to diffraction. This is further corroborated by the strong backlobe.

The calculated DUT transmission coefficient for the PEC plate in Figure 2.2 will be much higher than the actual Fresnel transmission coefficient for a PEC plate due to edge diffraction. It is expected that the feed network will not experience such significant diffraction due to the absence of a knife edge.

The presence of the feed network may result in a caustic, yielding a DUT transmission coefficient greater than 0 dB. The corporate feed shown in Figure 2.3 is a typical feed network for antenna arrays [16], [23]. The corporate feed divides into two or more transmission lines at regular intervals. Many corporate feeds adjust the impedance of the two split transmission lines to provide proper impedance matching to the previous transmission line [24].

Figure 2.4 shows the transmission coefficient with the crossed dipole positioned at different distances, $R$, behind a corporate feed network ($L = 7.5\lambda_0$). The transmission coefficient is taken at angle $\theta = 0^\circ, \phi = 0^\circ$. The crossed dipole boresight radiation
Figure 2.2: Transmission coefficient at $f_0$ for a crossed dipole behind a PEC sheet
Figure 2.3: Corporate feed for a 16x16 element array
with the corporate feed positioned at \( R = 5.7\lambda_0 \) and \( R = 6.8\lambda_0 \) is several dB above free space radiation for \( \hat{\theta} \) polarization. The crossed dipole at \( R = 13.8\lambda_0 \) yields a transmission coefficient that does not significantly exceed 0 dB for the entire frequency band.

Figure 2.5 shows the radiation pattern at 0.98\( f_0 \) and 1.5\( f_0 \) for a crossed dipole \((R = 13.8\lambda_0)\) blocked by a corporate feed. The \( \hat{\theta} \) polarization is parallel with the \( \hat{x} \) oriented dipoles on the far edge of the corporate feed (upper left and lower right in Figure 2.3). The \( \hat{\theta} \) polarization in Figure 2.5a shows significant ripples in the radiation pattern. A caustic is indicated by the presence of the ripples.

Figure 2.4 shows minimal attenuation from the corporate feed at 1.5\( f_0 \) and \( R = 13.8\lambda_0 \). The radiation pattern at 1.5\( f_0 \) is shown in Figure 2.5b and exhibits minimal diffraction. This supports the assertion that the 0 dB transmission coefficient at higher frequencies in Figure 2.4a \((R = 13.8\lambda_0)\) is not due to edge diffraction but due to the feed network being "transparent".

This analysis uses a evaluated DUT transmission coefficient that differs from the Fresnel transmission coefficient. The Fresnel transmission coefficient considers the electric field phase, where this evaluation does not consider the phase. The phase is considered in the next figure of merit, axial ratio. The transmission coefficient is also used to refer the \( S_{21} \) parameter of a transmission line. The signal distribution of the feed network will be explicitly referred to as the S parameters or feed network signal distribution.
Figure 2.4: Transmission coefficient at different $R$. 

(a) $\hat{\theta}$ polarization

(b) $\hat{\phi}$ polarization
Figure 2.5: Crossed dipole ($R = 13.8\lambda_0$) obstructed radiation pattern, x-z plane
2.2.3 Polarization

The feed network can affect the polarization of the transmitted signal. Circularly polarized signals can become linearly polarized or RHCP signals may become left hand circularly polarized if the feed network attenuation is polarization dependent. Note that the corporate feed transmission coefficient has different frequency nulls for orthogonal polarizations, see Figure 2.4. Some FSS are known to cause cross-polarization [2].

The polarization is characterized by the axial ratio (AR) [25]

\[
AR = \frac{E_{\text{max}}}{E_{\text{min}}} = \pm \frac{\text{major axis}}{\text{minor axis}}
\]

where \(1 \leq |AR| \leq \infty\). The axial ratio is positive for left hand circular polarization (LHCP) and negative for RHCP. The polarization can also be represented on the Poincaré sphere using the angles

\[
\gamma = \tan^{-1}\left[\frac{E_\phi}{E_\theta}\right]
\]

\[
\delta = \phi_\phi - \phi_\theta
\]

where \(\phi_\phi\) and \(\phi_\theta\) are the phases of \(E_\phi\) and \(E_\theta\) respectively. The angle \(2\gamma\) is the arc angle between a reference point on equator and the polarization point on the sphere. The angle \(\delta\) is the angle between the equator and the arc angle. The angles \((\delta, \gamma)\) are related to the axial ratio through the use of the Poincaré sphere latitude \(\epsilon\) [25]

\[
\sin(2\epsilon) = \sin(2\gamma) \sin(\delta)
\]

\[
AR = \cot(\epsilon)
\]

The axial ratio (AR) of the test RHCP crossed dipole (AR = −1) is evaluated with the feed network obstructing its radiation. The purpose is to determine whether
a RHCP signal will become linearly polarized or left hand circular polarized when being obstructed by certain feed networks. The angle $\epsilon$ is used in Chapter 5 as an optimization parameter. This dissertation will refer to AR in the linear scale in order to differentiate between LHCP and RHCP.

### 2.3 Transparency techniques

Several techniques to improve the transmission coefficient will be examined. Each technique needs to consider performance as a feed network capable of providing equal power and phase to each element of the array and a scattering structure with minimal attenuation and polarizing effects. The figures of merit relating to the scattering properties for the feed networks are simulated through the use of ESP 5.4. The performance as an antenna feeding network was simulated using a combination of HP-ADS, Ansoft HFSS, and handbook formulas.

The calculated transmission coefficient must be close to unity (0 dB) for the feed network to be useful as a transparent structure. Antenna arrays behave like an FSS at their resonant frequency [3]. An FSS reflector is often used in combination with an array of radiating elements in order to increase the gain [1]. Both the array of radiating elements and FSS structures have strong reflective properties at their operating frequency. Designing the feed network to be transparent at a specific bandwidth close to its antenna array radiating frequency ($f_0$) is necessary to allow other antennas operating at close frequencies. The frequencies at which a feed network becomes transparent are examined. The feed network may also be reflective at the resonant frequency of the antenna array. The reflective bandwidth will determine the permissible operational frequency bands of nearby antennas.
The proposed concepts for feed network design will result in multiple techniques capable of being incorporated into the design of an antenna array. This is done in Chapter 5. This feed network is later built and in Chapter 6. The next chapter examines corporate feeds and microstrip circuits. The effect of a finite ground plane is examined as well.
CHAPTER 3

CORPORATE FEED NETWORKS

Arrays are commonly fed using microstrip transmission lines over a substrate and PEC ground. Microstrip transmission lines are easy to design and fabricate, and they work well with many types of arrays [15], [16], [23], [24], [26]–[29]. The microstrip feed network uses a PEC ground plane and often implements a corporate feed layout. This chapter shows several corporate feed examples and a number of approaches are explored.

The ground plane and substrate size can be reduced to a finite width to obtain a better transmission coefficient without significantly degrading the transmission line characteristics [30], [31].

A parallel-strip method can be used to print the transmission lines [32]. This method offers similar benefits as traditional printed microstrip circuits but eliminates the PEC ground plane. A corporate feed network design is simulated using parallel-strips.

3.1 PEC Ground

The ground plane and substrate can be truncated without affecting the transmission line performance. The reduced ground plane allows for some transparency. A
A microstrip transmission line, shown in Figure 3.1, can use a finite ground plane without drastically affecting the characteristic impedance of the line [31]. The substrate dielectric ($\varepsilon_r$), the substrate edge separation over the transmission line width ($S_{sub}/W$), and the transmission line width ratio ($W/h$) contribute to the minimum separation between the ground and transmission line edges ($S_{GND}/h$). In general, the transmission line characteristics are unchanged if $S_{GND}/h \geq 2$.

There is a contribution to the fringe fields of the microstrip transmission line at small values of $S_{sub}$. The substrate cannot be removed at these distances without changing the propagation characteristics of the transmission line. For larger values of $S_{sub}/h$, the magnitude of charge density decreases and reduces the effect on the fields near the strip [30]. The propagation characteristics of microstrip transmission lines are not significantly changed from those of an infinite substrate transmission line for $S_{sub}/W \geq 1$. 

Figure 3.1: Microstrip transmission line
The corporate feed transmission line splits into $N$ transmission lines at regular intervals. This classical arrangement uses T-junctions that split the signal using impedance matching techniques. Corporate feeds can feed $N^M$ signals, where $M$ is an integer. The corporate feed used here feeds $16 \times 16 = 256 = 2^8$ elements.

The width of a microstrip transmission line is typically a few hundredths of a wavelength. The width of the corporate feed PEC ground in Figure 3.2 is $2S_{GND} + W = 5.7\lambda_0$. Using a common dielectric (Rogers Duroid 5580, $\epsilon_r = 2.2$, $h = 0.00045\lambda_0$) yields a $W = 0.0137\lambda_0$ for a 50\Omega transmission line. This results in a $S_{GND}/h = 4.8$ and $S_{GND}/W = 1.57$, and satisfies the previously mentioned ground plane separation criteria for maintaining transmission line characteristics.

The microstrip corporate feed, shown in Figure 3.2, was simulated using ESP 5.4 ($R=13.8\lambda_0$). Only the finite ground plane is included in the simulation. The microstrip cross-section in Figure 3.1 cannot be simulated in ESP 5.4, as ESP 5.4 cannot handle parallel plates less than 0.01\lambda apart or PEC plates on dielectric volumes [17].

The transmission coefficient for the corporate feed ground plane is shown in Figure 3.3. Different resonant frequencies are observed in the $\hat{\theta}$ polarization (Figure 3.3a and the $\hat{\phi}$ polarization (Figure 3.3b).

The PEC corporate feed network has a large reflective bandwidth. The value of $T_{\theta\theta}$ is worse than -1 dB below $f_0$ and does not increase above $-1$ dB until $1.3f_0$. The value of $T_{\phi\phi}$ component is below -1 dB for $0.5f_0 \leq f \leq 0.82f_0$. In both cases, the reflective bandwidth is over 30\% for both polarizations. The value of $T_{\theta\phi}$ and $T_{\phi\theta}$ for a corporate feed network are examined in Section 3.2.
Figure 3.2: Microstrip feed network with finite PEC ground plane

The ground plane separation ($S_{GND}$) has an effect on the reflective bandwidth and the reflective resonant frequency. The $\hat{x}$ oriented transmission lines between adjacent array elements act as resonant dipoles near $f_0$. The $\hat{y}$ oriented segments cause reflective behavior in $T_{\phi\phi}$ at a lower frequency. The cross-section width of the ground plane (0.057$\lambda_0$) corresponds to an equivalent circular cross-section radius of 0.014$\lambda_0$ [33]. This yields a length over diameter of $l/d = 18$, as opposed to a thin wire dipole which may have an $l/d \geq 50$. The reflectivity bandwidth is considered further when we examine a corporate feed built using thin wires in section 3.2.

