CALIFORNIA STATE UNIVERSITY, NORTHRIDGE

A Circularly Polarized Microstrip Antenna using an Open Stub Coupled-Line Phase
Shifter

A graduate project submitted in partial fulfillment of the requirements

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By

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Abstract

A Circularly Polarized Microstrip Antenna using an Open Stub Coupled-Line Phase Shifter

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Master of Science in Electrical Engineering

The design, simulation, fabrication, and measurement of a proximity-fed circularly polarized microstrip antenna is presented. The two major design components are (1) a stepped impedance open stub (SIOS) coupled line 90° phase shifter, and (2) a dual probe proximity fed circular patch antenna. The 90° phase shifter is a 4-port network consisting of two microstrip lines: the SIOS with coupled line, and a reference line. The antenna is a 64.4 mm diameter circular patch printed on top of a 140 mm x 140 mm x 0.8 mm upper layer FR4 substrate. The antenna is fed by a cascaded phase shifter and Wilkinson power divider on a 170 mm x 170 mm x 0.8 mm lower layer FR4 substrate. The top side of the lower substrate contains the ground plane and the feed network is printed on the bottom side. Connecting the upper and lower substrates are feeding probes which extend from the two outputs of the phase shifter to the bottom of the antenna substrate through 16.7 mm air

layer. Simulation results show a 3-dB axial ratio (AR) bandwidth of 22.2% from 1.68 GHz to 2.1 GHz, a 2-dB AR bandwidth of 19.3% from 1.69 GHz to 2.05 GHz, and a 10-dB return loss bandwidth of 60.9% from 1.6 GHz to 3.0 GHz. A prototype was fabricated and measured results show a 3-dB AR bandwidth of 38% from 1.66 GHz to 2.44 GHz, a 2-dB AR bandwidth of 14% from 1.8 GHz to 2.07 GHz, and a 10-dB return loss bandwidth of 60.9% from 1.6 GHz to 3.0 GHz.

Chapter 1: Introduction

1.1 Background

Microstrip antennas are commonly used in mobile devices, GPS, RADAR, and various other communication and navigation systems. Also known as patch antennas, these types of antennas are often attractive because they are low-profile, light weight, low-cost, and easy to fabricate. The basic components of a microstrip antenna are an etched conductor, a dielectric substrate, and a ground plane as shown in Figure 1.1. The etched conductor is a thin copper layer on top of the dielectric substrate that consists of a radiating patch and a feeding structure which is the connection between the antenna and its source. The dielectric substrate is an insulator or poor conductor sandwiched between the two thin metal layers (the antenna/feed and ground plane). When the patch is excited, an electric field is generated in the substrate mostly underneath the antenna, but some of the electric fields also exists above and around the edges of the antenna. These fields, called fringing fields, gives rise to microstrip antenna radiation. From Maxwell's equations, when there's a time-varying electric field, there's also an associated magnetic field, and as a result, an electromagnetic (EM) wave is produced and radiated from the antenna.

Figure 1.1 Microstrip antenna using a microstrip line feed

1.2 Antenna Parameters

1.2.1 Polarization

An important characteristic of antennas and EM waves is polarization, which is the orientation of the electric field. The two common forms of antenna polarization are linear and circular. A linearly polarized (LP) antenna is often identified as being either vertically or horizontally polarized. In linear polarization, the electric field oscillates up and down in a straight line on a one-dimensional plane. Circular polarization (CP) occurs when the two orthogonal electric field components have equal magnitude and a phase difference of $\pm 90^\circ$. A circularly polarized wave propagates in a circular fashion either in a clockwise or counterclockwise direction. If the thumb of the right hand points in the direction of propagation and fingers curve in the direction of the electric field rotation, this indicates a right hand circular (RHC) polarized wave. A left hand circular (LHC) polarized wave would be rotating in the opposite direction. Polarization is important because received signal strength is reduced when polarizations between the transmitting antenna and the receiving antenna are different. If the polarization of the receiving antenna is orthogonal to the transmitting antenna (e.g., if one is horizontal and the other is vertical), no signal is received. Circular polarization on the other hand is more robust in terms of antenna orientation compared to linear polarization. Multipath reflections and losses due to antenna orientation angles, inclement weather, and atmospheric conditions can be mitigated with CP antennas, effectively optimizing the received signal. However, a common challenge faced when designing circularly polarized microstrip antennas is achieving wide axial ratio (AR) and return loss bandwidths, which are defined and discussed in the following sections.

1.2.2 Axial Ratio

A measure of an antenna's circular polarization quality is the axial ratio. The closer the axial ratio is to 1 (or 0 dB), the closer it is to ideal circular polarization. An axial ratio of 1 means the magnitude of the orthogonal electric field components are equal. As mentioned earlier, circular polarization is achieved when the orthogonal electric field components are equal in magnitude and have a 90° difference in phase. Axial ratio in a mathematical description is the ratio of the major axis to the minor axis of the electric field ellipse:

$$AR = \frac{\text{major axis}}{\text{minor axis}} \quad (1.1)$$

If the axes are the same, the ratio becomes one indicating circular polarization. If the axial ratio is infinity, the polarization is linear. The frequency range over which acceptable circular polarization exists is the axial ratio bandwidth which is typically measured at 3 dB or less. This is also called the polarization (or co-polarization) bandwidth. Cross-polarization is the opposite, or the orthogonal, orientation of co-polarization. Low cross-polarization is desired and can also be another measure of antenna polarization performance.

1.2.3 Return Loss

Return loss is the reflected power due to impedance mismatches. Return loss is typically expressed in decibels as:

$$RL = -20\log|\Gamma| \text{ dB} \quad (1.2)$$

where Γ is the reflection coefficient. Impedance mismatches are typically measured using a vector network analyzer (VNA) and are described using the scattering parameter (S-parameter), S_{11} . S_{11} is the reflection coefficient seen at port 1 (input port) and has both a

magnitude and phase. Voltage standing wave ratio (VSWR) is the ratio of the maximum to minimum voltage wave amplitude between the load and generator. VSWR and $|S_{11}|$ are related and are interchangeable parameters to describe the impedance matched performance of an antenna. Often, a VSWR of 2 or less (1 is ideal) or an $|S_{11}|$ of -10 dB or less is considered matched (an acceptably minimal reflection). From the equation:

$$VSWR = \frac{V_{max}}{V_{min}} = \frac{1+|\Gamma|}{1-|\Gamma|} \quad (1.3)$$

a return loss of 10 dB is equivalent to a VSWR of 1.92 and a return loss of infinity translates to no reflection (a VSWR of 1). A return loss of 0 dB is total reflection which means all the incident power going to a transmitting antenna would be reflected back to the source. The concept behind impedance matching is to maximize power transfer by minimizing reflection so that all power generated is delivered to the load.

