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Design and Analysis of Receiver Systems in Satellite Communications and UAV Navigation Radar

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Design and Analysis of Receiver Systems in Satellite Communications and UAV Navigation

Radar

Matthew R. Morin

A thesis submitted to the faculty of
Brigham Young University
in partial fulfillment of the requirements for the degree of
Master of Science

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ABSTRACT

Design and Analysis of Receiver Systems in Satellite Communications and UAV Navigation Radar

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Master of Science

The design of a low cost electronically steered array feed (ESAF) is implemented and tested. The ESAF demonstrated satellite tracking capabilities over four degrees. The system was compared to a commercial low-noise block downconverter (LNBF) and was able to receive the signal over a wider angle than the commercial system. Its signal-to-noise ratio (SNR) performance was poor, but a proof of concept for a low cost ESAF used for tracking is demonstrated.

Two compact low profile dual circularly polarized (CP) reflector feed antenna designs are also analyzed. One of the designs is a passive antenna dipole array over an electromagnetic band gap (EBG) surface. It demonstrated high isolation between ports for orthogonal polarizations while also achieving quality dual CP performance. Simulations and measurements are shown for this antenna. The other antenna was a microstrip cross antenna. This antenna demonstrated high gain and quality CP but had a large side lobe and low isolation between ports.

A global positioning system (GPS) denied multiple input multiple output (MIMO) radar for unmanned aerial vehicles (UAVs) is simulated and tested in a physical optics scattering model. This model is developed and tested by comparing simulated and analytical results. The radar uses channel matrices generated from the MIMO antenna system. The channel matrices are then used to generate correlation matrices. A matrix distance between actively received correlation matrices to stored correlation matrices is used to estimate the position of the UAV. Simulations demonstrate the ability of the radar algorithm to determine its position when flying along a previously mapped path.

Keywords: GPS denied radar, physical optics scattering modeling, low profile antennas, dual circularly polarized antennas, electronic beam steered antennas, satellite communications
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CHAPTER 1. INTRODUCTION

The first active communications satellite was launched in 1962 [2]. Since then the satellite communications (SatCom) industry has become ubiquitous for modern society. SatCom is now being used for various applications such as television, telecommunications, navigation systems, and the internet.

As mentioned, one of the uses of SatCom is for navigation. The Global Positioning System (GPS) was the first globally operational satellite navigation system and was implemented in 1995. GPS has enabled accurate positioning capabilities to millions of military and commercial users worldwide. As with any technology there are limitations. This necessitates other means of navigation when dependence on GPS is not possible.

This thesis will discuss contributions made for earth based satellite communication receivers as well as a simulated performance analysis of a proposed GPS denied navigation radar system for UAVs.

1.1 Earth Based Satellite Communication Electronic Beam Steering Feeds and Dual Circularly Polarized Low Profile Antennas

For earth based stations in SatCom systems, parabolic reflector antennas are frequently used to receive or transmit signals. These parabolic reflector antennas have a feed antenna that illuminates the reflector which directs the signal toward the satellite.

Currently feed antennas are most commonly horn antennas and generally have high radiation, aperture, and spill-over efficiencies. However these antennas are large, heavy, and expensive to manufacture. Recently BYUs SatCom research group has demonstrated various low profile, light, and inexpensively manufacturable antennas that perform with similar efficiencies as horn antennas [3], [4].
One of the challenges in designing low profile antennas is achieving quality dual circular polarization (CP). There have been many satisfactory low profile single CP antennas, but adding the second port to achieve the orthogonal polarization while maintaining sufficiently high efficiencies and isolation has been problematic. Designs to address this problem will be presented and discussed in Chapter 4.

Due to non ideal movement of a satellite or an earth based antenna in a SatCom system, the signal can be decreased or lost. For example wind may cause the reflector to shake, temporarily causing the signal to be lost until the wind dies down. Or a satellites orbit can decay to the degree that it no longer receives a constant signal, thereby causing it to be abandoned. These issues can be remedied by using electronic beam steering technologies to track the signal. This could extend the life of satellites, allow for tracking in windy conditions, and reduce dish alignment costs. An electronically steered beam can be formed by shaping the primary beam with an array feed. An array feed design that adds this functionality to an earth based receiver is discussed in Chapter 3.