The reflective resonant frequency is related to the length and width of the transmission line connecting two adjacent array elements. The array elements are separated by $d = \frac{\lambda_0}{2}$. The feed network transmission lines between adjacent array elements act
Figure 3.3: Transmission coefficient for corporate feed with finite ground plane
as an array of dipoles with a reflective frequency at \( f_0 \). A microstrip ground plane has a width

\[
W_{GND} \approx W + 2S_{GND}
\]  

(3.1)

The width of the PEC ground plane decreases the effective length of the dipole and increases the resonant frequency of the reflectivity region. The dipole effective length becomes

\[
l_{eff} = \frac{\lambda_0}{2} - \frac{W_{GND}}{2}
\]  

(3.2)

We can calculate the resonant wavelength of the corporate feed

\[
\lambda_1 = \lambda_0 - 2W_{GND}
\]  

(3.3)

This results in a resonant frequency of the corporate feed

\[
f_1 = \frac{c}{\lambda_1}
\]  

(3.4)

\[
f_1 = \frac{c}{\lambda_0 - 2W_{GND}}
\]  

(3.5)

\[
f_1 = \frac{1}{1 - \frac{2W_{GND}}{\lambda_0}}f_0
\]  

(3.6)

The \( T_{\theta\theta} \) resonant reflectivity frequency for the corporate feed is higher than the array frequency. This is supported by ESP 5.4 results. Corporate feeds with different values of \( S_{GND} \) are simulated and shown in Figure 3.3. The \( T_{\phi\phi} \) lower reflective frequency null does not shift significantly due to the smaller relative change at the larger wavelengths.
We can improve the overall antenna transparency by overlapping the reflective bandwidth of the antenna array and the FSS. The microstrip corporate feed should be designed with $T_{\theta\theta}$ and $T_{\phi\phi}$ resonant frequency at the array frequency $f_0$.

The antenna designer generally has the option of choosing the relative positions of the antennas. For example, a designer designing multiple independent pointing antennas co-sited within a radome may choose to locate the lower frequency antennas in the outermost rotation. The higher frequency antennas are positioned behind the lower frequency antennas. The finite ground plane corporate feed may be appropriate in this scenario due to transparency at higher frequencies. The field transmitted through the finite ground plane has an axial ratio above -1.5 and individual $T_{\theta\theta}$ and $T_{\phi\phi}$ better than -1.0 dB for $f > 1.3f_0$. This is indicated in Figure 3.4.

The transparent antenna array might be required to operate with pre-existing antennas. The pre-existing antennas may operate at a lower frequency than the transparent antenna array. This is an issue as antennas continue to be added to already crowded sites. A new feed network design must be considered for these situations.

These results indicate that it is possible to use the corporate feed network under certain conditions. The required transparent frequency band should be higher the antenna resonant frequency. This scenario occurs when the corporate feed is blocking higher frequency antennas. The corporate feed example shown has a large reflective bandwidth, which is linked to the ground plane width. This limits the transparent frequency band resonant bandwidth and the allowable operating frequency of underlying antennas.
Figure 3.4: Axial ratio with finite ground plane corporate feed

The frequency sweep from $0.5f_0$ to $2.0f_0$ with 101 data points took over 31 hours to simulate on a AMD Athlon 2.2 GHz Processor with 1GB of RAM.

### 3.2 Parallel-Strip Circuits

The PEC ground plane can be completely eliminated from the microstrip circuit using parallel-strips. This technique prints microstrip transmission lines and other microstrip circuits on both sides of the substrate as shown in Figure 3.5a. This method requires a substrate that is twice the thickness of an equivalent PEC backed substrate. The mirror image physical circuit replaces the image theory equivalent that is normally produced by reflections from the PEC ground plane [25]. The characteristic impedance of a transmission line printed using this method is twice that of a similar microstrip line printed over a ground plane [32]. This can be expressed as
\[ Z_{0\text{parallel–strip}} = 2Z_{0\text{microstrip}}(h = d/2) \]  
\[ \epsilon_{\text{rparallel–strip}} = \epsilon_{r\text{microstrip}}(h = d/2). \]

where \( h \) is the height of PEC backed microstrip substrate, see Figure 3.1 and \( d \) is the height of the parallel-strip substrate as illustrated in Figure 3.5a.

This method provides a balanced transmission line, eliminates a PEC ground plane, and also allows us to control the transmission line impedance in order to obtain a close impedance match between the feed network and the array elements. The elimination of the ground plane also becomes significant in tuning the corporate feed reflective frequency to the resonant frequency of the antenna array.

A parallel-strip corporate feed was modeled using ESP 5.4. We are still limited to the previously mentioned ESP 5.4 restrictions on parallel PEC plates [17]. Electromagnetically transparent feed networks will favor thin substrates with low dielectric constants in order to minimize the phase shift caused by the feed network. A single wire approximation is reasonable if the distance between the parallel strips is electrically close,

\[ d \leq 0.01\lambda_{\text{sub}} \]  

where

\[ \lambda_{\text{sub}} = \frac{\lambda_0}{\sqrt{\epsilon_r}} \]  

The parallel strip separation can be expressed in terms of free space wavelength

\[ \frac{d}{\lambda_0} \leq 0.01 \frac{1}{\sqrt{\epsilon_r}} \]
An electromagnetically transparent feed network using a low dielectric substrate, \( \epsilon_r = 2.2 \), requires \( d \leq 0.0067\lambda_0 \). We can also express this requirement in terms of the dielectric constant

\[
\epsilon_r \leq 0.0001 \left( \frac{\lambda_0}{d} \right)^2
\]  

(3.12)

A thin dielectric substrate (\( d = 31 \) mils = 0.0045\( \lambda_0 \)) requires a substrate with a dielectric constant, \( \epsilon_r \leq 4.9 \). The parallel-strip feed network design implemented in Chapter 6 uses a substrate thickness of \( d = 31 \) mils. A Rogers Duroid 5880 (\( \epsilon_r = 2.2 \)) substrate satisfies the single wire approximation criteria.

The parallel-strip corporate feed transmission coefficient given in Figure 3.6 demonstrates both frequency and polarization dependent behavior. The parallel-strip corporate feed was simulated using both round and flat wires for comparison. The transmission coefficients are nearly identical for both round and flat wire feed networks.
The $\hat{\theta}$ polarization is better than -1.0 dB at at $1.05f_0$ and the $\hat{\phi}$ polarization is better than -1.0 dB at 0.82$f_0$. The frequency band where the parallel-strip corporate feed has an axial ratio better than -1.5 at $f \geq 1.15f_0$, see Figure 3.7. The cross-polarized components of the transmission coefficient ($T_{\theta\phi}$ and $T_{\phi\theta}$) are both more than -100 dB across the entire frequency band. The result is that an underlying antenna array fed by a parallel-strip corporate feed network can operate at a lower frequency compared than an antenna array operating underneath a finite ground plane corporate feed.

The $\hat{\theta}$ component transmission coefficient has a null near the array frequency ($1.04f_0$). The corporate feed behaves as a FSS reflector for a $\hat{\theta}$ polarized signal. This array and corporate feed is transmissive to an orthogonally polarized antenna transmitting through it. It may be possible to construct a polarization independent FSS ground plane can be constructed from the corporate feed by adding $\hat{\phi}$ polarized resonant elements to the feed.

The corporate feed network is connected to a $2^N \times 2^N$ antenna array. This array can be divided into 4 element groups (2x2). Each 2x2 group is fed by an I shaped corporate feed, see Figure 3.8a. The last transmission line segment between the feed network and array elements are $\hat{x}$ oriented $m = \lambda_0/4$ segments. These transmission line segments can be manipulated to adjust the transmission coefficient nulls.

Both the amplitude and resonant frequency of the nulls in the transmission coefficient can be reduced by increasing the transmission line length, $m$. The length of the leads can be increased from $0.5d \leq m \leq 0.71d$, or $0.25\lambda_0 \leq m \leq 0.35\lambda_0$ (see Figure 3.8b). An example of such corporate feed network is shown in Figure 3.9.
Figure 3.6: Transmission coefficient for parallel-strip corporate feed
The transmission coefficient of two corporate feeds with leads $m = 0.31\lambda_0$ and $m = 0.35\lambda_0$ are shown in Figure 3.10. The $m = 0.25\lambda_0$ case represents the typical I shaped corporate feed network discussed in previous sections. Both the resonant frequency and amplitude of $T_{\phi\phi}$ increase as the length of the feed network leads increase. This is consistent with the orientation of the leads changing with respect to $\hat{\phi}$. A transmission line lead length of $0.35\lambda_0$ resonates at $0.71f_0$, which agrees with the ESP 5.4 simulations in Figure 3.10.

The axial ratio for corporate feeds with increased values of $m$ are shown in Figure 3.11. The values of $T_{\theta\theta}$ and $T_{\phi\phi}$ for the $m = 0.25\lambda_0$ corporate feed are close to 0 dB above $f_0$. The same parallel-strip corporate feed network causes the crossed dipole axial ratio to degrade from -1 to -1.5 at $1.15f_0$ and falls off rapidly. The cutoff
Figure 3.8: 2x2 array corporate feed networks
frequency decreases to $0.95 f_0$ for the $m = 0.31 \lambda_0$ case. The $m = 0.31 \lambda_0$ case does not drop below -1.5 until $0.75 f_0$ and does not exhibit the same linearly polarizing and circularly polarized (CP) shifting behavior as the former two corporate feeds.

An example parallel-strip corporate feed has been shown with transparent characteristics at or above the feed network resonant frequency. The same corporate feed has polarization dependent reflective behavior at the resonant frequency. A corporate feed can be used as a polarization independent FSS by adding additional resonant elements. This would allow the corporate feed FSS to act as a reflector for the antenna array. This is explored in the next section.

3.3 Corporate Feed as a Frequency Selective Surface

A parallel-strip corporate feed was demonstrated with reflective properties at the array operational frequency. Figure 3.6a shows a null in $T_{\theta\theta}$ less than -10 dB near $f_0$ for the flat wire case. This behavior can be exploited in order to create a frequency
Figure 3.10: Transmission coefficient for corporate feed leads $m \geq \frac{\lambda_0}{4}$
Figure 3.11: Axial ratio for corporate feed network with leads $m \geq \frac{\lambda_0}{4}$
selective surface that will act as a reflector for the array and allow underlying antennas to operate freely.

A transparent feed network needs to be capable of providing both a uniform distribution to the array as well as behaving as an FSS capable of handling arbitrary polarization. The amplitude of $T_{\theta\theta}$ is consistent with the desired FSS behavior. The ratio between $T_{\theta\theta}$ and $T_{\phi\phi}$ must be improved. Figure 3.7 shows that the ratio between $T_{\theta\theta}$ and $T_{\phi\phi}$. Creating a null in $T_{\phi\phi}$ at $f_0$ will improve the $T_{\theta\theta} / T_{\phi\phi}$ ratio.

An array of dipoles can handle linear polarization with the E-field in the plane of the elements. Arbitrary polarized waves can be handled by orienting an array of dipoles such that their polarization is $90^\circ$ with respect to the parallel-strip corporate feed. Placing two orthogonally oriented arrays of non-touching dipoles in close proximity does not distort the reflection coefficient near the resonant frequency [34].

The dipole array will not affect the S parameter of the corporate feed network unless certain criteria are met. The separation between the dipole layer and the corporate feed layer must be close enough that the dipoles are within the fringe fields of the corporate feed, such that the dipole couples to the corporate feed. The dipoles will couple to the feed network if this occurs. The spacing between the dipole array and the corporate feed must be large enough in order to avoid coupling, but small enough to reduce cross-polarization effects. The corporate feed parallel-strip transmission line fringe fields will be largely confined to the substrate between the two mirror image feed networks.

Two-line couplers operate on a similar principle. A two-line coupler uses a geometry that uses a $\frac{\lambda}{4}$ transmission line segment parallel to the microstrip transmission line to couple power from the feed network [24], [35]. A known limitation of this type
of coupler is its inability to couple tightly to the microstrip transmission line [11]. Two line couplers implemented in high dielectric constant substrates yield larger separation between the parallel transmission line segments [24], [36], [37].