1.2.4 Directivity and Gain

Directivity is the ratio of maximum power density radiated to that of an isotropic antenna which radiates equally in all directions [1]. Gain is the ratio of maximum power density radiated to power supplied to the antenna over the surface area of a sphere [1]. The difference between gain and directivity is that gain accounts for antenna losses. Directivity does not, so gain would equal directivity if there were no losses. Antenna losses can be from internal losses due to the substrate and the conductor. Thus, gain is an important parameter because it characterizes the efficiency and performance of the antenna. Gain can be explicitly defined, as:

$$G = \eta_{rad}D \quad (1.4)$$

where gain (G) is the product of radiation efficiency (η_{rad}) and directivity (D) [2]. Radiation efficiency is the ratio of radiated power to the power supplied to the antenna [1]. Impedance

mismatches are accounted for in realized gain. Microstrip antenna gain is typically in the range of 6 to 9 dBi (decibels relative to isotropic). The gain of a microstrip antenna can be increased by integrating multiple antennas or antenna elements into an antenna array. A common type is a phased array which uses phase shifters to electronically steer the antenna beam by controlling the phase in each antenna element.

1.2.5 Radiation Pattern

The radiation pattern paints a picture of what the antenna transmit and receive beam looks like and is usually described by polar or rectangular plots of various planes of the 3D pattern (often called “cuts”). Figure 1.2 shows an example of an E-plane antenna pattern using a polar plot. The straight dashed lines from the center measure angles. The circular dashed lines are power levels in decibels. The antenna pattern reveals many important details about the antenna characteristics such as the mainlobe, sidelobes, directivity, and 3-dB beamwidth. This information can be used to determine the type of antenna and its application. The mainlobe is the lobe with the highest power level. The lower level peaks on each side of the mainlobe are called sidelobes. The 3-dB beamwidth is the width of the mainlobe at the half-power levels (a 3 dB drop from the maximum level), which is commonly used as a measure of antenna beamwidth. Based on the beamwidth of the mainlobe and the size of its sidelobes, the directivity and antenna application can be approximated. For instance, a narrow beam pattern with very low sidelobes would imply high directivity and potential usefulness in radar applications.

Figure 1.2 Antenna radiation pattern on a polar plot (from [2])

1.3 Design Goals

The goal of this project was to design an antenna for space satellite applications operating in the L and S band. More specifically, the objective was to design and fabricate a circular polarized microstrip antenna (single element) with the following performance requirements:

- Operate at least from 1.75 to 2.12 GHz (19.5 % bandwidth) with
 - $|S_{11}| < -10$ dB
 - $AR < 3$ dB
 - Gain of at least 6 dBic

1.4 Report Overview

This report will cover the design, simulation, fabrication, and measurements of a new circularly polarized microstrip antenna. Different antenna and feed network options will also be examined and compared with in terms of its design configuration, advantages, disadvantages, and simulation performance.

Chapter 2: CP Antenna Feed Methods and Techniques

The signal that the patch antenna radiates must be first fed to the antenna. The common feed methods used are coaxial (also known as probe or pin fed), microstrip line, and aperture coupled. Coaxial is a direct injection method where one end of the probe makes direct contact with the antenna as shown in Figure 2.1. When fabricating a probe-fed patch antenna, there will be drilling involved because the probe must go through the substrate and be soldered onto the patch. One of the drawbacks with using coaxial is that the inductance from the length probe can cause matching problems. The helpful advantage with coaxial is that the wire can be easily located for optimal impedance matching. A microstrip line feed shown in Figure 1.1 is another direct contact technique, but instead of running a probe through the substrate, a transmission line is printed on the same surface as the patch where one end connects to the patch, making it easy to fabricate. The disadvantage with the microstrip line method is that the feed line's proximity to the patch can cause an increase in spurious radiation and cross polarization effects.

Figure 2.1 Coaxial feed method (side view and top view)

Unlike microstrip line and coaxial feeds, aperture coupling feeds the antenna indirectly by way of proximity coupling where EM fields are coupled or transferred from the feed to the antenna through an aperture. This feed method does not have physical contact with the antenna. Instead, the EM fields from the feed are coupled through an aperture slot cut into the ground plane. Aperture coupling consists of three main layers: the upper substrate containing the patch, the lower substrate containing feed line, and an aperture slot-etched ground plane sandwiched in between as shown in Figure 2.2. Proximity coupling, shown in Figure 2.3, can also be applied with coaxial and microstrip lines, but the feed does not contact the antenna. Aperture coupling is a type of proximity coupling defined by a ground plane with the aperture slot configuration in between two substrates. Proximity coupling can exhibit wide bandwidth performance, but it can also be difficult to fabricate.

Figure 2.2 Aperture coupling feed method (from [3])

Figure 2.3 Proximity coupling (from [3])

2.1 Single Feed Circularly Polarized Antennas

Microstrip antennas come in a variety of shapes from a simple square, rectangular, circular, and triangular shape, to a more unconventional shape such as the L-shaped, H-shaped, and F-shaped antenna. The different shapes are part of the various techniques used to achieve circular polarization and impedance matching. The fundamental baseline for the techniques is meeting the criteria for circular polarization: two field components of equal magnitude and 90° out of phase. Using a single feed is the simplest technique because it does not require an external feed network and therefore the antenna structure tends to be more compact in size and easier to fabricate. However, since there's only one feed, the patch design will need to be modified in a way to produce two field components of equal magnitude and 90° out of phase. This can be done using perturbation segments shown in Figure 2.4. Several examples of perturbation segments include F-shaped slot, C-shaped slot, S-shaped slot, cross-slot cut into a square patch and fed via aperture coupling. Other single feed techniques are stacked antennas, truncated corners (or edges), and a combination of stacked, truncated corners (or edges), and slots. The underlining disadvantage with using single feeds is typically a narrow bandwidth. Such is the case

with the F-shaped slot antenna in [4] which exhibited a measured 3 dB AR bandwidth of 3.2% from 2.17 to 2.24 GHz. In [5], a CP square patch with truncated corners achieved a 3-dB AR from 1.566 to 1.582 GHz. Another single fed antenna can be seen in [6] using the stacked and truncated edges technique and exhibited a 10% AR bandwidth.

Figure 2.4 Perturbation segments (from [3])

2.2 Dual Feed Circularly Polarized Antennas

Fortunately, the bandwidth limitation using single feeds can be improved by adding another feed. Dual feeds are more commonly used because the circular polarization concept is easier to implement over broader bandwidths. The feed lines typically need to be placed orthogonal to each other with both lines configured using a network that will send signals of equal magnitude and 90° out of phase. Additionally, the antenna can be a simple circular or square patch instead of truncated edges and complex shaped slots. Dual feeds can result in wider bandwidths compared to single feeds, but at the cost of size and complexity from adding an external feed network needed to generate the two signals. If dual or multi feed techniques are utilized, an external feed network is needed. An external feed network includes phase shifters, power dividers, and couplers.

2.2.1 Phase Shifter

In addition to the different feeding methods and techniques mentioned previously, there are also various 90° phase shifters designs, each with its own advantages and disadvantages. A simple phase shifter design can be made with two transmission lines of different electrical lengths. For example, one line could have an electrical length of 180° ($\lambda/2$) and the other 90° ($\lambda/4$). The phase difference between the two lines would be 90° . The drawback with this simple phase shifter design is its narrow bandwidth performance. Consequently, several alternative phase shifter designs exist with much wider bandwidth performance such as the Schiffman [7], T-shaped open stub [8], and parallel open-short stubs [9,17]. The Schiffman shown in Figure 2.5 is a widely used 90° phase shifter known for its wide-band performance, but also comes with fabrication difficulties due to very tight spacing between traces. The Schiffman is made up of a quarter-wave coupling section (C-section) and a three-quarter wave reference line. The 90° phase difference bandwidth of the Schiffman is a function of its electrical lengths and tight coupling degree. One of the advantages with the SIOS and coupling phase shifter design for this project to be discussed shortly is the loose coupling spacing.