1.2 Performance Analysis of a GPS Denied Radar using a Physical Optics Model

In recent years there has been a dramatic increase in the use of unmanned aerial vehicles (UAVs). Their utility is demonstrated by the diversity of applications that they are used for, such as military reconnaissance and attacks, scientific research, and wild fire management. Though the most commonly known uses tend to be military, it is predicted that their commercial applications will increase dramatically [5]. With this predicted expansion, as well as the demands for military purposes, there are many technologies that need to be developed. One of these is positioning methods that can be used when the global positioning system (GPS) is unavailable.

UAVs typically use GPS to track their location. While in general GPS provides an accurate and dependable service, there are times when it is ineffective. To achieve accurate location estimation the UAV needs to be in the line of sight of four GPS satellites and not in a strong multipath area [6]. If the UAV is in a strong multipath area the GPS signals can be delayed, causing the position calculations, which depend on the timing of arrival of the signal, to be incorrect. If a UAV is flying through a natural or urban canyon, the UAV likely will not have a line of sight to four GPS satellites, either rendering the UAV partially or completely unable to identify its location.
In addition to errors in position estimation that may occur due to the UAVs immediate environment, there may be hostile means such as jamming or spoofing which can render the GPS system on the UAV incapable of fulfilling its purpose. GPS spoofing is where a false GPS signal is created to fool the UAV into thinking that it is at a different position than it really is, allowing the spoofer to direct the UAV [7].

To avoid these potential hazards that come from dependence on GPS, a position recognition radar system and corresponding algorithms are being developed. As the UAV flies it will actively transmit and receive the backscattered signal. This signal will be a function of the propagation environment, which would include various terrain features. If the multipath environment is rich enough, the backscattered signal will vary with position in temporal and spatial characteristics. The signals would then correspond to locations of the UAV when the signals were received, and therefore allow position tracking.

To give insight into the performance of the radar and corresponding algorithm a physical optics scattering model is developed. This model will simulate antennas generating electric fields which reflect off of a randomly generated terrain, and return to the receive antennas. This is done at various positions where the electric fields at the receive antennas are saved. This generated data is then used to analyze the performance of the algorithm. The physical optics scattering model and the analysis of the GPS denied radar is discussed in Chapter 5.

1.3 Contributions

The following is a list of contributions discussed in this thesis:

- A design of an electronically steered array feed (ESAF) for a dish is built and tested. It demonstrated signal tracking capabilities for a SatCom application. This is discussed in Chapter 3.

- An antenna design is proposed that demonstrated that a low profile native dual CP antenna can achieve quality CP, high radiation efficiency, and high isolation between ports. This will be discussed in Chapter 4.

- An analysis of a MIMO radar algorithm using a physical optics scattering model is discussed. It is demonstrated that this algorithm can track the position of the UAV in the physical optics
model, proving that valuable position information can be obtained using the MIMO radar algorithm. This is shown in Chapter 5.
CHAPTER 2. BACKGROUND

2.1 Antenna Parameters

Since there are many different types of antennas and antenna properties, there are various parameters to describe and measure their performance. The following will give a brief introduction to the antenna parameters that are used in this thesis.

2.1.1 Scattering Parameters

Since the antennas discussed in this thesis have two ports there are four S-parameters. The S-parameters $S_{11}$ and $S_{22}$ indicate how much power is rejected when fed into the respective antenna ports. The power accepted into the antenna either is radiated, absorbed into the antenna materials, or coupled into the other port. The parameters $S_{12}$ and $S_{21}$ are a measure of the latter. Since the antennas are made of isotropic materials these two parameters should be identical. Clearly we want all the power to be accepted by the antenna and radiated, therefore we want to minimize all of these values over the working bandwidth.

2.1.2 Polarization

In practice all waves are elliptically polarized to some degree but can approach either linear or circular polarization (CP). The polarization quality can be described by the axial ratio (AR), which is defined as

$$AR = \frac{|E_{\text{major}}|}{|E_{\text{minor}}|},$$  \hspace{1cm} (2.1)

where $E_{\text{major}}$ and $E_{\text{minor}}$ are the major and minor axes of the polarization ellipse [8]. Ideally, for linear polarization AR is infinite, but AR is not generally used when discussing linear polarization.
For an ideal CP antenna the magnitude of $E_{\text{major}}$ and $E_{\text{minor}}$ are the same and so AR should be exactly one. The AR of an antenna varies with the elevation and azimuthal angles. A CP antenna will generally be required to have AR below a ceiling value for a certain beam width centered on the main beam. For reflector antennas AR is typically used to characterize CP of an antenna feed, or the primary pattern.