The onset of frequency grating lobes can be delayed by reducing the distance between adjacent dipoles [2]. The bandwidth of the reflection coefficient also increases as we reduce the element spacing. A balance must be made between reducing sidelobes and keeping the bandwidth comparable to the corporate feed network. An array of dipoles with $\lambda_0$ spacing in the $\hat{y}$ polarized direction and $\frac{\lambda_0}{2}$ spacing in the orthogonal direction was modeled in ESP 5.4. The transmission coefficient for an array of $\frac{\lambda_0}{2}$ dipoles oriented in the $\hat{y}$ direction is shown in Figure 3.12.
As expected, the dipole array is transparent to cross polarized signals. The array is reflective to co-polarized signals at $f_0$. The dipole array is positioned below the corporate feed in Figure 3.13a. The dipoles are located a few percent of a wavelength below the corporate feed as shown in Figure 3.13b. The vertical separation avoids creating shorts in the corporate feed. Cross polarization may occur unless the separation is small [2], [34]. The smaller the separation between the two layers, the lower the cross-polarization.

The transmission coefficient for the corporate feed with the supplemental dipole layer is shown in Figure 3.14. Both polarizations demonstrate a null at $f_0$. The result is a reflective surface for RHCP signals at $f_0$. The transmission coefficient behavior at frequencies other than $f_0$ is similar to the corporate feed in Figure 3.6. The axial ratio with the modified corporate feed blockage is shown in Figure 3.15 and is similar to the axial ratio with corporate feed blockage. These figures support our earlier
Figure 3.14: Transmission Coefficient for corporate feed with supplemental dipoles
Figure 3.15: Axial ratio for RHCP crossed dipole underneath corporate feed with supplemental dipoles

assertion that the dipole array does not change the transmission coefficient behavior of the corporate feed.

The corporate feed has transparent behavior for a RHCP signal. The addition of a orthogonally polarized FSS layer improves the layer’s ability to handle right hand circular polarization. This corporate feed FSS design was developed using classical FSS design techniques. Chapter 5 demonstrates a similar design using optimization methods.
3.4 Summary

Examples of unmodified corporate feeds have been shown to be transparent above the resonant frequency. The cutoff frequencies are different for corporate feeds built with a PEC ground plane or built using parallel-strips.

The unmodified corporate feed transmission coefficient also demonstrate a strong sensitivity to polarization. The $\hat{\theta}$ polarization is reflective near the resonant frequency while the $\hat{\phi}$ polarization is reflective well below the resonant frequency.

A method for building a feed network with polarization independent FSS at the array resonant frequency was demonstrated. This feed network used the addition of $\frac{\lambda_0}{2}$ dipoles in order improve the polarization sensitivity. This type of feed network would be suitable for use as both a feed network and array ground plane. This method is developed further in Chapter 5

The corporate feed is acceptable for situations where higher frequency transparency is desired. The designer may be allowed to choose the relative antenna location. Lower frequency antennas positioned in front of the higher frequency antenna arrays avoids the problem of the poor bandpass behavior in the regions $f \leq f_0$. This is done in some designs in order to avoid antenna array blockage [9], [10]. This type of transmission coefficient is reminiscent of many bandstop FSS where the transmission coefficient has better passband characteristics above the reflective frequencies [2].

The corporate feed structure demonstrates reflectivity at resonant frequencies at or below $f_0$. This is due to the resonant physical features of the feed network are related to the array spacing and design. The feed network topology needs to be decoupled from the antenna array spacing and its associated frequencies. This is examined in the next chapter.
Chapter 3 showed typical corporate feeds used to feed antenna arrays. Corporate feed examples were demonstrated with reflective behavior at the array frequency and polarization dependent nulls. This behavior is dependent on the feed network physical geometry. The feed network geometry is linked to the antenna array geometry. This chapter presents a method which decouples the feed network geometry from the antenna array geometry.

Rotational symmetry can preserve the polarization of the underlying antenna. This method is most effective when the blocking feed network is restricted to a full blockage scenario. Feeding the array in subsections using orthogonally oriented feed networks creates feed networks with rotational symmetry.

New feeding schemes have been introduced with polarization independent reflective behavior. This can be exploited to create a feed network that can act as an FSS. An alternative is to reduce the FSS behavior by orienting the feed network plane orthogonally relative to the blocked antenna.

The choice of dielectric substrate affects the transmission coefficient. Substrates with high dielectric coefficients can result in a large phase shift for the incident plane waves and lossy dielectrics can cause attenuation. This creates a desire for thinner
substrates with lower dielectric constants. Techniques which allow the use of thinner and lower dielectric constant substrates will improve the feed network transparency.

4.1 Vertical orientation

The corporate feed networks shown in Chapter 3 are printed on a single planar layer. Feed networks using microstrip couplers and multiple parallel transmission lines will be introduced later. Feed networks are commonly printed on dielectric substrates using microstrip or parallel-strip techniques. A series in parallel feed network (discussed in Section 4.2.2) is shown in Figure 4.1a. These single layer feed networks can be oriented perpendicular to the array in a non-planar orientation. Figure 4.1b shows a similar microstrip feed network has been rotated 90° (x-axis rotation) relative to the blocked antenna.
The horizontal feed network demonstrates significant nulls for both polarizations and are indistinguishable in Figure 4.2. The horizontally printed feed network has a significant reflective bandwidth relative the resonant frequency. The planar feed network shown in Figure 4.1a possesses a strong similarity to an FSS of square loops. The square couplers of the feed network have a $\lambda_0$ long perimeter similar to square loop FSS elements. The axial ratio with the horizontal feed network blockage is shown in Figure 4.3. Rotating the feed network transmission lines vertically results in an improved polarization purity from the crossed dipole.

The vertically oriented feed network in Figure 4.1b uses a microstrip transmission line to bridge the vertical gap between the feed network and antenna array [11], [13],
Figure 4.3: Axial ratio for RHCP crossed dipole blocked by horizontally oriented and vertically oriented feed networks

[14]. This simplifies the transmission line design and construction. The vertically oriented feed network transmission coefficient shows improvement over the planar feed network. No significant reflective bandwidth is present for the vertically oriented feed network.

4.2 Feed Network Layout

The corporate feed network geometry is dependent on the antenna array element spacing. The array spacing is chosen to reduce grating lobes and maximize aperture size [38]. This spacing is a function of the radiating frequency wavelength. One
result is that the corporate feed structure demonstrates reflective behavior at regions of interest.

A feed network design can result in unequal signal paths to each array element if the feed network is asymmetrical. The feed network may provide a uniform or controlled signal distribution, but yield unequal phase to each element. The array will not produce the desired far field radiation in this case. Compensation for the unequal feed network phase output must be provided. Phase shifters are often used in electronically scanned or phased arrays. Phase shifters often use high dielectric substrates which unsuitable for a transparent feed network [26], [28]. Phase shifters may also be unsuitable for low cost arrays [9], [10].

Phase compensation techniques and frequency scanning can result in beam squint. As stated previously, this feed network is limited to feeding narrow band antenna arrays. Phase compensation methods which rely on the signal phase are suitable for narrow band applications.

4.2.1 Unequal phase compensation

Corporate feeds are commonly used for antenna arrays [23], [27]. They provide equal phase to each array element. Chapter 3 showed examples of corporate feeds with transparency at frequencies above the array radiating frequency. The typical application allows the designer of multiple antenna systems to position the higher frequency antennas behind the lower frequency antennas. This may not always be the case. The feed network designer may be required to design a feed network capable of transparency at lower frequencies. New feed network methods must be explored.
The feed network phase output, $\phi$, can be calculated given the design parameters or measured with a vector network analyzer. Rotating individual circularly polarized array elements around the axis of propagation shifts the element reference coordinates with respect to the array coordinate axis. This rotation causes a relative phase shift in the element far field radiation with respect to the other array elements, see Figure 4.4.

The phase compensation technique rotates each circularly polarized array element based on the calculated feed network phase output. Rotating a right hand circularly polarized array element (clockwise) by a $2\pi - \phi$ angle results in a far-field radiation pattern for the array. A right hand circularly polarized (RHCP) array fed by a phase matched feed network is shown in Figure 4.5a. A RHCP array fed with unequal phase uses rotated elements for phase compensation and is shown in Figure 4.5b.

The far field antenna parameters of an array fed by a phase matched feed network or a phase compensated series feed networks was simulated in ESP 5.4 (see Table 4.2.1 and see Figure 4.6). The boresight CP gain for both feed networks differ by less than 0.2 dB with similar sidelobe levels. The radiation pattern when using either feed network is very similar, see Figure 4.6. The 1st sidelobe levels for the phase matched feed $\hat{\theta}$ polarization are slightly higher than those of the sidelobe levels for the phase compensated array. The sidelobe levels relative to the main beam are similar. The sidelobe levels and pattern are nearly identical for the $\hat{\phi}$ polarization.

The antenna array axial ratio improves from -2.18 to -1 with the phase compensation technique. This is due to the physical rotation of the elliptically polarized array elements. The phase output provided by the feed network has a relatively uniform distribution from $0 \leq \phi \leq 360^\circ$. The uniform distribution results in array elements oriented in pseudo-random directions. The pseudo-random orientations of the array
elements causes an element by element shift in the polarization tilt angle \((\tau)\). The cumulative shifts in \(\tau\) and subsequent far field summation results in a nearly perfect far field CP signal as shown by Table 4.2.1. The axial ratio is preserved across the main beam half power beamwidth.

This phase compensation method requires CP (right hand or left hand) antenna elements. Linearly polarized array elements cannot be rotated relative to the propagation axis and still provide linear polarized radiation. Linearly polarized array elements can only be rotated \(\pm 180^\circ\) while maintaining polarization. This requires
another phase shift from $0^\circ \leq \theta \leq 180^\circ$. Phase shifters are undesirable due to the effect on the transmission coefficient. Creating a phase delay with additional microstrip transmission line would also cause problems since the additional lengths may cause transmission coefficient nulls at frequencies $f > f_0$. 

Table 4.1: Antenna Parameters with and without Phase Compensation

<table>
<thead>
<tr>
<th></th>
<th>phase matched</th>
<th>phase compensated</th>
</tr>
</thead>
<tbody>
<tr>
<td>$G_\theta$</td>
<td>25.7 dBi</td>
<td>23.6 dBi</td>
</tr>
<tr>
<td>$G_\phi$</td>
<td>23.4 dBi</td>
<td>23.6 dBi</td>
</tr>
<tr>
<td>$G_{RHCP}$</td>
<td>26.8 dBi</td>
<td>26.6 dBi</td>
</tr>
<tr>
<td>1st sidelobe</td>
<td>-17.3 dB</td>
<td>-17.0 dB</td>
</tr>
<tr>
<td>AR</td>
<td>-2.18</td>
<td>-1.00</td>
</tr>
</tbody>
</table>
Figure 4.6: Radiation pattern for phase compensated array
Phase compensation allows array feed networks to be designed without individually adjusting the transmission line to each element for equal phase. Feed networks will be demonstrated which take advantage of this technique. A feed network that uses microstrip couplers in series is also presented.

### 4.2.2 Series Feed Network

A series feed network is developed using microstrip couplers. Series couplers allow the number of transmission lines to be reduced. This configuration connects each array element to two adjacent array elements via sequentially connected couplers. This is illustrated in Figure 4.7. Each coupler has a unique coupling ratio in order to achieve equal magnitude at the feed outputs.

![Coupler Diagram](image)

**Figure 4.7: Tuned couplers in series**

Couplers are 4 port devices with an input (in), a isolated port (iso), a through port (out), and a coupled port (coup). The isolated port is connected to the ground (via a 50 Ω resistor). Multiple microstrip couplers are connected in series, see Figure 4.7. The coupler output is fed into the subsequent coupler’s input. The coupled port is the transmission line output. This output is connected to either an array element...
or may be connected to another set of couplers in series. Each coupler has a tuned coupling ratio. Lower coupling ratios are necessary for couplers closer to the input in order for to achieve a uniform power distribution. An amplitude of $\frac{1}{N}$ is obtained for each output by progressively increasing the coupling ratio. Appendix A describes the coupling ratio calculations in detail.