Figure 2.5 Schiffman 90° phase shifter (from [10])

Many phase shifters have similarities in their design structures such as the T-shaped open stub and the conventional open stub design shown in Figure 2.6. Both have a main line and a reference line. In a conventional open stub design, the half-wavelength main line is loaded with an open stub in the middle and has a reference line of a half-wavelength (plus the desired phase difference). The phase difference and impedance bandwidths of the open stub phase shifter is a function of the main line and open stub component. However, a wider bandwidth was found to be based on the slow change in the electrical length of the open stub, so the T-shaped open stub phase shifter was introduced. The T-shaped open stub is made up of a quarter-wave open stub and quarter-wave triangular patches on each side, which all together appears as a half-wave open stub like the conventional open stub phase shifter. The SIOS and coupled line phase shifter for this project also uses a similar design structure (main line and reference line), but with a combination of features described in the Schiffman and open stub phase shifters. The main line has an open stub component including a coupled line used to increase the phase difference bandwidth. Phase shifters are typically the final cascaded stage in a feed network configuration which means its input signal often comes from a power divider.

Figure 2.6 Open circuit stub (left) and T-shaped open stub phase shifter (right) (from [8])

2.2.2 Power Divider

The power divider is another integral part of a feed network. A power divider basically takes an input power and splits it into two equal output powers such as the simple 3-port network known as a T-junction. A common power divider is a quadrature hybrid coupler (also known as a branch line coupler) shown in Figure 2.7. The quadrature hybrid coupler is a 4-port network and is unique in that not only does it perform equal power division functions but also provides 90° phase difference at the output ports. Power entering port 1 is split at ports 2 and 3 while no power enters the isolated port (port 4). The Wilkinson power divider shown in Figure 2.8 is also a 3-port network but has properties that makes it a popular choice in microstrip design. The Wilkinson provides (ideally) matched impedance at all three ports ($S_{11} = 0$, $S_{22} = 0$, $S_{33} = 0$) as well as isolation between the output ports ($S_{23} = S_{32} = 0$) [2]. Any reflected power from the output ports is dissipated in the $100\ \Omega$ resistor making the Wilkinson appear lossless. With good matching, the Wilkinson has a wide impedance bandwidth. However, bandwidth may still be limited for a single power divider or coupler, so many designs are seen cascaded in multiple sections to increase the bandwidth. In this project, the Wilkinson power divider was used to split power equally into the input ports of the phase shifter. With equal magnitude and phase provided by the Wilkinson, the phase shifter takes that signal and produces a phase by 90° phase difference to satisfy the required circular polarization conditions. With a basic understanding of phase shifters and power dividers, the next several sections will explore the various IEEE feed network designs and their performance.

Figure 2.7 Microstrip layout of a quadrature hybrid coupler (from [2])

Figure 2.8 Microstrip layout of the Wilkinson power divider (from [2])

2.3 Feed Network Design

During the early stages of the design, feed networks from several IEEE papers were explored and compared to see which would be best suited for this project in terms of performance as well as being feasible for design and fabrication. The feed network used in [9] consisted of a Wilkinson power divider and a parallel open-short stub phase shifter. The simulated impedance bandwidth and $90^\circ (\pm 3^\circ)$ phase difference bandwidth were reported to be 86% from 1.0 to 2.5 GHz and 76% from 1.10 to 2.45 GHz, respectively. The feed network in [10] consisting of a Wilkinson power divider and a Schiffman showed a simulated impedance bandwidth of 187.6% from 0.9 to 2.81 GHz and a $90^\circ (\pm 5^\circ)$ phase difference bandwidth of 66.7% from 1.3 to 2.6 GHz. A 90° hybrid coupler was also

simulated with the same Schiffman in [10] and compared to the Wilkinson. The hybrid coupler exhibited a narrower impedance and phase difference bandwidth of 30.9% from 1.53 to 2.09 GHz and 32% from 1.47 to 2.03 GHz respectively. The excellent performance demonstrated from these results utilizing the Wilkinson power divider along with its useful properties including comparisons with the hybrid coupler indicated that the Wilkinson was the best choice as the power divider for this project. Despite some of the excellent bandwidth performances, the noticeable disadvantage seen in a lot of the circular polarized antennas in the literature was the design complexity which involved multiple layers and cascaded circuits.

Seven different potential 90° phase shifters were designed and scrutinized based on their simplicity, 10-dB RL bandwidth, 90° phase difference bandwidth, and amplitude balance. Comparisons were made in Advanced Design System (ADS) simulations, an electronic design automation software for RF and microwave applications. ADS was used to integrate microstrip components, run simulations, and assess performance. ADS also provides a tool called ADS LineCalc that was used to calculate microstrip line dimensions. Entering the substrate information, characteristic impedance, and electrical length on ADS's LineCalc produces the designed microstrip line length and width. Table 2.1 shows ADS results for the seven different phase shifters that were designed and simulated. As described briefly in Section 2.2.1, the Schiffman, open stub, SIOS and coupled line, and T-shaped open stub phase shifters were similar in design structure. Each one differs at the main line where one has a conventional open stub, a T-shaped open stub, a coupled line or C-section (Schiffman), as well as both open stub and coupled line (SIOS and coupled line). The main line which includes the stub and coupled line components are the key enablers

for the wide phase difference bandwidth performance of each phase shifter. The parallel open-short stubs phase shifter is another example which consist of a 61.9Ω half-wave length main line with 125.6Ω one-eighth wavelength parallel open-short stubs (two open and two short stubs). The reference line is a 50Ω three-quarter wavelength transmission line. The characteristic impedance values of the main line and stubs are from Figures 2 and 3 of [17]. Out of the five phase shifters mentioned, the hybrid coupler (see section 2.2.2 and Figure 2.8) and the 3-port divider don't have the main line and reference line configuration. The 3-port divider is simply a 50Ω input port that splits into two transmission lines of an electrical length difference of 90° .

	Hybrid Coupler [3]	Simple 3-port Divider	Schiffman [7]	Parallel Open-Short Stubs [17]	Open Circuit Stub [8]	SIOS and Coupled Line [11]	T-shaped Open Circuit Stub [8]
10-dB Return Loss Bandwidth (GHz)	1.6-2.2	Above 10 dB	1.6-2.5	Fluctuates above 10 dB	1.6-2.5	1.6-2.5	1.63-2.2
$90^\circ (\pm 5^\circ)$ Phase Difference Bandwidth (GHz)	1.6-2.2	1.7-1.9	Deviates $\pm 10^\circ$ from 90°	1.6-2.5	1.6-2.5	1.6-2.5	1.63-1.84
Amplitude Difference (± 1 dB) (GHz)	1.65-2.15	1.6-2.2	1.6-2.01	1.6-2.5	1.6-2.5	1.6-2.5	1.6-2.5
Substrate	0.8 mm Rogers RO4003 ($\epsilon_r = 3.38$)	0.8 mm Rogers RO4003 ($\epsilon_r = 3.38$)	0.8 mm Rogers RO4003 ($\epsilon_r = 3.38$)	0.8 mm Rogers RO4003 ($\epsilon_r = 3.38$)	0.508 mm Rogers RO4003 ($\epsilon_r = 3.38$)	0.762 mm Rogers 4350B ($\epsilon_r = 3.48$)	0.508 mm Rogers RO4003C ($\epsilon_r = 3.38$)

Table 2.1 Comparison of phase shifters simulation results

The phase difference results were also consolidated on a single plot as shown in Figure 2.9. Based on results of ADS simulations, the SIOS and coupled line with weak coupling phase shifter was selected as the phase shifter for this project since it has the tightest 90° phase deviation of $\pm 1.5^\circ$. The ease of implementation was another factor since geometry and dimensions of the SIOS and coupled line phase shifter were simpler to design and fabricate compared to the Schiffman and T-shaped open stub phase shifters. The SIOS and coupled line phase shifter has the advantage of a loosely coupled phase shifter as opposed to the tightly spaced coupled line dimensions of the Schiffman phase shifter.