Another parameter used for determining the quality of the polarization of an antenna is the cross polarization isolation (XPI). This is defined as the power received in the co-polarization of an antenna given an antenna transmitting in the co-polarization, divided by the power received in the cross polarization of an antenna given a transmitting in the cross polarization with the same gain [9]. The co-polarization is the polarization that the antenna, or a specific port on an antenna, is designed to radiate in [8], while the cross-polarization is the field component that is orthogonal to the co-polarization. Generally XPI is used to characterize CP for the secondary pattern of an antenna feed and reflector.

Orthogonally polarized waves can propagate in the same path and not interfere with each other. By exploiting this property, different signals can be sent on orthogonal polarizations, thereby doubling the amount of data sent with just one polarization. The challenge is in achieving high isolation between ports of an antenna that receives or generates the orthogonal polarizations.

The antenna designs presented in this thesis will all be dual CP. CP has a couple of advantages over linear polarization. First, aligning antennas is easier. As long as antennas are pointing in the right direction the polarizations are aligned, whereas with linear polarization they have to be oriented properly as well. Furthermore, when linear polarized waves propagate through the ionosphere the polarization can be rotated, causing the magnitude of received signal to decrease. This is known as Faraday rotation [10, p. 130]. Faraday rotation adds a phase shift to a CP signal but it still maintains the same polarization, which means that for a CP wave signal is not lost. This is not a serious issue when dealing with frequencies above 10 GHz [10, p. 130].
2.1.3 Antenna Efficiencies

Various antenna efficiencies help with the management of power, which is essential in a SatCom system. Radiation efficiency, $\eta_{\text{rad}}$, is [8],

$$\eta_{\text{rad}} = \frac{P_{\text{rad}}}{P_{\text{in}}},$$

(2.2)

where $P_{\text{in}}$ and $P_{\text{rad}}$ are respectively the accepted and radiated powers of the antenna. Radiation efficiency can only be adversely affected by loss in the materials, thus having designs and materials that are low loss is essential to achieving high radiation efficiency.

Antenna efficiency is

$$\eta_{\text{ant}} = \frac{A_e}{A_p},$$

(2.3)

where $A_e$ is the effective area of the antenna and $A_p$ is the physical area of the antenna. Antenna efficiency can be related to the aperture efficiency, $\eta_{\text{ap}}$, by

$$\eta_{\text{ant}} = \eta_{\text{ap}} \eta_{\text{rad}}.$$

(2.4)

The aperture efficiency of an aperture antenna is a measure of how well the aperture is illuminated. For an aperture antenna the maximum possible aperture efficiency is one.

Spillover efficiency, $\eta_{\text{spill}}$, is the ratio of power that is incident on the dish to the total power that is radiated by the feed antenna [8]. The aperture and spillover efficiencies, measure the degree to which the antenna feed’s beam pattern focuses the radiated energy in the correct direction.

2.1.4 Gain and Beam Pattern

The directivity of an antenna is the ratio of the radiated power in one direction to the power radiated in the same direction by an isotropic antenna. The gain of an antenna is radiation efficiency multiplied by the directivity [11, p. 414]. The gain changes over different directions of propagation, therefore mapping the gain as a function of direction gives a sense of where the
transmitted power is going, and is also known as the beam pattern. In this thesis a beam pattern will be typically shown by two orthogonal cuts of the beam pattern.

2.2 Types of Antennas

Various types of antennas are discussed in this thesis. Some are used in the designs presented in Chapter 4. Others are antennas that are used commonly for the different applications discussed. Below is an overview of these different types of antennas.

2.2.1 Rectangular Patch Antennas

The rectangular patch antenna is the most common type of microstrip antenna [12, p. 816]. It consists of a rectangular patch over a ground plane. There are various ways to feed this antenna, the most common of which is the microstrip-line feed. The dimensions of the patch antenna depend on the substrate thickness and its relative permittivity. In the direction of the linear polarization the antenna needs to be approximately half a wavelength with respect to the relative permittivity, where air is above and the substrate is below.

The patch antenna has a broad radiation pattern with low side lobes, with the main beam at boresight. The gain at boresight is around 6-7 dB. Due to the symmetric quality of the patch antenna it is a good candidate for an element in arrays. The bandwidth for patch antennas is limited but can be expanded by having thicker substrates. This however can cause more loss, due to the added substrate, therefore decreasing the radiation efficiency which is already an issue for patch antennas.