One concern with the series feed is achieving reasonable and manufacturable coupling ratios for the microstrip couplers. A single series coupler transmission line connected to a large array yields a large range of coupling ratios with minute changes in value. Coupling ratios are determined by microstrip trace width or line separation [24], [39]–[42]. Achieving the necessary tolerances for controlled signal distribution can become difficult. Extreme coupling ratios also require thick dielectrics with high dielectric constants [12]. Thick dielectrics and high dielectric constants can have a negative effect on the transparency.

An array of NxN elements fed with a single series feed has a minimum change in coupling ratio, $\Delta C$

$$\Delta C = \frac{1}{N^2 - 1} - \frac{1}{N^2} \quad (4.1)$$

$$\Delta C = \frac{N^2}{N^2(N^2 - 1)} - \frac{N^2 - 1}{N^2(N^2 - 1)} \quad (4.2)$$

$$\Delta C = \frac{1}{N^2(N^2 - 1)} \quad (4.3)$$

The required coupling ratios can be reduced by using parallel sets of series transmission lines. One series transmission line is used to feed multiple sets of series transmission lines in parallel. The main line is a series transmission line whose outputs are connected to N branch lines, see Figure 4.8a. The branch line is connected to N radiating elements. Each series transmission line only has N outputs as opposed
to the $N^2$ outputs in the case of the single series transmission line. This means that the coupling ratio for each individual main or branch series feed is $\frac{1}{N} \geq C \geq \frac{1}{2}$.

When an $N \times N$ array is fed with two sets of series transmission lines, $\Delta C$ becomes

$$\Delta C = \frac{1}{N - 1} - \frac{1}{N}$$  \hspace{1cm} (4.4)

$$\Delta C = \frac{1}{N(N - 1)}$$  \hspace{1cm} (4.5)

The use of a main line and multiple branch lines results in a $N(N + 1)$ change in the coupling ratio requirement. The smallest coupling ratios for a single series feed for a $16 \times 16$ array are $\frac{1}{256}$ and $\frac{1}{255}$. The individual coupling ratios differ by as little as $1.5 \times 10^{-5}$. Feeding the same $16 \times 16$ array with a combination of a main and branch lines results in a minimum coupling ratio difference of is $400 \times 10^{-5}$. The smallest coupling ratio for the main and branch combination is $\frac{1}{16}$. The value of $\Delta C$ can be further reduced by dividing the array into subdivisions [14]. This is shown in Section 4.3.

Non-rectangular arrays, such as octagons, can be fed by varying the number of feed points on the branch line. An octagon array is implemented in Chapter 6. The unequal number of elements in each row requires new coupling ratio calculations. These calculations are covered in Appendix A.2.

Multiple types of microstrip couplers are capable of achieving the required coupling ratios [24], [41]. A feed network is built in Chapter 6 using quadrature hybrid couplers [13], see Figure 4.10. The quadrature hybrid coupling ratio is adjusted by controlling the impedances of the vertical and horizontal legs, $Z_{0B}$ and $Z_{0A}$ respectively [41]. This can be written as
Figure 4.8: Series in parallel feed network
Figure 4.9: Series feed network joint

Figure 4.10: Quadrature Hybrid Coupler
\[ Z_{0A} = Z_0 \times \sqrt{\frac{P_A}{P_B} \frac{1 + \frac{P_A}{P_B}}{P_A P_B}} \]  \hspace{1cm} (4.6)

\[ Z_{0B} = Z_0 \times \sqrt{\frac{P_A}{P_B}} \]  \hspace{1cm} (4.7)

where \( P_A \) is the through port power output and \( P_B \) is the coupled port power output as shown in Figure 4.10. The microstrip coupler dimensions are found using (4.6), (4.7), and (3.7) to calculate the required impedance and then applying standard microstrip transmission line equations [24]. The joint between the main line and branch line using quadrature hybrid couplers is shown in Figure 4.9.

Microstrip coupling ratios are frequency specific. This makes the feed network inherently narrowband. This is acceptable for narrowband antenna arrays. Broadband couplers are needed for arrays with larger bandwidths. The bandwidth for a quadrature hybrid coupler can be increased by using two couplers in a double box configuration [41].

A parallel series feed network for a 16x16 array is shown in Figure 4.8b. The \( \hat{\phi} \) component is parallel (upper right in Figure 4.11) with the main transmission line and demonstrates less than 1 dB attenuation. The \( T_{\theta\theta} \) curve exhibits more attenuation than the \( T_{\phi\phi} \) component which is consistent with the alignment of the branch feed networks.

The axial ratio with a series in parallel feed network blockage is shown in Figure 4.12. The phase compensation technique allows the number of transmission lines to be reduced. The RHCP radiation of the crossed dipole is preserved over a large portion of the frequency sweep.
Chapter 3 shows examples of corporate feed networks commonly used in practice. These corporate feed networks affect the axial ratio. This affects the electromagnetic wave polarization incident on the feed network. This section examines rotational symmetry in the feed network. The result is that the axial ratio of the obstructed crossed dipole is preserved at boresight ($\theta = 0^\circ, \phi = 0^\circ$). Both previously used and new feed network designs are shown with rotational symmetry.

This technique only yields perfect symmetry when the feed network is centered and directly over the blocked array. Full blockage is more common when two apertures have fixed pointing angles relative to each other. An example is when a transparent
antenna array is added in the presence of a preexisting antenna system. The rotational symmetry technique does not provide perfect symmetry for independently pointing application since partial blockage is a more common scenario.

This method has the advantage of reducing the coupling ratio requirements in the series feed network. A 16x16 array can be divided into 4 subarrays of 8x8. The orthogonally oriented 8x8 sub-array feed network only has to feed 64 elements instead of 256 elements. This reduces the coupling requirement and allows a thinner substrate with a lower dielectric constant.

An MxN array is divided into L subarrays. Each subarray is fed by a feed network rotated 90° around the center axis. An example is shown in Figure 4.13a using a series
feed network. The 16x16 array has been divided into 4 subarrays. The feed networks for each subarray have been rotated 90° with respect to the vertical axis. The result is a rotationally symmetric feed networks are shown in Figure 4.13. Both parallel series and corporate feeds are shown as an example.

The transmission coefficient of rotationally symmetric feed network for a 16x16 array is shown in Figure 4.14. The orthogonal components of the transmission coefficient for both types of feed network are indistinguishable. Figure 4.15 shows the axial ratio for both the parallel series feed network and the corporate feed. The RHCP radiation is preserved after blockage by the rotationally symmetric feed networks. Compare this to the parallel series feed network axial ratio in Figure 4.12 and the corporate feed axial ratio in Figure 3.6. A CP incident electromagnetic wave does not become linearly polarized due to transmission through the feed network at the lower frequencies.

4.4 Final Comments

This chapter presented several feed network design techniques. The goal was to improve the transmissivity of the feed network, while maintaining the performance of the feed network. The phase compensation technique is particularly important because it allows for the feed network physical geometry to be decoupled from that of the array. Resonant feed network segments created by the array element spacing can be eliminated.

Chapter 5 shows an example of a feed network design that implements several of the techniques. This new feed network design is compared to the commonly used corporate feed network. This feed network is constructed and measured in Chapter 6.
Figure 4.13: Rotationally Symmetric Feed Network for a 16x16 element array
Figure 4.14: Transmission Coefficient vs. frequency for Rotationally Symmetric Series Feed Network
Figure 4.15: Axial ratio vs. frequency for rotationally symmetric feed networks
CHAPTER 5

FEED NETWORK DESIGNS

Multiple techniques have been presented for the design of electromagnetically transparent antenna feed networks. A feed network for a RHCP array has been designed using previously presented techniques. This feed network design is fabricated, integrated, and measured in Chapter 6.

This chapter examines the use of an optimization technique in designing a corporate feed network for an 8x8 RHCP antenna array. The corporate feed is designed to act as a reflector for the antenna array and be transparent for frequencies $f \geq f_0$.

5.1 Corporate FSS Optimization

Previous chapters demonstrated electromagnetically transparent feed networks developed using classical design techniques. The corporate feed FSS is one example. This section uses a local optimization technique in order to design a feed network to meet several performance goals.
Optimization techniques are used to solve the nonlinear constrained problem (NCP) of the general form [43]

\[
\begin{align*}
\min & \quad F(x) \\
\text{subject to constraints} & \quad c_i(x) = 0, \quad i = 1, 2, \ldots, m' \\
& \quad c_i(x) \geq 0, \quad i = m' + 1, \ldots, m
\end{align*}
\]

(5.1)

where \( F(x) \) is the objective or cost function, constraint functions \( \{c_i(x)\} \) are real scalar values. The optimization techniques find values of \( x \) which achieve local minima of \( F(x) \). The local optimization values are dependent on initial conditions and as such these initial conditions have to be chosen carefully. The optimization technique has no knowledge of the problem not contained within the definition in Equation 5.1. The design parameters, initial conditions, and cost function must be carefully chosen to represent the true problem. A poor choice of optimization parameters and cost functions can lead to a poor local minimum.

Chapter 3 presented a design for a corporate feed supplemented with dipoles in order to create a FSS ground plane for the radiating antenna array. Optimization is employed to further refine this design. Several objectives are met. The corporate feed FSS is designed to be transparent for frequencies \( f_{\text{trans}} = 1.3f_0 \) and to be reflective at \( f_0 \). The antenna array axial ratio should be the same with or without the corporate feed FSS. The original purpose of developing the corporate feed as a FSS is to improve the antenna array RHCP gain at \( f_0 \).

5.1.1 Algorithm

A quasi-Newton method is used for the optimization algorithm [18]. The optimization code was implemented in MATLAB using the optimization toolbox. ESP 5.4 was
used to evaluate the figures of merit. Upper and lower boundary constraints are used to limit the search space and to maintain sensible optimization values. Constraints were not placed on the optimization variables.

Newton’s method uses the first derivative of the objective function to calculate the next approximation of the local minimum. Another similar technique uses the Hessian matrix (H) of the objective function, to calculate updates to the next approximation of the local minimum [44], [45]. Direct calculation of the Hessian matrix is computationally intensive and requires multiple evaluations of the objective function.

The quasi-Newton method calculates a new Hessian matrix by using the values calculated during the previous iteration [46]. Several methods exist for updating the Hessian matrix, of which the Brodyen-Fletcher-Goldfarb-Shanno (BFGS) technique is used in this optimization [18].

5.1.2 Variables

The dipole array has multiple design parameters, x, see Table 5.1 and Figure 5.1. Number of elements, dipole length, spacing between the corporate feed and the dipole arrays were some variables initially chosen for optimization.

The dipole array consisted of \( \hat{y} \) oriented dipoles located underneath the corporate feed network. The \( \hat{x} \) dipoles were excluded in order to reduce both the search space and computation time. Chapter 3 shows several examples of corporate feeds. These corporate feeds demonstrate a transmission coefficient null at \( f_0 \) in the \( \hat{\theta} \) polarization (\( \phi = 0^\circ \)). The \( \hat{\phi} \) transmission coefficient component at \( f_0 \) was close to 0 dB (\( \phi = 0^\circ \)). Only \( \hat{y} \) dipoles will be needed in order to achieve a polarization independent corporate feed FSS.
<table>
<thead>
<tr>
<th>Variable</th>
<th>Description</th>
<th>lower bound</th>
<th>upper bound</th>
</tr>
</thead>
<tbody>
<tr>
<td>$l_{dip}$</td>
<td>dipole length</td>
<td>0.4$\lambda_0$</td>
<td>0.6$\lambda_0$</td>
</tr>
<tr>
<td>$d_x$</td>
<td>dipole spacing in the $\hat{x}$ direction</td>
<td>0.2$\lambda_0$</td>
<td>$\lambda_0$</td>
</tr>
<tr>
<td>$d_y$</td>
<td>dipole spacing in the $\hat{y}$ direction</td>
<td>0.2$\lambda_0$</td>
<td>1.2$\lambda_0$</td>
</tr>
<tr>
<td>$M$</td>
<td>Number of dipoles in the $\hat{x}$ direction</td>
<td>4</td>
<td>18</td>
</tr>
<tr>
<td>$N$</td>
<td>Number of dipoles in the $\hat{y}$ direction</td>
<td>4</td>
<td>17</td>
</tr>
</tbody>
</table>

Table 5.1: Dipole Array Parameters

The distance between the corporate feed and dipole array, $z_{off}$, should be made small while avoiding coupling to the corporate feed. This reduces the cross-polarization caused by the separate layers at $f_{trans}$ [47]. This corporate feed and dipole array separation was not included for optimization due to the desire to keep the value as small as possible.