Figure 2.9 ADS simulation results displaying phase difference of various phase shifter designs

As mentioned earlier, the feed network was designed and simulated first using the ADS software. The microstrip line dimensions that produced the desired results in ADS were then used to build a more accurate model in the EM simulation software package

FEKO. FEKO contains a wide range of EM computational tools such as finite-difference time-domain (FDTD), and finite element method (FEM), but utilizes the methods of moments (MoM) as its primary solver. FEKO uses a mesh to create triangular elements on every part of the simulation model, and uses the calculated current in each triangular element to develop the radiation pattern, return loss, and other EM results. Simulation design dimensions, details, and performance for the selected feed network are described in the next sections.

2.3.1 SIOS and Coupled Line 90° Phase Shifter

The SIOS and coupled line phase shifter is 4-port network consisting of two microstrip lines sections: a main line and a reference line. A diagram of the SIOS and coupled line phase shifter is shown on Figure 2.10. The main line contains port one and port two. Port three and port four are for the reference line. The main line is made up of an SIOS and an odd-mode coupled line. The main line has an electrical length of 90° with stepped characteristic impedance, Z_1 of 72.9 Ω and an open stub characteristic impedance, Z_2 of 17.1 Ω . The coupled line used an odd mode characteristic impedance, Z_o of 25.3 Ω with a coupling coefficient of 11.7 dB (weak coupling compared to the Schiffman).

Figure 2.10 Diagram of the SIOS and coupled line phase shifter (dimensions unit are in mm)

If the main line consisted of only the SIOS, according to [11], the impedance bandwidth increases as the impedance ratio is increased. However, that simultaneously degrades the phase difference performance. By adding the coupled line to the SIOS, the inverse effects of the increased impedance ratio on the phase difference bandwidth is negated while still maintaining wide impedance bandwidth. This illustrates that the wide impedance and phase difference bandwidths is a direct function of the impedance ratio (g) and coupling coefficient (p) parameters of the SIOS and coupled line respectively:

$$g = \frac{(Z_1 - Z_2)}{(Z_1 + Z_2)} \quad (g \geq 0) \quad p = \frac{Z_e}{Z_o} \quad (p \geq 1) \quad (2.1)$$

where Z_1 is the characteristic stepped impedance and Z_2 is the characteristic impedance of the open stub; Z_o is the odd mode and Z_e is the even mode characteristic impedance of the coupled line. Initial conditions for g and p were chosen to be 0 and 1 respectively to simplify calculations to $Z_1 = Z_2 = Z_s = 45 \, \Omega$ and $Z_e = Z_o = Z_m = 33 \, \Omega$. The stub impedance, Z_s and main line impedance, Z_m are the optimum values for a 90° differential phase shifter from Table 1 in [8]. The SIOS and coupled line characteristic impedances and the new values of g and p were then computed using the following equations:

$$Z_1 = Z_s (1 + g) \quad Z_2 = Z_s (1 - g) \quad (2.2)$$

$$Z_e = Z_m \sqrt{p} \quad Z_o = \frac{Z_m}{\sqrt{p}} \quad (2.3)$$

The calculated design parameters can also be found on Table 1 of Example B in [11]. The reference line has an electrical length of 270° and a characteristic impedance of $50 \, \Omega$. All four ports have a characteristics impedance of $50 \, \Omega$. For design and fabrication considerations, microstrip lines were added to each side of the coupled line parallel to the

reference line. However, the additional microstrip lines had also increased the total electrical length of the main line. In order to maintain the 90° and 270° electrical lengths between the mainline and reference line respectively, an equivalent electrical length adjustment was added to the reference. This resulted in the same phase difference as the unmodified configuration while having more design and fabrication flexibility.

The design parameters were based on a center frequency of 1.9 GHz and using a 0.8 mm thick FR4 substrate with a relative permittivity of 4 and a loss tangent of 0.018. Despite its high loss properties and different permittivity values, FR4 was used because it was inexpensive and easy to have manufactured. Using these design parameters including electrical lengths, characteristic impedances, and coupling coefficient, ADS microstrip lengths, widths, and coupling spacing dimensions were determined. The microstrip components were then placed in the ADS schematic and connected as shown in Figure 2.11.

The same dimensions were used in FEKO and later adjusted to optimize the results. Edge ports were used in FEKO to represent the input and output ports and were also the reference points for the simulation, as shown on Figure 2.12. S-Parameter calculations were configured with $50\ \Omega$ impedance for all four ports. The S-Parameter solutions were viewed in POSTFEKO using various plots. In addition, a phase shifter of a similar design and configuration, but without the SIOS and coupled line components was constructed for performance comparison. This design shown on Figure 2.13 simply consisted of two transmission lines with two electrical lengths that differed by 90° ($\lambda/4$).

Figure 2.11 ADS schematic of the SIOS coupled-line phase shifter

Figure 2.12 Edge port used in FEKO at the input and output ports

Figure 2.13 Phase shifter without the SIOS and coupled line for performance comparison

2.3.2 Wilkinson Power Divider

In order to provide the two input signals required by the phase shifter, a Wilkinson power divider was designed using the same process and substrate as the phase shifter including the use of edge ports. The design started on ADS as shown on Figure 2.14. and then was fine-tuned in FEKO, shown on Figure 2.15. The Wilkinson is a 3-port network consisting of one input and two output ports. All the ports have a characteristic impedance of $50\ \Omega$. The input line was extended in order to properly solder the center pin of the SMA connector. Attached to the input port forming a wedge (“V” configuration) are two quarter-wave rectangular microstrip lines with characteristic impedances of $\sqrt{2}Z_0$ or $70.71\ \Omega$. This configuration can also be seen in Figure 2.7. At the ends, and in between the wedge, are two $100\ \Omega$ microstrip lines to support the $100\ \Omega$ surface mount resistor. The $100\ \Omega$ lines form the connection between the $100\ \Omega$ resistor, two quarter-wave microstrip lines, and $50\ \Omega$ output ports. In addition to the two $100\ \Omega$ lines are two small rectangular microstrip paddings designated as soldering points for the surface mount resistor as shown on Figure 2.16. A wire port used to model the $100\ \Omega$ resistor is also shown on Figure 2.16. The

output port traces were extended in order to create distance between the two $100\ \Omega$ lines and the edge of the substrate.