A major benefit of the patch antenna, and generally of microstrip antennas, is the ease and inexpensive cost of manufacturing. Because it can be made on a printed circuit board (PCB) it can also be easily integrated with other circuitry. It is also light and small. These characteristics make it a good candidate for using as a reflector feed antenna.
2.2.2 Horn Antennas

The horn antenna is commonly used for SatCom reflector antenna feeds. Horn antennas are open ended pipes with a flair at the open end. They are usually fed by a probe which inserts into a waveguide which is attached to the horn.

Horn antennas are used for several reasons including their high radiation, aperture and spillover efficiencies. For example, they are reported to have aperture efficiencies up to 75-80% [12]. This means that the beam patterns are well shaped. Wide bandwidths are also achievable by horn antennas.

Horn antennas can transmit and receive dual linear polarization, and by using a polarizer they can be dual CP. When there are stricter requirements for isolation between ports often an orthomode transducer (OMT) is used. These can be quite large and complicated structures compared to the horns themselves. Horn antennas can be heavy, bulky, and expensive, especially if an OMT or polarizer is needed.

2.2.3 Dipole Antennas

Dipole antennas consist of two rods that are aligned end to end with a gap between them. They are linearly polarized antennas and have omnidirectional beam patterns. Nulls form in the directions that the rods point, and max gain occurs in the direction perpendicular to the rods. Dipoles on their own do not have highly directive beam patterns and so are not suitable for a feed in a dish configuration. By placing the dipole a quarter wave above a metal plane, the reflected and transmitted waves add constructively, and a more directive pattern is created.

The feed to a dipole needs to be balanced. If the transmission line that is delivering the signal is unbalanced, a balun is used to create a balanced signal that is suitable for feeding the dipole. For example, a coaxial cable can be used to feed a dipole where one rod is connected to the center pin and the other is connected to the shielding. If no balun is used, some of the current on the rod connected to the outer conductor will propagate down the outside of the coaxial cable, causing the current on the two rods to not be balanced. This causes the pattern to be distorted. Adding a balun causes the impedance down the outside of the outer conductor to become large, which restricts current flow. Thus current is pulled mainly from the rod and thereby balances the
current on the rods allowing the dipole to radiate properly. A more complete explanation as well as various balun designs can be found in [12, p. 538-541].

2.3 SatCom System Overview

A SatCom system consists of both earth based stations and satellite stations. We will only be proposing designs that pertain to earth based systems. To understand the motivation of these designs it is necessary to understand the properties of the SatCom system as a whole.

There are various orbits that satellites can take but we will be only discussing the geostationary orbits. Geostationary orbits, as their name implies, are orbits where the satellite stays in the same position relative to the earth. Satellites in this orbit therefore lie in the equatorial plane, and to stay at the same position relative to the earth have a geostationary height of around 36,000 km [10, p. 77]. These orbits are used because an earth based antenna needs only to point in one direction at all times.

Due to the ellipsoidal shape of the earth, and also the gravitational pull of the moon and the sun, satellites in orbit can drift out of their desired positions and the orientation of the satellites can change. Different stabilization methods are used on the satellites to remedy this [10, p. 199,209]. Nevertheless satellites still can have non ideal movement. This can cause problems for earth based stations that do not have either a wide enough beam or active steering capabilities.

The earth based systems are used for uplink and downlink. They generally consist of various components including reflectors, feed antennas, power amplifiers, filters, and frequency mixers. For many of these systems the feed antenna is a horn and is paired with a reflector.

The antenna systems in this thesis were designed for a parabolic reflector with an offset feed. The parameter $f/D$ is used to characterize dishes, where $f$ is the focal length of the dish and $D$ is the diameter. It defines the opening angle of the dish and how wide the ideal gain pattern of the feed antenna should be. The ideal illumination pattern for the reflector is for all power transmitted to hit the dish and to be evenly distributed across the area of the dish, which forms a rect like pattern. Aperture and spillover efficiencies measure how well the dish is illuminated.

The spillover efficiency means different things for a transmitter than a receiver. For a transmitter it is a loss of power that is transmitted. As a receiver poor spillover efficiency means
that the signal received has more noise from signals coming from around the dish edges, thus decreasing SNR.

### 2.3.1 Sensitivity

The sensitivity is a measure of the quality of a receiver, and is related to the signal-to-noise ratio (SNR). SNR is defined as the power received from the signal divided by the power received from the noise. SNR can be improved by either increasing the strength of the received signal, which requires more power, or by decreasing the noise. Therefore, minimizing noise in a system is a problem with a large focus in industry, because the lower the noise power means the lower the power that needs to be transmitted.