The number of elements in the dipole array is a discrete value. The optimization techniques used are designed to finding local minima using derivatives of the objective function. These techniques require continuous variables over a closed interval. The number of elements in the array is a crucial requirement. We can capture the element number in the form of the continuous element spacings $d_x$ and $d_y$, see Figure 5.1.

The number of elements in the $\hat{x}$ direction is

\[
(M - 1)d_x = L_x
\]  

\[
M = \frac{L_x}{d_x} + 1
\]

69
where \( L_x \) is the length of the corporate feed in the \( \hat{x} \) direction. Similarly, the number of elements in the \( \hat{y} \) direction is

\[
(N - 1)d_y + l_{\text{dip}} = L_y \hspace{1cm} (5.4)
\]

\[
N = \frac{L_y - l_{\text{dip}}}{d_y} + 1 \hspace{1cm} (5.5)
\]

with the additional restriction that \( d_y > l_{\text{dip}} \), where \( l_{\text{dip}} \) is the dipole length. Both the dipole length and dipole array distance from the corporate feed were optimized.

The design variables can be forced to lie within acceptable values using upper and lower bounds. Bounds can also be used to avoid local minima that do not result in the global minimum. The upper and lower bounds are given in Table 5.1. Design variables (x) used in the optimization were \( l_{\text{dip}}, d_x, \) and \( d_y \).

5.1.3 Cost function

The choice of the cost function, or objective function, for optimization should be carefully considered. Choosing the wrong objective function to minimize can result in a design that is minimized but is impractical or performs poorly.

A corporate feed structure which serves as both a feed network and a FSS ground plane requires reflective behavior at \( f_0 \) and transparent behavior at another frequency. For the purposes of this optimization, the transparent frequency is chosen to be \( 1.3f_0 \). Optimization techniques assume the goal is to find the minimum of the cost function, see equation 5.1. Design goals must be formulated such the desired value, \( g_i \), is a minimum of the figure of merit, \( y_i(x) \). The figure of merit, \( y_i(x) \) is a function of the design parameters (x).

The initial cost function used was based on the sum of the transmission coefficient components at the bandstop and bandpass frequencies \( f_0 \) and \( 1.3f_0 \). The desired
Figure 5.1: Optimization Parameters for Corporate Feed FSS geometry
transmission coefficient values, \( y_0(x) = T_{\theta\theta}(f_0) \) and \( y_1(x) = T_{\phi\phi}(f_0) \), are minimum at \( f_0 \). We wish the corporate feed and FSS layer to be transparent at 1.3\( f_0 \), so we chose to minimize the values of \( y_2(x) = -T_{\theta\theta}(1.3f_0) \) and \( y_3(x) = -T_{\phi\phi}(1.3f_0) \). The initial cost function tried was

\[
F(x) = y_0(x) + y_1(x) + y_2(x) + y_3(x) \tag{5.6}
\]

\[
F(x) = T_{\theta\theta}(f_0) + T_{\phi\phi}(f_0) - T_{\theta\theta}(1.3f_0) - T_{\phi\phi}(1.3f_0) \tag{5.7}
\]

the sum of these values. Optimization of this objective function met with limited success. This was due in part to the null behavior at \( f_0 \). The null magnitude can vary significantly with small changes in the design parameters. A small shift in the null location can result in a large change in the transmission coefficient value at \( f_0 \). The transmission coefficient magnitude in the null can vary significantly. A one percent increase in the null frequency could result in a 5 dB increase in transmission coefficient.

The objective function in equation (5.7) can also result in poor bandpass performance at 1.3\( f_0 \). The importance of the transmission coefficient is dependent on the magnitude. For example, suppose the \( i^{th} \) iteration achieves a -15 dB transmission coefficient at \( f_0 \) and -3 dB transmission coefficient at 1.3\( f_0 \). The next iteration may result in improvement at either metric. Achieving a -18 dB transmission coefficient at \( f_0 \) is relatively inconsequential since this will have little effect on the performance of the corporate feed FSS as a reflector. A 3 dB improvement at 1.3\( f_0 \) is quite significant.

The first cost function chosen was flawed. A cost function must properly relate the design criteria to the desired performance goals. Another cost function was developed which better represents the desired performance.
A null in the transmission coefficient at $f_0$ is only a measure of reflectivity. A null in the transmission coefficient is not the final goal. Optimizing the reflection coefficient is another possibility. We wish to obtain a unity reflection coefficient as it pertains to building a reflector to improve the antenna array gain. Optimizing gain of a static RHCP antenna array located over the corporate feed FSS gives us a direct measure of how the corporate feed FSS will behave as a reflector. The gain of the array is expected to vary smoothly with respect to geometry changes.

The gain of an 8x8 array of crossed dipoles is $y_0(x) = G_{RHCP} = 20.0\text{dBiC}$. A 3 dB increase in gain is expected from a perfect reflector. The RHCP array gain goal is $g_0 = 23.0\text{dBiC}$. The goal axial ratio for the RHCP gain of the crossed dipole array is $AR = -1$.

Axial ratio is a poor choice for a optimization parameter due to the discontinuous domain. The axial ratio is defined from $1 \leq |AR| \leq \infty$. The axial ratio does not exist over the interval (-1,1). As the polarization state of an electric field transitions from right hand circularly polarized to left hand circularly polarized, the axial ratio changes from -1 to $-\infty$ and from $\infty$ to +1. The discontinuous interval and asymptotic behavior near $\pm \infty$ will cause problems for the derivative based quasi-Newtonian optimization method. The axial ratio can be mapped to the Poincaré sphere angle $\epsilon$ using equation (2.15). The value of $\epsilon$ is defined over the interval $[-45^\circ, +45^\circ]$ and is chosen as $y_1(x)$, with $g_1 = -45^\circ$.

The cost function must also consider the corporate feed transparency when blocking a radiating source (RHCP crossed dipole). Similar to the prior analysis, the transmission coefficient components and crossed dipole axial ratio are used to evaluate the FSS behavior. The third metric is the minimum of $T_{\theta\theta}(1.3f_0)$ and $T_{\phi\phi}(1.3f_0)$. 

73
Preserving the axial ratio of the RHCP crossed dipole is also important and is captured through \( y_3(x) = \epsilon_{x\text{dip}} \).

We have four performance metrics to evaluate for the objective function: gain of an array over a corporate feed FSS \( (y_0(x) = G_{\text{RHCP}}) \), polarization of an array over a corporate feed FSS \( (y_1(x) = \epsilon_{\text{array}}) \), polarization of a RHCP crossed dipole underneath the corporate feed \( (y_2(x) = \epsilon_{x\text{dip}}) \), and transmission coefficient \( (y_3(x) = -\min[T_{\theta\theta},T_{\phi\phi}]) \).

Several goals have been established for the corporate feed FSS. Not all goals are achievable and some measure of deviation from the goals must be used. One function that accomplishes this is called the regret function and is defined as [46]

\[
F(x) = \left[ \sum_{i=1}^{N} (y_i(x) - g_i)^p \right]^{1/p}
\]

where \( y_i(x) \) is the evaluated figure of merit, \( g_i \) is the goal value, and \( x \) are the design parameters of the optimized structure. A logical choice for \( p \) is \( p=2 \). This results in a formula similar to the standard deviation. Choosing an even value for \( p \) can also create a minimum value of 0 for the performance metrics defined less than zero and have a desired value at the maximum of the metric (for example, \( y_3 \)). The optimum value of the regret function is \( F(x) = 0 \). The cost function using the regret function with the previously stated metrics and goals yields

\[
F(x) = (\text{Gain}_{\text{RHCP}} - 23)^2 + (\epsilon_{\text{array}}(1.3 f_0) + \frac{\pi}{4})^2 + \min(T_{\theta\theta}(1.3 f_0), T_{\phi\phi}(1.3 f_0))^2 + (\epsilon_{x\text{dip}}(1.3 f_0) + \frac{\pi}{4})^2
\]

The cost function is mapped with fixed values of \( d_x \) and \( d_y \) in Figure 5.2. The objective function varies smoothly and a minimum exists near \( l_{\text{dip}} = 0.5 \lambda_0 \). This is
consistent with the expectation that the dipole array will resonate at $f_0$ when it is approximately half a wavelength long.

![Figure 5.2: Objective function map with respect to $l_{dip}$](image)

The cost function is plotted in Figure 5.3 with $l_{dip} = 0.5 \lambda_0$. The presence of multiple local minimums makes the local optimization technique sensitive to the initial estimate of the minimum.

The transmission coefficient and other performance metrics were evaluated with ESP 5.4. Earlier attempts at optimization evaluated the figures of merit at several frequencies. This was computationally intensive and resulted in lengthy computation times. The scope of the problem is narrowband. Each metric can be evaluated at a
Figure 5.3: Objective function map with respect to $d_x$ and $d_y$ and $l_{dip} = 0.5\lambda_0$
Table 5.2: Optimization Results

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Optimized</th>
<th>no FSS</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\text{Gain}_{\text{RHCP}}$</td>
<td>22.5 dBiC</td>
<td>21.3 dBiC</td>
</tr>
<tr>
<td>$\text{AR of radiating array}$</td>
<td>-1.1</td>
<td>-1.2</td>
</tr>
<tr>
<td>$\min(T_{\theta\theta}, T_{\phi\phi})$</td>
<td>0.5 dB</td>
<td>-0.8 dB</td>
</tr>
<tr>
<td>$\text{AR of blocked cross dipole}$</td>
<td>-1.2</td>
<td>-1.4</td>
</tr>
<tr>
<td>$F(x)$</td>
<td>0.828</td>
<td>1.89</td>
</tr>
</tbody>
</table>

single frequency point without mischaracterizing the optimization objective. Reducing the frequency sweep allows the computation time to be reduced. This becomes significant when mapping the cost function over a range of values for multiple design variables.

5.1.4 Results

The optimized dipole array uses a dipole length of $0.52\lambda_0$. The dipole spacing was $d_x = 0.52\lambda_0$ and $d_y = 0.75\lambda_0$. This yields an 8x5 array of dipoles. The corporate feed FSS designed in Chapter 3 yielded a dipole length of $0.48\lambda_0$ and an array spacing of $d_x = d_y = 0.5\lambda_0$. The optimization took less than 2 hours.

The objective function value at each iteration is shown in Figure 5.4. The quasi-Newton converges quickly and the optimization ended after a set number of function evaluations. A summary of the optimization results are given in Table 5.2.

The axial ratio and gain are plotted for an array of RHCP crossed dipoles is shown in Figure 5.5. The RHCP crossed dipoles are oriented $45^\circ$ with respect to the $\hat{x}$ and $\hat{y}$ axes. The optimized corporate feed show a small improvement in both gain and axial ratio.
Figure 5.4: Optimization objective function
Figure 5.5: Radiating characteristics at $f_0$ for crossed dipole array over optimized corporate feed/FSS
The transmission characteristics at $1.3f_0$ are given in Figure 5.6. The axial ratio of the crossed dipole through the optimized feed improves by a dB over that blocked by the corporate feed alone.