Figure 2.14 ADS schematic of the Wilkinson power divider

Figure 2.15 FEKO model of the Wilkinson power divider

Figure 2.16 FEKO model of the Wilkinson's $100\ \Omega$ resistor using a wire port

Figures 2.17 and 2.18 show the integration of the Wilkinson power divider and the phase shifter in ADS and FEKO, respectively. The feed network is a $138.9\text{ mm} \times 65.05\text{ mm} \times 0.8\text{ mm}$ 3-port network consisting of an input port from the Wilkinson and two output ports from the phase shifter. The phase shifter's output ports are the feed network's direct connection to the load, which will be the antenna. If the impedance at the output ports does not match the input impedance at the ends of the probes which feed into the antenna, then only a fraction of the power generated will be delivered. The main function of the SIOS and coupled line phase shifter is to provide the antenna with two feeds that have an accurate 90° phase difference. The Wilkinson power divider that is connected to the phase shifter ensures an equal power split and isolation between output ports. Essentially, the phase shifter provides the 90° phase difference and the Wilkinson provides the equal magnitude needed to produce currents on the patch that results in circular polarization.

Figure 2.17 ADS schematic of the feed network

Figure 2.18 FEKO model of the feed network

2.3.3 Simulation Results

In this section the simulation results presented are of the feed network (integration of the phase shifter and Wilkinson power divider). Figure 2.19 shows an ADS plot of how well the feed network is matched based on the schematic from Figure 2.17. The input port is matched as shown by the $|S_{11}|$ below -10 dB across the band. The power level difference between the outputs can be seen in the amplitude balance plot shown on Figure 2.20 which shows about a 0.35 dB deviation. Figure 2.21 shows the phase difference between the output ports deviating from a minimum 88.6° to a maximum of 96.3°.

Figure 2.19 ADS simulation input impedance matching results

Figure 2.20 ADS simulated feed network amplitude difference between the outputs

Figure 2.21 ADS simulated feed network phase difference results

The next several figures show the simulation results in FEKO. Figure 2.22 shows the $|S_{11}|$ below -10 dB from 1.68 to 2.4 GHz and an amplitude difference of 1.1 dB shown

on Figure 2.23. The phase difference shown on Figure 2.24 deviates between 81° and 89° from 1.68 to 2.4 GHz. Figure 2.24 also shows how the phase shifter on Figure 2.13 compares to the SIOS and coupled line phase shifter. Results show that the phase difference of the phase shifter without the SIOS and coupled line at $90^\circ(\pm 5)$ spans only from 1.95 to 2.17 GHz whereas the proposed phase shifter's deviation at $90^\circ(\pm 5)$ widens from 1.68 to 2.3 GHz .

Figure 2.22 FEKO simulated feed network input impedance matching results

Figure 2.23 FEKO simulated feed network amplitude difference between output ports

Figure 2.24 FEKO simulated feed network phase difference between designs shown in Figure 2.10 and Figure 2.13

2.3.4 Fabrication

The completed feed design was fabricated on a printed circuit board (PCB) through PCBway.com online. Prior to placing an order on PCBway.com, Gerber files had to be created, zipped, and uploaded. Three Gerber files were required to be uploaded for the two-layered (two sided) substrate: outline, feed network, and ground plane. The outline file contains the outer borderlines of the substrate material. The feed network file contains the feed network copper traces on the bottom side of the substrate shown in Figure 2.25. The ground plane file contained a filled-in rectangle as shown in Figure 2.26.

Figure 2.25 PCB feed network copper trace on ADS Layout

Figure 2.26 PCB ground layer on ADS Layout

To generate the Gerber files, the feed network design was exported from FEKO (.dxf file extension) and imported to ADS Layout. The outline file was generated by drawing a rectangle on ADS Layout. Gerber files were uploaded after a list of PCB specifications were selected on PCBway.com. Certain items such as solder masks and silk screen were optional and can add cost to the order as well as create differences from the simulation model. Solder mask is a protective coating of insulating material added to the entire board except the soldering areas, which wasn't included in the simulation model and could change the performance of the circuit. Cost was the main reason for using FR4. FR4 is a commonly used substrate and was much cheaper to manufacture at around \$30 compared to the approximately \$300 Rogers substrate fabrication cost. However, this

benefit also comes at a cost of a very lossy material indicated by its 0.018 loss tangent and a large tolerance on permittivity values. The fabricated PCB prototype is shown in Figure 2.27.

Figure 2.27 PCB prototype of the feed network

Four parts were soldered onto the feed network PCB: three female subminiature version A (SMA) connectors and one 0603 surface mounted device (SMD). The 0603 SMD is a 1.626 mm x 0.813 mm x 0.381 mm, 100 Ω ($\pm 0.1\%$ tolerance), high frequency Vishay thin film resistor ordered from digikey.com. SMA is a commonly used 50 Ω coaxial RF connector rated up to 18 GHz. The signal pin was soldered onto the input port trace of the Wilkinson while the other ground pins were soldered to the ground plane. The other two SMA connectors were soldered onto the output ports of the phase shifter applying the same process as the first SMA connector. The SMD was soldered in place on the designated pads of the Wilkinson power divider. The soldered feed network is shown on Figure 2.28.

Figure 2.28 Fabricated feed network with components soldered

The feed network PCB was measured using a 2-port VNA. The VNA and its cables and connectors had to be calibrated first to remove any error for accurate measurements. Following successful calibration, the input port was connected to one cable and an output port to another. A second output port was terminated with a $50\ \Omega$ load. All the connectors were tightened with an RF torque wrench appropriate for SMA. Since the VNA is a 2-port network, the output ports had to be measured one at a time. Hence, this configuration measures how well a signal passes through the device from the input port to one of the output ports. The same process was repeated to measure the power going into the output port previously terminated.

2.3.5 Measured Results

Both amplitude and phase were measured and the results are shown in Figures 2.29, 2.30, and 2.31. Figure 2.29 shows similar $|S_{11}|$ results measured from the input ports and

demonstrates the fabricated feed network is matched from 1.68 to 2.4 GHz. The amplitude difference between ports 2 and 3 varied from a 0.46 dB to -0.9 dB as shown on Figure 2.30. The amplitudes deviated 1.36 dB, which is slightly worse, but quite close to the simulated amplitude deviation of 1.1 dB. Figure 2.31 shows a measured phase difference varying from -86° to -94° over 1.68 to 2.4 GHz compared to the simulated phase difference of 81° to 89° from 1.68 to 2.4 GHz. Both behaved similarly in a sinusoidal pattern. However, the measured phase difference is more contained within $\pm 5^\circ$ of 90° . After testing the feed network and verifying that it works, the next step was designing the antenna.

Figure 2.29 Measured input impedance matching results of the fabricated feed network

Figure 2.30 Measured amplitude difference of the fabricated feed network

Figure 2.31 Measured phase difference of the fabricated feed network

Chapter 3: Antenna Design

Like the feed network, several IEEE papers involving circularly polarized antennas were studied for performance and design and fabrication suitability. Many patch antennas demonstrated excellent AR and impedance bandwidth performance but involved components and layers that increased the design complexity such as the quadruple L-probe in [9] and [10]. The quadruple L probe fed circular patch antenna from [9] showed a simulated 2-dB AR bandwidth of 56% from 1.35 to 2.4 GHz with an impedance bandwidth of 94.1% from 0.9 to 2.5 GHz. The quadruple L-probe fed antenna from [10] showed a simulated 2-dB AR bandwidth of 48.8% from 1.33 to 2.18 GHz and impedance bandwidth of 73% from 1.07 to 2.3 GHz. In addition to a quadruple probe feed, the same patch antenna design in [10] was also simulated using dual L-probe feeds at which demonstrated a 3-dB AR bandwidth of 39% from 1.26 to 1.87 GHz and an impedance bandwidth of 59.52% from 1.18 to 2.18 GHz. Based on results from [10] and its design simplicity, the proximity dual probe fed antenna design idea was chosen for this project with a few modifications.