Noise comes from many different sources. Noise can be generated in the receiver from different components such as antennas, amplifiers, and filters. It can also be received from sources outside the system. An analogy is made between noise generated by a resistor to noise in general. A resistor’s noise power increases with an increase in temperature. Different sources of noise are referred to as having an equivalent temperature, which is related to how much noise power they generate. Some contributing sources of temperature in a system and their corresponding symbols are receiver temperature, $T_{rec}$, external noise, $T_a$, and the physical temperature of the antenna $T_p$. These all contribute to the overall system temperature [13, p. 645]

$$T_{sys} = \eta_{rad} T_a + (1 - \eta_{rad}) T_p + T_{rec}. \quad (2.5)$$

For a reflector the external noise can be expressed as

$$T_a = (1 - \eta_{spill}) T_{ground} + T_{sky}, \quad (2.6)$$

where $T_{ground}$ is the temperature of the ground and $T_{sky}$ is the temperature of the sky. As can be seen in these two equations the antenna efficiencies impact the system temperature dramatically.

Given the importance of SNR for receivers, understanding how the antenna effects it is essential. This brings us to a key parameter for the performance of a receiver called the receiver sensitivity. It is the gain of the antenna divided by the system temperature, $G/T_{sys}$. For receivers
it is also known as G over T, or the figure of merit [8]. The latter name gives an idea of the significance of this parameter. Sensitivity is proportional to the SNR, and is purely an attribute of the receiver.

2.3.2 Noise Figure

Recall that equivalent temperature is a measure of how much noise a component, or some other source, contributes. An equivalent representative of the measure of noise of a component is the noise figure. It is defined as

\[ F = \frac{\text{SNR}_i}{\text{SNR}_o}, \]  

(2.7)

where \( \text{SNR}_i \) is the SNR of the input signal, and \( \text{SNR}_o \) is the SNR of the signal at the output of the device [13, p. 493]. The noise figure is greater than one because all real devices add noise.

The noise figure is directly related to the equivalent temperature, by

\[ F = 1 + \frac{T_e}{T_0}, \]  

(2.8)

where \( T_e \) is the equivalent temperature of the device. The noise figure assumes that the input noise is from a matched resister with a temperature of \( T_0 = 290 \text{ K} \) [13, p. 494].

A significant result for noise figure is the noise figure of a cascaded system. This can be written generally as,

\[ F_{\text{cas}} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots, \]  

(2.9)

where \( F_k \) and \( G_k \) are the noise figure and the gain, respectively, of the \( k \)th device in the cascade. From this equation we see that the noise figure and gain of the first device has the most impact on the overall noise figure, thus it motivates the design of the initial device to have a low noise figure and decent gain.
2.3.3 Link Budget Analysis of a SatCom System

It has been mentioned that the efficient management of power in a SatCom system is essential to making it cost effective. In order to understand the amount of power that is available at a receiver, a link budget analysis is needed.

There are various factors that affect the link budget. First, the equivalent isotropic radiated power (EIRP), where the square brackets imply that the values are in dBW, is

\[
[EIRP] = [G_t] + [P],
\]

(2.10)

where \( G_t \) and \( P \) are respectively the gain and power transmitted of the transmit antenna [8]. As the name suggests, it is the power density that an isotropic antenna would radiate with an input power of \( G_t P \). Next we have the free space loss (FSL) of the antenna, in dBW

\[
[FSL] = 10\log_{10}\left(\frac{4\pi r}{\lambda}\right)^2,
\]

(2.11)

where \( r \) is the distance from the source and \( \lambda \) is the wavelength [10, p. 353]. The loss increases with the square of the distance, and for smaller wavelengths. There are other sources of loss such as alignment losses, atmospheric absorption losses, but the FSL is the major source of loss in a SatCom system, and so we will disregard the other losses. Also, we have the gain of the receive antenna, \( G_R \).

The link budget can be represented as follows in dBW,

\[
[P_R] = [EIRP] + [G_R] - [FSL],
\]

(2.12)

where \( P_R \) is the power at the output of the receive antenna [10, p. 353]. The power from the signal must be large enough so that the required SNR can be achieved.