Antenna arrays require a feed network. Antenna array analyses typically do not include the feed network [3]. The integrated FSS and feed network design introduces a small change in the feed network design which yields improvement in the antenna array radiating and transparency performance. Inclusion of the FSS with the feed network layer allows elimination of a separate FSS ground plane. The CP array gain is improved over the gain for the typical array configuration.

This design is further refined by the use of optimization techniques. Implementing the optimization techniques required understanding of the underlying technical problem. In this case, optimization is best used to refine a design derived through classical design techniques.

5.2 Non-Optimized Feed Network Design

A transparent feed network design has been developed using some of the previous presented feed network design techniques. This non-optimized design was developed with classical FSS and feed network design techniques. This feed network will be used for antenna arrays (100+ elements) in a multi-band antenna system. Results of the integration stages and final measurements performed in the are given in Chapter 6.

Design guidelines for a electromagnetically transparent feed network are given in Figure 5.7. The guidelines begin with the antenna array specifications and guide the designer towards a final feed network configuration.
Figure 5.6: Transmission characteristics for crossed dipole underneath optimized corporate feed/FSS
Figure 5.7: Design guidelines for a feed network for an electromagnetically transparent antenna array
The antenna array size and geometry helps determine the choice between corporate feed network and series in parallel feed network. A corporate feed network usually feeds an rectangular array. The octagonal antenna array in Chapter 6 is difficult to feed with a corporate feed, but is easily fed with a series in parallel feed network. The corporate feed also places restrictions on the number of elements in the antenna array. The example corporate feed shown in this dissertation uses a 2-way power divider to feed a 16x16 array. The corporate feed usually feeds rectangular antenna arrays in a $2^M \times 2^N$ configuration, otherwise a series in parallel feed network should be used.

The transparent frequency, $f_{\text{trans}}$, also informs the feed network decision. The corporate feed network was demonstrated to have poor reflectivity below $1.3f_0$. A series feed network should be used if the transparent frequency, $f_{\text{trans}}$ is close to $f_0$. However a corporate feed network may be required due to other conditions, such as a linearly polarized antenna array. The reflective bandwidth of the corporate feed network can be tuned by adjusting the length of the corporate feed segments as discussed in Section 3.2.

The design guidelines illustrated in Figure 5.7 should result in an electromagnetically transparent feed network configuration. The particular idiosyncrasies of a particular design problem may call for deviation from these guidelines. This calls for the designer to keep in mind the specifics of their specific problem rather than blindly following directions.

The feed network is designed to be employed with an antenna array with a separate FSS layer, see Figure 1.2. An array and FSS ground plane combination results in an antenna that is transparent outside of the operating bandwidth.
Separating the feed network from the array layer prevents the feed network structure from interfering with the radiation of the array it is trying to feed. The separate FSS layer is located $\frac{\lambda_0}{4}$ below the array. The FSS is designed to provide narrowband reflectivity at $f_0$. The FSS is transparent with little or no attenuation outside the reflective bandwidth. The frequency dependent behavior allows the FSS to behave as a ground plane reflector for the antenna array while remaining transparent to the underlying antennas.

5.2.1 Feed Network Transparency Techniques

The FSS design used with this feed network and in Chapter 6 is an array of crossed dipoles. The open space between the FSS elements allow for vertical transmission lines from the feed network to pass through the FSS layer and connect to the array elements. These vertical transmission lines do not affect the FSS behavior as a reflector for the antenna array. The antenna array radiation pattern with the FSS is identical regardless of the presence of the feed network.

The feed network is located $\frac{\lambda_0}{8}$ below the FSS layer. This feed network uses a parallel series feed network. The parallel series feed network is implemented using a single main line and multiple branch lines. The phase compensation technique is required due to the feed network not being phase matched. The associated array uses right hand circularly polarized (RHCP) antenna elements. The transmission lines linking the feed network have been designed in order to tolerate rotation of the array elements.

The rotated angle for each array element is determined by prior calculation of the output phase for each feed point. This is easily done given the characteristics of the
substrate and the feed network geometry. A comparison of an antenna array with rotated elements for a series feed network with a corporate feed network is shown in Figure 4.5b.

The phase compensation techniques provides an advantage beyond the far field phase matching of the array signal. The rotation of each individual array element results in an array of RHCP antennas with pseudo-random orientations. The pseudo-random orientations average the tilt angle of the CP signal, which results in an axial ratio close to ±1. This allows the array designer to trade axial ratio for gain when selecting an array element. The antenna array will still yield an axial ratio close to −1 (for RHCP).

This feed network uses microstrip circuits printed on a substrate. Quadrature hybrid couplers are printed using the parallel-strip technique discussed in Section 3.2. This eliminates the PEC ground plane. The feed network substrate and printed circuits are oriented perpendicular with respect to the array face, see Section 4.1.

The feed network is divided into 4 sub panels. Each panel is oriented orthogonally around the center axis as discussed in Section 4.3. This creates rotational symmetry for the transmission coefficient. This reduces the required coupling ratios for the parallel strip couplers. The reduced coupling ratio allows the use of a lower dielectric substrate. An advantage of a 4 panel feed network is the capability for monopulse tracking.

The substrate choice depends on the microstrip coupler design and manufacturing tolerances. Two-line microstrip couplers have difficulty attaining tight coupling ratios (reference) and need a high dielectric constant to achieve realistic separation [11], [12]. Quadrature hybrid couplers may have difficulty with extremely low coupling ratios.
Coupling ratios that vary from one another by thousandths may not be achievable with manufacturing tolerances. A high dielectric substrate is then required in order to manufacture the required couplers. Early designs required a high dielectric constant, $\epsilon_r = 10.2$ and $d=250$ mils, in order to achieve uniform distribution. A high dielectric substrate negatively affects the transmission coefficient.

The coupling ratios, and therefore the microstrip coupler, depend on the antenna array size. Larger arrays require couplers with a lower coupling ratio in order to provide a controlled power distribution. The initial couplers need to tap a smaller percent of the energy for a transmission line with more elements. For a uniform distribution, the last series coupler is always a 3 dB coupler with the equal outputs connected to the last two array elements.

The coupling ratio calculations presented in Appendix A account for the desired power distribution. A common series feed network issue is calculating the coupling ratios to properly account for the substrate loss. The non-optimized feed network design in Chapter 6 deals with in two ways. A thin and low loss substrate is chosen to reduce the overall losses. HP ADS Momentum software was used to calculate the S parameters for a feed network line on a lossy substrate. The non-optimized feed network coupling ratios were tuned based on these values.

The non-optimized feed network uses a Rogers Duroid 5880 substrate ($\epsilon_r = 2.2$, $h = 31$ mils). This feed network is more transparent than previous designs ($\epsilon_r = 10.2$ and $d=250$ mils) due to the lower dielectric constant and thinner substrate [13]. A thin vertical strip of Rogers Duroid substrate can withstand 180° rotation. This flexibility allows the use of phase compensation and separate the feed network and antenna array layers.
5.2.2 Non-optimized Feed Network Simulations

The transmission coefficient for the non-optimized transparent feed network design is given in Figure 5.9. This result is compared against both the PEC ground corporate feed network and parallel-strip corporate feed network developed in Chapter 3. The nulls are smaller compared to that of the corporate feed nulls.

This complex feed network for a 16x16 array took over 42 hours to simulate using the ESP 5.4 code. The most significant null for the non-optimized feed is at $f_{null} = 0.7f_0$ in both $T_{\theta\theta}$ and $T_{\phi\phi}$ components. This null is due to resonance in the vertical transmission lines that connect the feed network to the antenna array. The vertical transmission line stubs are $0.375\lambda_0$. This vertical stub behaves as a quarter wavelength monopole near $0.67f_0$, very close to where we see the transmission coefficient null. This hypothesis is validated by simulations of the feed network without the vertical transmission lines which lack the null at $0.7f_0$. 

Figure 5.8: Non-optimized feed network design
Figure 5.9: Transmission coefficient for non-optimized transparent feed network
Figure 5.10: Transmission coefficient for FSS with and without feed network present

The transmission coefficient for an FSS is shown in Figure 5.10. The FSS demonstrates a null near the resonant frequency. A feed network is added underneath the FSS with vertical transmission lines through the FSS spaces. The presence of the feed network does not make a significant difference in the behavior of the FSS.

Figure 5.11 compares the axial ratio of a RHCP crossed dipole obscured with non-optimized transparent feed network and corporate feed examples. Both corporate feed network examples yield polarizing behavior above and below the feed network design frequency. The corporate feed networks can attenuate one polarization to the extent that RHCP polarization becomes linearly polarized. The corporate feed wire example does not yield an axial ratio better than -1.5 until $1.3f_0$ which limits the operational
frequencies of the underlying the antennas. A phase shift can also occur which causes the signal to become left hand elliptically polarized at $f_0$ and $0.5f_0$.

The non-optimized feed example preserves the polarization state of the RHCP signal from the crossed dipole across the the entire frequency band. The axial ratio for the non-optimized feed network does not get worse than -1.07 across $0.5f_0 \leq f \leq 2.0f_0$. The non-optimized feed network design is implemented on two RHCP antenna arrays with similar operating frequencies in Chapter 6.
5.3 Comments

Examples of antenna array feed networks have been developed using optimization and classical design techniques. A optimization procedure presented a method of designing a corporate feed with supplementary dipoles. This optimized corporate feed improves the gain at \( f_0 \) while maintaining transparency at some frequency, \( f_{\text{trans}} \). The procedure was demonstrated with \( f_{\text{trans}} = 1.3 f_0 \). The regret function was used to combine multiple figures of merit into a single cost function for optimization.

A feed network design has been developed with classical design techniques (non-optimized) using methods for improving transparency presented in prior chapters. This design has been constructed and tested at The Ohio State University. The process and results are documented in the next chapter.
CHAPTER 6

TRI-BAND ANTENNA EXAMPLE

A non-optimized antenna array feed network has been designed using previously shown transparency techniques. An L band antenna array and S band antenna array have been built using this design. These concentric antennas point independently in a multi-band antenna system, shown in Figure 6.1. The two outer antennas arrays are designed to be transparent at the operating frequencies of the lower antennas.

The multi-band antenna system (shown in Figure 6.2) was tested at the compact range at the ElectroScience Laboratory at the Ohio State University [48].

6.1 Multi-band Antenna Design

The tri-band antenna presented in this chapter uses 3 independently pointing antennas, each operating at a unique frequency band [9], [10], [12], [13], see Figure 6.1. Each antenna array radiates at a lower frequency than the antenna behind it. The outer two antennas must be designed to be transparent above their operating frequency. The L band array is the outermost antenna and must be transparent at S and X band. The S band array is located between the L and X band arrays and must be transparent to the X band array below it.
Figure 6.1: Multi-band antenna test configuration (top view)
Figure 6.2: Multi-band antenna for L, S, and X band data


6.1.1 Array and FSS Design

The antenna is a planar octagonal array as shown in Figure 4.5b. The octagonal shape was chosen to allow the antennas to be closely positioned and avoid collision during independent steering. The unequal number of elements in each row is considered when designed the feed network, see Appendix A.

The array elements are cut dual rhombic loop elements [49], shown in Figure 6.3. This array element is appropriate for this application due to the dimensions and boresight gain. The cut loop element uses a balanced feed and a balun is not needed when using the parallel strips method. The dual rhombic loop is also right hand circularly polarized (RHCP) which allows the use of the far field phase compensation technique.