The antenna was designed to operate at a bandwidth of 1.7 to 2.13 GHz with a center frequency of 1.9 GHz. The antenna section is made up of the upper substrate, feed probes, and ground plane as shown on Figure 3.1. The circular patch antenna is 68.2 mm in diameter and sits on top of a 140 mm x 140 mm x 0.8 mm FR4 substrate with a relative permittivity of 4 and a loss tangent of 0.018. The bottom side of the substrate are two orthogonal 21.45 mm x 13 mm rectangular strips embedded underneath as shown on Figure 3.2. The ground plane parallels the upper substrate and has the same length and width dimensions. The ground plane and upper substrate are separated by 18 mm long cylindrical probes and air layer. The dual cylindrical probe feeds are 3.2639 mm in diameter and are

in accordance with the American Wire Gauge 8 (AWG 8) standards [18]. Unlike the feed network design, the antenna section was designed only using FEKO and did not have a prototype fabricated until integration with the feed network.

Figure 3.1 Side view of the antenna design.

Figure 3.2 Side view of the antenna design in 80% opacity display

3.1 T-Probe Feeds

The notable physical differences compared to [10] are the T-shaped probes and that its horizontal section is embedded on the bottom of the antenna substrate with only a small

part of it directly underneath the patch. The antenna uses the proximity fed method as opposed to coaxial, microstrip line, or aperture coupled. The circular patch is excited by two proximity-fed T-shaped probes underneath the antenna substrate. The dual T-probes are formed by two orthogonal rectangular strips parallel to the substrate and attached to cylindrical probes perpendicular to the substrate as shown in Figure 3.2. The rectangular strips act as a capacitance to compensate for the inductive impedance due to the probe's vertical length. Balancing the inductance and capacitance is essential in achieving impedance matching and wideband performance. Circular polarization was achieved by placing the probes orthogonal to each other and varying their dimensions and positions relevant to the circular patch. The 21.45 mm x 13 mm rectangular horizontal strips are 90° apart from each other and positioned 44.795 mm from the center of the patch as shown on Figure 3.3. The cylindrical probes are 18 mm long with a diameter of 3.2639 mm based on the AWG 8 standard wire system. The cylindrical probes are positioned about 1.325 mm off-centered with respect to the length of the horizontal strip and 43.47 mm from the center of the patch also shown on Figure 3.3. The top end of the cylindrical probes makes direct contact with the horizontal strips forming the T-probe shape. The bottom end connects to input ports configured with simulation voltage sources of 50 Ω impedance. One port was configured with 0° phase and the other with 90° phase to model an ideal feed network.

Figure 3.3 Top view of the antenna design in 40% opacity display to show the rectangular strips on the bottom of the substrate

3.2 Impedance Match and Axial Ratio

If the input impedance to the antenna is 50Ω , measured by the input port looking toward the antenna, the load is matched which means the transmitted power is delivered to the antenna and radiated with minimal reflection. An $|S_{11}|$ of less than -10 dB is a commonly used indication of this. Hence, the $|S_{11}|$ plot is the primary means of assessing impedance performance. Another plot that can aid in revealing how closely matched the antenna is to 50Ω is an impedance plot that displays both the real part and imaginary part. An imaginary part greater than zero ($X_L > 0$) indicates a positive (inductive) reactance. An imaginary component of less than zero ($X_C < 0$) indicates a negative (capacitive) reactance. The load impedance is composed of a resistance and reactance ($Z_L = R + jX$). The idea is to cancel out the reactance ($jX = 0$) while driving the real part as close to 50Ω as possible ($Z_L = R = 50$). The way to control the reactance is to adjust the parameters that affect the inductance and capacitive nature of the design. These parameters are the patch radius, substrate permittivity and thickness (including air layer in between), and the dimensions and

locations of the T-probes.

Impedance matching was achieved by varying the probe radius and the air dielectric height which simultaneously changes the length of the probe. However, the probe thickness and dielectric substrate design were limited to specific fabrication dimensions available. The dielectric substrate was based on the PCBway.com selection of substrate material and thickness while also considering costs. The cylindrical probe was based on the AWG standard size wires which varies in discrete diameter steps. The wire size also needed to be reasonably thick so that it would not be difficult to fabricate. The patch radius was varied to add or remove more capacitance as necessary.

All of the design parameters affect the AR and $|S_{11}|$ performance simultaneously, but some have a more pronounced effect on one than the other. The optimal AR was found by adjusting the width and length of the horizontal strips connected to the probe including moving the horizontal strips outwards away from the patch (see Figure 3.4). However, this also depended on the state of the other parameters. Another adjustment that had a large effect was moving the probe towards the center of the horizontal strip. The initial placements of the probe and horizontal strips were based on [10], but this did not produce desired results after running simulations. Hence, adjustments such as one made from an L-probe to a T-probe were necessary.

Figure 3.4 A zoomed-in top view of horizontal strip in 40% opacity display

Gain was also monitored but was not as sensitive to the design parameter variations as the AR and impedance bandwidth. The AR and gain were simulated with two orthogonal far field cuts with the following settings: one with $-90^\circ < \theta < 90^\circ$ for $\phi = 0^\circ$ and another with $-90^\circ < \theta < 90^\circ$ for $\phi = 90^\circ$. Phi, ϕ is the azimuth angle measured from the x -axis. Theta, θ is the elevation angle measured from the z -axis where 0° is the center and perpendicular to the patch, which is where the maximum radiation occurs. As the radiation pattern rotates 90° towards the edge of the substrate, radiation is at its minimum. Radiation is not intended to be behind the patch or in the $-z$ -direction, so θ was set to only an 180° span.

3.3 Simulation Approach

During the antenna design process, several simulation techniques were employed while using FEKO. One was running the simulation in discrete frequencies and using an infinite multilayer plane/ground to reduce simulation time. Another key feature to save simulation time in FEKO was the parameter sweep function. The parameter sweep function performs multiple simulation runs for a combination of parameters requested to sweep as opposed to a single simulation run. The feed probes were constructed using cylinders for

accurate meshing. Local meshing, which applies to a specific section or geometry faces, was used to smooth out the sharp edges on the probes from the meshing to retain its cylindrical form. Otherwise, the simulation may not be accurate for a cylindrical wire and could impact the results. Local meshing was also applied on the feed network trace during its design. However, too fine of a mesh can lead to an exceedingly long simulation time, so a balance between a fine mesh and simulation time had to be used.

The total bandwidth of the circular patch antenna depended on both the impedance bandwidth and the AR bandwidth. Either one of these performance parameters can limit the operating bandwidth of the antenna. Often during simulations, as the AR bandwidth performance enhanced, the impedance bandwidth degraded and vice versa. Hence, the antenna design stage required extensive optimization to find a balance between the impedance and the AR. This was done by sacrificing the performance in one area to improve the performance of another, so that both can meet the required bandwidth. Once the desired bandwidth performance was attained, the infinite plane layer was replaced with a finite ground plane to be more representative of the fabricated microstrip antenna. Adding a finite ground plane increases the simulation time because of the additional mesh elements, but results were similar to the infinite layer results. A few final optimizations returned the results back to the desired performance.