The cut loop elements were printed using a silver ink on mylar. The boresight gain for a L and S band single element was measured using a coaxial cable and balun for the feed. This measurement is compared against an ESP 5.4 simulation in Figure 6.4.
The ESP 5.4 simulation uses a dielectric sleeve to approximate the effects of the mylar substrate. The dielectric sleeve dimensions were determined by correlating the L band gain measurement with the ESP 5.4 results. The dielectric constant of the sleeve was set at $\varepsilon_r = 3.2$ (Mylar). The thickness of the dielectric sleeve was adjusted until the frequency of the null in simulated gain matched the measured results. The sleeve thickness and dielectric values were used in the S band single element simulations and the FSS simulations and were consistent with measured results.

![Graph showing single element RHCP gain](image)

Figure 6.4: Single element RHCP gain

The FSS layer is an array of crossed wire dipoles, shown in Figure 6.5. The dipoles are printed on both sides of a mylar dielectric substrate. The FSS layer is located between the antenna array and the feed network, see Figure 1.2.
The L band FSS was measured using a pair of linearly polarized AEL 1 to 12.4 GHz horns. The FSS measurements are compared to ESP 5.4 results, shown in Figure 6.6. The ESP 5.4 results used the same dielectric sleeve dimensions as derived from the L band single element gain measurements. The agreement between ESP 5.4 simulation and measurement in Figure 6.4 and Figure 6.6 indicate that the substitution of a dielectric sleeve for a dielectric sheet is reasonable.

The transmission characteristics of the L band and S band FSS were measured both separately and together, see Figure 6.7. A metal plate of the same width and length of the FSS was also measured for comparison. A well designed FSS will still yield some measure of transmissivity due to the finite size. The metal plate represents the best case reflectivity the FSS can achieve. Both the L and S band FSS achieve nulls of the same magnitude as the metal plate. Note that the L band FSS does not produce significant blockage at the higher S band frequency.
Figure 6.6: Simulated and Measured Transmission characteristics of L band FSS
Figure 6.7: Measured Transmission characteristics of L band and S band FSS
6.1.2 Feed Network Design

The feed networks for the L band and S band arrays are divided into four sections each. The four sections can be fed with a four-way power divider or be used to implement monopulse tracking. The four L band sections are connected to 30 elements and each section for the S band array is connected to 43 elements resulting in 120 elements and 172 elements for the L and S band arrays, respectively. This allows for implementation of the "pin wheel" feed network.

The feed network used parallel series fed transmission lines fed by a single series feed network, see Figure 6.8a. The series feed networks are implemented using quadrature hybrid couplers [50]. The coupling calculation was initially calculated by hand and fine tuned using HP-ADS Momentum, see Appendix A.

The feed network quarter panels were manufactured in sections to allow easier testing and simply manufacturing. The smaller sections allowed us to measure the output for individual couplers using a vector network analyzer. These measured results were then used to synthesize the quarter panel output at each feed point of the array. The synthesized quarter panel output for the L band feed network at the array radiation frequency is shown in Table 6.1.2. The desired output at each feed point is close to the desired level of -14.8 dB. A taper is present from the input to the outer feed points. This taper will result in a reduced sidelobe and increased beamwidth for the radiation pattern. The S band feed network output has a desired coupling level of -16.3 and is shown in Table 6.1.2.

The use of series couplers fed in parallel results in an unequal phase output. This requires the use of the phase compensation introduced in Chapter 4. The feed network phase output was calculated based on the feed network geometry, coupler output, and
Figure 6.8: Feed network quarter panel for L band array
<table>
<thead>
<tr>
<th>Main Branch Index</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>-14.8</td>
<td>-14.4</td>
<td>-16.0</td>
<td>-16.2</td>
<td>-16.5</td>
<td>-17.6</td>
</tr>
<tr>
<td>3</td>
<td>15.0</td>
<td>-14.6</td>
<td>-16.2</td>
<td>-16.4</td>
<td>-16.7</td>
<td>-17.8</td>
</tr>
<tr>
<td>4</td>
<td>-16.0</td>
<td>-16.1</td>
<td>-16.3</td>
<td>-16.5</td>
<td>-17.6</td>
<td>n/a</td>
</tr>
<tr>
<td>5</td>
<td>-16.4</td>
<td>-16.6</td>
<td>-16.8</td>
<td>-17.9</td>
<td>n/a</td>
<td>n/a</td>
</tr>
<tr>
<td>6</td>
<td>-16.5</td>
<td>-16.2</td>
<td>-17.0</td>
<td>n/a</td>
<td>n/a</td>
<td>n/a</td>
</tr>
</tbody>
</table>

Table 6.1: Coupling (dB) for L band quarter panel

<table>
<thead>
<tr>
<th>Main Branch Index</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-15.3</td>
<td>-17.7</td>
<td>-14.5</td>
<td>-16.4</td>
<td>-16.8</td>
<td>-16.7</td>
<td>-17.0</td>
</tr>
<tr>
<td>2</td>
<td>-16.4</td>
<td>-18.7</td>
<td>-15.6</td>
<td>-17.4</td>
<td>-17.8</td>
<td>-17.8</td>
<td>-18.0</td>
</tr>
<tr>
<td>3</td>
<td>-14.8</td>
<td>-17.1</td>
<td>-13.0</td>
<td>-15.8</td>
<td>-16.3</td>
<td>-16.2</td>
<td>-16.5</td>
</tr>
<tr>
<td>4</td>
<td>-15.1</td>
<td>-17.4</td>
<td>-14.3</td>
<td>-16.1</td>
<td>-16.6</td>
<td>-16.5</td>
<td>-16.8</td>
</tr>
<tr>
<td>5</td>
<td>-18.6</td>
<td>-14.9</td>
<td>-16.8</td>
<td>-17.2</td>
<td>-17.2</td>
<td>-17.5</td>
<td>n/a</td>
</tr>
<tr>
<td>6</td>
<td>-18.9</td>
<td>-17.7</td>
<td>-18.2</td>
<td>-18.2</td>
<td>-18.4</td>
<td>n/a</td>
<td>n/a</td>
</tr>
<tr>
<td>7</td>
<td>-19.0</td>
<td>-19.4</td>
<td>-19.5</td>
<td>-19.7</td>
<td>n/a</td>
<td>n/a</td>
<td>n/a</td>
</tr>
</tbody>
</table>

Table 6.2: Coupling (dB) for S band quarter panel
the effective dielectric constant [24]. The phase compensation technique was tested on a single quarter panel of the L band array. The quarter panel was measured outdoors and resulted in 20.3 dBiC gain at boresight for the 30 element L band quarter panel. The horizontally polarized gain for a phase compensated quadrant is compared against a uncompensated quadrant in Figure 6.9. This indicates a 5.7 dBiC single element gain based on the array factor. The single element boresight gain is reduced from the initial measurements of approximately 8 dBiC, shown in Figure 6.4. The cut loop single element gain decreases further during the time elapsed between the outdoor measurement and the compact range measurement of the antenna arrays. This is discussed further in the next section.

![Figure 6.9: Single quadrant gain for horizontal polarization](image)

Figure 6.9: Single quadrant gain for horizontal polarization
6.1.3 Antenna Measurements

The integrated multi-band antenna was tested at the compact range at the ElectroScience Laboratory at The Ohio State University [48]. These tests measured the calibrated gain and far field radiation patterns for the S band array. The S band array radiation patterns were measured with multiple levels of blockage from the L band antenna array, see Figure 6.1. The L band array radiation patterns were not measured. The L band array uses fewer elements (120 vs 172) than the S band, but otherwise uses the same array, FSS, and feed network design. Measurements of the S band array should give some indication as to the L band performance.

The calibrated gain measured in the compact range was lower than expected. This was due to increased resistivity observed in the individual radiating element. The resistivity across the gap in the cut loop element was measured at less than 5 Ω at time of delivery. The resistance across the gap at the time of the compact range measurement was approximately 90 Ω. An individual cut loop element produced 7-9 dBiC of gain, shown previously in Figure 6.4. The coaxial cables and power divider used to combine the 4 quarter panels were found to produce 4 dB of loss. The S band array feed network was measured with 1 dB of loss at the resonating frequency. The compact range calibrated gain is consistent with the ESP 5.4 simulation when these losses are accounted for.

The normalized radiation pattern of the S band array is shown in Figure 6.10. The beamwidth of the main beam is consistent between the measured radiation pattern and the ESP 5.4 results. This indicates that the array is achieving the expected directivity since this is related to the 3 dB beamwidth [21]. This antenna array
pattern indicates that the feed network is providing a near uniform power distribution and that the phase compensation is working as expected.

Figure 6.10: S band array radiation pattern, free space

The S band array radiation patterns with various amounts of blockage from the L band are shown in Figure 6.11 through 6.14. The main beam shifts by 0.5° as the L band array obscures the main beam from the opposing direction. The L band array provides -2.5 dB of attenuation at full blockage (Figure 6.14), but preserves the axial ratio across the main beam, shown later. The first sidelobes are approximately -12.5 dB below the free space boresight gain. The L band array blockage does not reduce the side lobe levels and causes the sidelobe levels to rise relative to the main beam when blockage is present.
Figure 6.11: S band array radiation pattern, 25% blockage
Figure 6.12: S band array radiation pattern, 50% blockage
Figure 6.13: S band array radiation pattern, 75% blockage
Figure 6.14: S band array radiation pattern, 100% blockage
The boresight axial ratio over the S band array bandwidth is shown in Figure 6.15a with different amounts of blockage. The L band array is rotated toward the negative azimuth as the blockage is decreased. The boresight axial ratio is better than -1.5 for over 300 MHz. The main beam axial ratio at 2230 MHz is given in Figure 6.15b. The axial ratio is better than -1.3 over the 3dB beamwidth. The axial ratio begins to decay at \(-4^\circ\) azimuth which is consistent with the location of the first null of the radiation pattern.

The RHCP boresight gain attenuation for the X band commercial off the shelf (COTS) antenna is shown in Figure 6.16. Simulations was not performed at this frequency band. The structure is electrically large at this frequency and the computation time would have been prohibitive. These measured results demonstrate that the L and S band arrays provide less attenuation at higher frequencies. The L and S band array blockage causes the X band antenna axial ratio to shift by less than 10%, see Figure 6.17.

6.2 Summary

Balanced feed networks have been designed, built, and tested for circularly polarized narrowband L and S band arrays. These feed networks have been designed to be transparent by use of previously presented techniques. The feed network is a series feed network with unequal phase output and uses a phase compensation technique that takes advantage of circular polarization. The feed network is on a separate layer from the array and uses the parallel strip method to eliminate a PEC ground plane.

Measurements have been performed on three high gain (approx. 30 dBiC) antennas operating in close proximity. Each antenna is capable of independent pointing and
Figure 6.15: S band array axial ratio

(a) boresight over bandwidth

(b) main beam at 2230 MHz
Figure 6.16: Normalized RHCP boresight gain at X band with L and S band array blockage
Figure 6.17: X band axial ratio with L and S band array blockage
provides minimal attenuation to other antennas despite obscuring their line of sight. This system reduces the footprint needed to operate three independently pointing antennas by a factor of 3. This reduction allows for a system upgrade while simultaneously maintaining capability of legacy systems. This is a distinct advantage when deploying this system on platforms with stringent real estate requirements.
CHAPTER 7

SUMMARY AND CONCLUSIONS

Several techniques for the design of array feeding networks are presented with the goal of improving the transmission coefficient for incident plane waves. These range from simple techniques, such as reducing the ground plane size; to more complex, such as using unequal phase compensation. An optimized feed network design for a combined FSS and feed network is developed. A non-optimized feed network is developed using classical design techniques. The non-optimized feed network uses several transparency techniques. This non-optimized feed network is simulated, integrated, and measured.