3.4 Simulation Results

Figure 3.5 shows the $|S_{11}|$ from 1.68 to 2.2 GHz. As can be seen from this figure, $|S_{11}| < -10$ dB from 1.75 to 2.15 GHz indicates matched performance at that particular band. Figure 3.6 also shows the imaginary part of the input impedance falling off significantly from 0.5Ω at 2.1 GHz to -46.2Ω at 2.2 GHz. In a 100 MHz span, the impedance changed

by $45.7 \, \Omega$. A correlation can be observed between the $|S_{11}|$ plot and the impedance plot. Based on the two plots from 2.13 GHz to 2.2 GHz, the input impedance became highly capacitive as indicated by the green imaginary line significantly dropping down to $-46.2 \, \Omega$ and that consequently drove the $|S_{11}|$ above -10 dB. As a result, the $|S_{11}|$ limited the overall operating bandwidth of the antenna. Figure 3.7 shows the gain and axial ratio performance. The total gain peaked at 8.04 dBi at 1.7 GHz and then gradually reduced to 4.73 dBi at 2.2 GHz. The AR was 1.2 dB at 1.68 GHz and improved to 1 dB at 1.7 GHz to 2.2 GHz. Figure 3.8 shows the total gain pattern at 1.9 GHz with ϕ at 0° . Peak gain of 8.1 dBi can be seen slightly to the right at about 8° . Figure 3.9 shows the LHC gain pattern antenna and its RHC gain cross polarization pattern at 1.9 GHz with $\phi = 0^\circ$. The higher gain performance shown with LHC in contrast to the RHC pattern clearly indicates that the designed patch antenna is LHC polarized. Based on the impedance, AR, and gain performances presented, the simulated antenna design has achieved the desired bandwidth requirements.

Figure 3.5 FEKO simulated antenna input impedance match

Figure 3.6 Real and imaginary parts of the simulated antenna input impedance

Figure 3.7 Simulated total gain (dBi) and axial ratio (dB) of the antenna

Figure 3.8 3D view (left) and polar plot (right) of the total gain (dBi) at 1.9 GHz

Figure 3.9 LHC gain (blue) and RHC gain (green) radiation pattern at 1.9 GHz

Chapter 4: Feed Network and Antenna Integration

The final stage in the design process was integrating the feed network and antenna from the outputs of the phase shifter to the bottom of the cylindrical probes of the antenna section. The ground plane platform of the antenna section was replaced with the phase shifter which has a ground plane on the top side of the substrate. The ground side up configuration was to isolate the feed network from the antenna radiation. Several other design modifications were made during integration. The diameter of the patch was decreased to 64.4 mm and the probe length were decreased to 17.5 mm to optimize the AR performance. The probes were directly connected to the output ports of the phase shifter which equated to an air layer of 16.7 mm. Since the ground plane was on the top side of the substrate and the phase shifter was on the bottom, holes in the lower substrate had to be created so that the probes can contact the output ports of the phase shifter. However, any probe and ground plane contact would generate an unintended electrical path to the ground and ultimately shorts the circuit. For this reason, the ground plane was modified where a circular area of substrate was etched around the hole to allow space between the probe and the ground plane as shown on Figure 4.1.

Another distinct change was the phase shifter's reference line layout with respect to the Wilkinson's output port and the cylindrical probe as shown in Figures 4.2, 4.3, and 4.4. Moving the probes can change the AR and $|S_{11}|$ performance established in the antenna design stage, so the probes were not changed. The main priority was aligning the phase shifter outputs to the probes without moving the probes. This was done by rearranging multiple sections of the reference line while maintaining the same width and total length. The length can be changed but this would also have to be done to the main line. Other

considerations for reconfiguring the reference line were the size of the substrate and its position relative to the Wilkinson power divider's outputs. The 103 mm long reference line (if extended) connected to the Wilkinson can increase the size of the lower substrate. Furthermore, the Wilkinson would have to be angled because the reference line and main line have different lengths making the design more difficult.

Following the phase shifter, the Wilkinson power divider had minor design changes. The length of the 100 Ω lines were adjusted based on the distance apart between the main line and reference line. The input port length of the Wilkinson shown on Figures 4.2, 4.3, and 4.5 was extended to meet the desired substrate dimensions. The feed network substrate and the antenna substrate individually have the same thickness but different lengths and widths. Initially, the lower substrate's length and width was changed to reflect the upper substrate, but the gain dropped dramatically. As a result, the size of the ground plane was increased thereby improving the antenna gain. The integrated feed network and antenna consisted of a 170 mm x 170 mm x 0.8 mm lower substrate and a 140 mm x 140 mm x 0.8 mm upper substrate as shown on Figures 4.6 and 4.7.

Figure 4.1 A zoomed-in view of the no-copper area on the ground plane for the integrated antenna

Figure 4.2 Top view of the integrated antenna design in 40% opacity display

Figure 4.3 Bottom view of the integrated antenna design

Figure 4.4 A zoomed-in view of the phase shifter in 60% opacity display

Figure 4.5 A zoomed-in view of the Wilkinson power divider

Figure 4.6 Top view of the patch antenna design

Figure 4.7 Side view (tilted) of the integrated antenna design

4.1 Simulation Results

Figure 4.8 shows the $|S_{11}|$ below -10 dB from 1.68 to 2.4 GHz. Figure 4.9 shows the gain and axial ratio performance at which the total gain peaked at about 7.74 dBi at 1.9 GHz and then gradually reduced below 7 dBi from 2.11 GHz to 4.1 dBi at 2.4 GHz. The 2-dB AR was from 1.7 to 2.06 GHz. The AR increased above 3 dB at 2.11 GHz and then back down below 4 dB at 2.22 GHz. Figure 4.10 shows the total gain pattern at 1.9 GHz with ϕ at 0° . Figure 4.11 shows the LHC gain pattern and its RHC gain cross polarization pattern at 1.9 GHz with $\phi = 0^\circ$.

Figure 4.8 FEKO input impedance matching results of the integrated antenna design

Figure 4.9 Total gain (dBi) and axial ratio (dB) of the integrated antenna design at $\theta = 0$ and $\phi = 90^\circ$ and 0°

Figure 4.10 3D view (left) and polar plot (right) of the total gain (dBi) at 1.9 GHz

Figure 4.11 LHC gain (blue) and RHC gain (green) radiation pattern at 1.9 GHz

4.2 Fabrication

The fabrication process for the PCBs of the integrated design was similar to Chapter 3. The Gerber files were created and exported from ADS Layout and the PCB prototype was ordered on PCBway.com. The difference being two circuit boards instead of one: the upper substrate and lower substrate. The Gerber files formatting for the upper substrate was similar to the feed network in Chapter 3. It consisted of a substrate outline, top layer (circular patch shown on Figure 4.12), and the bottom layer file (rectangular strips shown on Figure 4.13). The lower substrate on the other hand had a drill file in addition to the same three files as the upper substrate. The drill file contained two filled circles imported as shown on Figure 4.14. Since the drill holes (based on the feed probe's diameter) were larger than the outputs of the phase shifter, circular pads were added at the output ends in

order to solder the feed probes onto the outputs of the phase shifter as shown on Figure 4.15. The top layer which contains the ground plane was also modified to include the no-copper areas to prevent any ground plane contact with the feed probe as shown on Figure 4.16. The no-copper areas indicated by two hollow squares within a filled rectangle encompass the drill holes. Figures 4.17 and 4.18 shows the fabricated upper and lower substrate PCBs, respectively.

Figure 4.12 Circular patch antenna copper trace layout for PCB fabrication

Figure 4.13 Rectangular probe strips copper trace layout for PCB fabrication

Figure 4.14 Drill holes layout for PCB fabrication

Figure 4.15 Feed network copper trace layout for PCB fabrication

Figure 4.16 Ground plane layout for PCB fabrication

Figure 4.17 PCB prototype of the top (left) and bottom (right) of the upper substrate (antenna)

Figure 4.18 PCB prototype of the top side (left) and bottom side (right) of the lower substrate (feed network)

Four parts were soldered onto the integrated PCB: one female SMA connector, one 0603 SMD, and two probe feeds. The 0603 SMD was soldered first following the feed

probes. An AWG 8 solid copper wire was used as the feed probe. About a 1.8 to 1.9 cm long piece of wire was cut from a two feet long AWG 8 wire. The jagged ends of the wire were filed as flat as possible to replicate the flat end of the simulated feed probe. The wire was measured to be as close to 17.5 mm long as possible and then inserted through the drill hole in the ground plane, flushed with the circular padding on the feed network output ports. Electrical tape was used to hold the probe in place while being soldered to the feed network output ports. Since the rectangular strips did not have the probe soldering area marked, measurements were taken carefully and marked by hand. With both probes soldered to the phase shifter, the other ends of the probe were aligned and soldered on the marked soldered location on the horizontal strips. The last component soldered was the SMA connector which serves as the input port for the integrated PCB. The soldered integrated circularly polarized antenna is shown in Figures 4.19, 4.20, and 4.21.

Figure 4.19 Side view of the soldered integrated circularly polarized antenna

Figure 4.20 Bottom view of the soldered integrated circularly polarized antenna showing the feed network

Figure 4.21 Top view of the soldered integrated circularly polarized antenna showing the antenna, horizontal strips, and ground plane below

4.3 Measured Results

The $|S_{11}|$ was measured using the VNA at CSUN. The $|S_{11}|$ VNA result was plotted in MATLAB as shown on Figure 4.22 and demonstrates that the integrated PCB is matched from 1.6 to 3.0 GHz. The simulated input impedance is also matched from 1.6 to 3.0 GHz. The measured results looks similar to the simulated results, but the measured resonances seems to be shifted slightly to the right of the simulated resonances. The AR and gain were measured in the anechoic facility in CSUN's Microwave and Antenna's Laboratory. The receiving patch antenna was rotated 360 degrees while pointing at the transmitting linearly polarized horn antenna. If the patch antenna is circularly polarized, the received signal amplitude should not vary regardless of either antenna's orientation. The AR was measured by taking the difference of the maximum and minimum variation one frequency at a time. The results were recorded and then plotted on MATLAB as shown on Figure 4.23.

Figure 4.22 Measured and simulated input impedance matching results

Figure 4.23 Measured and simulated AR

The AR results shows a 3-dB bandwidth from 1.66 to 2.44 GHz and a 2-dB bandwidth from 1.8 GHz to 2.07 GHz. The best AR results was 1.38 dB at 2.0 GHz. The AR performance behaved somewhat similar to the simulation where there were subtle dips at 2.0 GHz and 2.35 GHz before slightly climbing towards 2.4 GHz. However, for the most part, the AR remained steady between 1.8 dB to 2.4 dB whereas the simulation exhibited a more pronounced sinusoidal pattern. Although the AR bandwidth results were better than simulations, performance could still be improved. There were many variables that affect the AR performance. One involves fabrication by hand which can certainly be inaccurate. The soldering, probe placements, probe length and air gap measurements were all done manually. As mentioned in the design process, for parameters such as the probe position relative to the horizontal strip and the air gap between the two substrates, accuracy is imperative to AR performance.

Another critical factor was the accurate alignment of the measurement system. It was discovered in the lab that the transmitting antenna was not pointed directly at the receiving patch antenna. In simulation, a 5° offset can cause a 0.5 to 1 dB degradation in AR performance. Since the AR simulated result was based on θ centered at 0° with $\varphi = 0^\circ$ and 90° , manual adjustments were made to both receiving patch antenna and transmitting horn antenna to ensure the main beam was pointed at the patch antenna. Therefore, measured AR performance is highly dependent on the accuracy of the manual fabrication process as well as the lab equipment available. Antenna gain was measured by taking the S_{21} magnitude and phase of the patch antenna oriented 90° vertical and 0° horizontal as the turntable rotated from -90° to 90° in azimuth. The S_{21} magnitude of the standard gain horn antenna was also measured to provide a reference for shifting the patch antenna gain appropriately. The magnitude and phase collected were imported to MATLAB and used to generate the measured antenna LHC gain shown on Figure 4.24. Results show the measured LHC gain is slightly less and narrower than the simulated LHC gain. The measured gain peaked at 6.8 dBi and the simulated gain peaked at 7.7 dBi.

Figure 4.24 Measured and simulated LHC gain including the RHC cross polarization at 2.0 GHz, $-90^\circ < \theta < 90^\circ$, and $\varphi = 0^\circ$

Chapter 5: Conclusion

A proximity-fed circularly polarized microstrip antenna was designed, simulated, fabricated, and measured. The 64.4 mm diameter circular patch was printed on top of a 140 mm x 140 mm x 0.8 mm upper layer FR4 substrate and fed by a SIOS coupled line phase shifter on a 170 mm x 170 mm x 0.8 mm lower layer FR4 substrate. Simulation results show a 3-dB AR bandwidth of 22.2% from 1.68 GHz to 2.1 GHz, a 2-dB AR bandwidth of 19.3% from 1.69 GHz to 2.05 GHz, and a 10-dB return loss bandwidth of 60.9% from 1.6 GHz to 3.0 GHz. Measured results show a wideband 3-dB AR of 38% from 1.66 to 2.44 GHz, a 2-dB AR bandwidth of 14% from 1.8 to 2.07 GHz, and the same 10-dB return loss bandwidth of 60.9% from 1.6 GHz to 3.0 GHz. In simulation, the AR performance is highly sensitive to where the probe is positioned relative to the patch and the horizontal strip including the length of the probe and the air layer separation. The accuracy in lab measurement equipment can also have a huge impact in performance. Therefore, if the entire fabrication was done precisely and with accurate measurement equipment, performance could be further improved. In addition to the wideband impedance matching and 3-dB AR bandwidth performance, the presented antenna exhibited a slightly wider 2-dB AR bandwidth than the similar dual L-probe feed patch antenna design in [10] which has a 2-dB bandwidth of about 11%. Moreover, it's worth mentioning that this design was also built with a lossy FR4 substrate and is much less complex than other wideband circularly polarized designs. For these reasons, the design overall was a success and could be used as a foundation to build upon for space satellite antenna applications.

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