The feed network provides a controlled power distribution to a CP antenna array. This is accomplished by means of individual rotation of each array element in order to compensate for the feed networks unequal phase output. This feed network results in better transparency near the array radiating frequency. This allows the underlying antennas to operate at closer relative frequency.

7.1 Results

Several examples of typical corporate feed networks have been designed and simulated in ESP 5.4. These feed networks generally have poor transparency below
the radiating frequency of the antenna array fed by the feed network \((T_\theta < -1\, \text{dB}, T_\phi < -1\, \text{dB} \text{ for } f < 1.3 f_0)\). The feed network blockage can also have an effect on the polarization state of the RHCP radiation. A corporate feed network was designed that was capable of transparent behavior with polarization purity \((T_\theta > -1\, \text{dB}, T_\phi > -1\, \text{dB}, -1.5 \leq \text{AR} \leq -1)\) for \(f > 0.95 f_0\). This would allow underlying antennas with higher operating frequencies to radiate without blockage.

Feed network designs were developed which preserved the polarization of a RHCP signal under blockage conditions. These designs relied on a technique that provided far field phase matching for a circularly polarized antenna array. This technique requires prior calculation of the feed network non-uniform phase output, but provides similar radiating characteristics. This technique also improves the antenna array radiating axial ratio \((\text{AR} = -2.18 \text{ to } -1)\) as a result of the “pseudo-random” orientation of the array elements. The feed network design examples presented offer several improvements on the traditional corporate feed design. The RHCP radiation of an antenna is preserved when blocked by these feed networks \((-1.5 \leq \text{AR} \leq -1.0)\).

Several of these techniques have been implemented on L and S band antenna arrays for a multi-band antenna system. The L band array blocks the S band array line of sight under certain pointing angles. The L band array attenuates the peak gain of the S band array but preserves the beamwidth. This attenuation is limited to the main beam and results in the sidelobe levels increasing with respect to the main beam under blockage. These antenna arrays yield RHCP radiation with an axial ratio better than -1.3 over the operating bandwidth and half power beamwidth, regardless of blockage conditions. The L band and S band array each provide approximately 1 dB of attenuation while preserving the X band RHCP polarization.
7.2 Further work

Measurement of the other feed network types developed here would help validate the ESP 5.4 simulations performed. The combined FSS and feed network is a new concept that would benefit from measured data for comparison against simulations.

The multi-band antenna has areas in need of further development. The manufacturing process yielded defects which resulted in deterioration of gain and reliability. These issues need to be resolved before the multi-band antenna can be used in a real world environment.

Clearly, many more potential techniques for improving the transmission coefficient exist. As is the case with many design problems, the solution is not unique and other potential approaches can be investigated.
APPENDIX A

COUPLING RATIOS

We can feed an MxN array using a main line with M outputs, and M branch lines with N outputs each. The output of port m of the main line is connected to a branch line input. This MxN feed network can also be combined with 3 other similar feed networks to feed an array of 2Mx2N. We can also subdivide the feed network into 4 feed networks of $\frac{M}{2} \times \frac{N}{2}$.

The calculation of these coupling ratios assume zero insertion loss. The coupling ratios are calculated for the feed network of Chapter 6, where the couplers are connected in series using the coupled port as an output, and connecting the through port to the input of the following coupler.

A.1 MxN Array

For a uniform distribution array, we desire an power output of $\frac{1}{MxN}$ at each feed network port. The main line requires M-1 couplers, and the branch line requires N-1 couplers. For both the main and branch lines, the final coupler in the series is a 3 dB coupler. Both coupled and through ports yield the same amplitude output. The coupled port is connected in the typical manner, and the through port is connected to the next output in series.
The coupling ratio of the individual couplers 1 through $m$ of the main line is $C_{m,0}$. We desire the total output $O_m$ measured from the main line to be $\frac{1}{M}$. The branch line outputs $B_n$ should be $\frac{1}{N}$ of the original signal when the branch line is measured independent of the main line. When the branch lines are connected to the main lines, the final output at each feed port will yield $\frac{1}{MN}$

If the individual coupling ratios for the main line are set to,

$$C_{m,0} = \frac{1}{M + 1 - m}$$ \hspace{1cm} (A.1)

then the power of the signal fed into coupler $m$ will be the product of the through outputs of all the preceding couplers.

$$I_{m,0} = \prod_{1}^{m-1} (1 - C_{k,0})$$ \hspace{1cm} (A.2)

This results in an output at port $m$ of the main line:

$$O_m = \frac{1}{M} = C_{m,0} \prod_{1}^{m-1} (1 - C_{k,0})$$ \hspace{1cm} (A.3)

Similarly for the individual microstrip couplers 1 through $n$ of branch $m$, the coupling ratios are

$$C_{m,n} = \frac{1}{N + 1 - n}$$ \hspace{1cm} (A.4)

where the input for that coupler is

$$I_{m,n} = \prod_{1}^{n-1} (1 - C_{m,k})$$ \hspace{1cm} (A.5)

This yields a measured branch line output

$$B_n = C_{m,n} \prod_{1}^{n-1} (1 - C_{m,k}) = \frac{1}{N}$$ \hspace{1cm} (A.6)
### Table A.1: Coupling ratios for a 6 element main line

<table>
<thead>
<tr>
<th>Main Index, $m$</th>
<th>$1$</th>
<th>$2$</th>
<th>$3$</th>
<th>$4$</th>
<th>$5$</th>
<th>$6$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{m,0}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{5}$</td>
<td>$\frac{1}{4}$</td>
<td>$\frac{1}{3}$</td>
<td>$\frac{1}{2}$</td>
<td>$1$</td>
</tr>
<tr>
<td>$1 - C_{m-1,0}$</td>
<td>n/a</td>
<td>$\frac{5}{6}$</td>
<td>$\frac{4}{5}$</td>
<td>$\frac{3}{4}$</td>
<td>$\frac{2}{3}$</td>
<td>$\frac{1}{2}$</td>
</tr>
<tr>
<td>$O_{m,0}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
</tr>
</tbody>
</table>

The final result is a measured power output for the feed network of

$$F_{m,n} = O_mB_n$$  \hspace{1cm} (A.7)

$$F_{m,n} = C_{m,0}C_{m,n} \left[ \prod_{1}^{m-1} (1 - C_{k,0}) \right] \left[ \prod_{1}^{n-1} (1 - C_{m,l}) \right]$$  \hspace{1cm} (A.8)

$$F_{m,n} = \frac{1}{MN}$$  \hspace{1cm} (A.9)

#### A.1.1 6x5 array

Assume we have a 30 element array with $M=6$, $N=5$. We can feed this with a main branch of 6 output ports, and branch lines with 5 ports each. The main line will provide the branch lines with a signal of $O_m = \frac{1}{6}$ power of the original signal. The individual coupling ratios couple tighter to the signal as they progress further from the input as seen in Table A.1.1.

The branch line ports yield outputs of $B_n = \frac{1}{5}$. The calculated coupling ratios for the branch line are given in Table A.1.1. When this output $B_n$ is combined with the input signal provided by the main trunk $O_m$, we get a signal $F_{m,n} = \frac{1}{30}$. 

120
Table A.2: Coupling ratios for a 5 element branch line

<table>
<thead>
<tr>
<th>Branch Index, n</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{m,n}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{4}$</td>
<td>$\frac{1}{3}$</td>
<td>$\frac{1}{2}$</td>
<td>1</td>
</tr>
<tr>
<td>$1 - C_{m,n-1}$</td>
<td>n/a</td>
<td>$\frac{4}{5}$</td>
<td>$\frac{3}{4}$</td>
<td>$\frac{2}{3}$</td>
<td>$\frac{1}{2}$</td>
</tr>
<tr>
<td>$B_{m,n}$</td>
<td>$\frac{1}{5}$</td>
<td>$\frac{1}{5}$</td>
<td>$\frac{1}{5}$</td>
<td>$\frac{1}{5}$</td>
<td>$\frac{1}{5}$</td>
</tr>
<tr>
<td>$O_{m,0}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
<td>$\frac{1}{6}$</td>
</tr>
<tr>
<td>$F_{m,n}$</td>
<td>$\frac{1}{30}$</td>
<td>$\frac{1}{30}$</td>
<td>$\frac{1}{30}$</td>
<td>$\frac{1}{30}$</td>
<td>$\frac{1}{30}$</td>
</tr>
</tbody>
</table>

### A.2 Polygonal Array

For a polygonal array such as the one used in Chapter 6, we define $N_m$ as the number of elements in branch line $m$. The calculation of the branch line coupling ratios will remain similar with the substitution of $N_m$ for $N$.

The main line coupling ratios need to be weighted to account for the number of elements in each row. Branch lines with more elements need a larger percentage of the signal, and elements with fewer elements do not need as much.

\[
O_m = \frac{N_m}{\sum_{k=1}^{M} N_k} \quad (A.10)
\]

The input to the coupler from the preceding coupler can be expressed as

\[
I_{m,0} = \frac{\sum_{l=1}^{M} N_l}{\sum_{k=1}^{M} N_k} \quad (A.11)
\]
We can express the main line output as

\[ O_m = C_{m,0} I_n \]  \hspace{1cm} (A.12)

\[ \frac{N_m}{\sum_i^M N_k} = C_{m,0} \frac{\sum_i^M N_l}{\sum_i^M N_k} \]  \hspace{1cm} (A.13)

\[ C_{m,0} = \frac{N_m}{\sum_i^M N_l} \]  \hspace{1cm} (A.14)

### A.2.1 30 element Polygon Array

Given a 6x6 planar array with 6 elements removed from one corner, \( N_m = \{6, 6, 6, 5, 4, 3\} \), we have a 5 sided array with 30 elements. Combined with 3 similar arrays, we have the 120 element hexagon L band array used in Chapter 6.

As mentioned previously, the coupling ratios for the branch lines are similar to the MxN case. The 6 element branch lines have coupling ratios identical to the main lines of MxN case, see Table A.1.1. The 5 element branch line coupling ratios are identical to the 5 element branch lines of the MxN case, see A.2.1. The 4 and 3 element branch lines coupling ratios are simply the last 4 and 3 coupling ratios of any of the longer branch lines.

### A.3 Final Notes

Keep in mind that when designing microstrip couplers, that the design equations may be given in voltage ratios rather than power ratios. These calculated coupling ratios assume zero loss. The coupling ratios are used to calculate the microstrip coupler dimensions using Equations (4.6), (4.7), and (3.7) to calculate the required impedance and then applying standard microstrip transmission line equations [24]. These microstrip dimensions were simulated in HP-ADS Momentum using the manufacturer specifications for the dielectric, including loss tangent. The coupling ratios
<table>
<thead>
<tr>
<th>Main</th>
<th>0</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>m</td>
<td>1</td>
<td>(\frac{1}{5})</td>
<td>(\frac{1}{5})</td>
<td>(\frac{1}{4})</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{2})</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>(\frac{1}{7})</td>
<td>(\frac{1}{6})</td>
<td>(\frac{1}{5})</td>
<td>(\frac{1}{4})</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{2})</td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{4})</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{2})</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>(\frac{1}{12})</td>
<td>(\frac{1}{5})</td>
<td>(\frac{1}{4})</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{2})</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>(\frac{4}{7})</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{2})</td>
<td>1</td>
<td>n/a</td>
<td>n/a</td>
</tr>
<tr>
<td></td>
<td>6</td>
<td>(\frac{1}{3})</td>
<td>(\frac{1}{2})</td>
<td>1</td>
<td>n/a</td>
<td>n/a</td>
<td>n/a</td>
</tr>
</tbody>
</table>

Table A.3: Coupling ratios for a 30 element Polygonal Array

and coupler dimensions were refined to yield uniform distribution for each series feed network through iterative simulations.
BIBLIOGRAPHY