2.4 Multiple Input Multiple Output

Multiple input multiple output (MIMO) techniques enable the use of multiple signal paths, increasing the amount of information that can be sent wirelessly on the same frequency band [14].
For a MIMO system we can have $M$ transmit and $N$ receive antennas. We can represent the system as

$$x = Hs + w,$$

(2.13)

where $x$ represents the output voltages of the receive antennas in an $N \times 1$ column vector, $s$ is an $M \times 1$ column vector of the transmitted voltages at the input terminals, $H$ is called the channel matrix and is $N \times M$, and $w$ is the noise received in an $N \times 1$ column vector. The channel matrix is determined by the signal path and properties of both the receive and transmit systems. The properties of this channel matrix determine the potential of the MIMO system, for example the rank of $H$ determines how many independent signals can be sent [14].
CHAPTER 3. ELECTRONICALLY STEERED SATCOM RECIEVER

3.1 Introduction

As discussed in Chapter 2, when either a satellite or an earth based antenna is misaligned, communications between them can be disrupted. To remedy this an ESAF for dynamic tracking of a SatCom satellite has been designed.

Using ESAFs for SatCom is not widely researched. ESAFs are commonly used in radio astronomy, but these systems tend to be too expensive for commercial applications [15, p. 4].

Mechanically steered dishes are often used for tracking satellites but these are bulky and can be slower than electronically steered systems. An example of a hybrid, using a phased-array antenna as well as mechanical steering, is the TracVision A7 by KVH [16]. This is made to be attached to a moving vehicle to allow for satellite television on the go. While it performs well, this system is still expensive.

An example of a phased-array system only is the Connexion by Boeing [17]. This worked and was implemented on some aircraft to provide satellite television and internet, but was discontinued due to its high cost [18].

The ESAF system proposed is intended to provide steering over smaller angles for SatCom at low cost. The radio astronomy community has demonstrated the utility of ESAFs with extremely low power signals. It is the intent of this research to develop similar technology for SatCom. This is believed to be possible by simplifying the ESAF design for an application that has much higher SNR than is common in radio astronomy.

3.2 Requirements

While the ESAF system is designed to add the functionality of tracking a moving satellite, it also needs to achieve SNR performance comparable to industry feeds. This will be tested by
measuring the performance of the ESAF as well as a standard industry low-noise block downconverter feedhorn (LNBF), and comparing the results.

The cost of the system, both parts and assembly, must be low enough so that it can be a practical purchase given different competing products. The system also needs to be light weight and compact enough to attach to current parabolic reflectors.

A summary of the system requirements is:

- Track signals using electronic beam steering.
- Achieve SNR performance comparable to traditional systems.
- Achieve a weight and size that are small enough to attach to a reflector.
- Achieve a Low cost system design.

### 3.3 The System Design

A block diagram of the system design is shown in Figure 3.1. Moving left to right it begins with a linear array of antennas, which forms the feed for the offset dish. Each of the antenna elements are followed by a low noise amplifier (LNA), and then by a variable gain amplifier (VGA). The LNAs play an important role by providing gain to the signal while adding limited noise, which is especially important for the first gain stage in a noise sensitive system.

The VGAs are what enables the system to shape the beam for the primary pattern which then steers the beam for the secondary pattern. These are controlled by a DC voltage input. The signals are combined after the VGA stage and the low-noise block downconverter (LNB) amplifies and mixes the signal down to the intermediate frequency (IF). The signal is then split between the receiver and the beam control segment.

As the signal continues through the beam control segment it passes through a DC block which is used to protect the RF amplifier. The signal is filtered before it passes through the power detector which converts the RF signal into a DC signal. This DC signal corresponds to the signal power. The DC output is then amplified to be in a range that the ADC can work with. The output of the DC amp is used to determine the signals to be sent to the VGAs.
The system is designed to dynamically steer the beam until it maximizes the SNR. Since there are other satellites nearby in the geostationary orbit the beam could be steered to a neighboring satellite. There are regulations made to limit possible interference from other satellites, therefore they are spaced such that each one has about a two to four degree slot in the orbit [10, p. 402]. The beam steering range for the ESAF system is designed to only be around four degrees, thereby reducing the possibility that the beam steers to the wrong satellite.

Since the design uses a four by one array, it has the capability to only steer in one dimension. The antennas are aligned so that the steering occurs along the geostationary orbit. This steering is achieved by controlling the gain of the individual VGAs. This is shown by using a simple expression known as the array factor,

\[ A = \sum_{n=1}^{N} I_n e^{jk \cdot \hat{r}_n}. \]  

(3.1)

where \( N \) is the number of elements in the array, \( I_n \) is generally the complex excitation for the \( n^{th} \) element, \( \vec{k} \) is the wave vector, and \( \hat{r}_n \) is the position of the \( n^{th} \) element. Since the system does not have phase shifters, \( I_n \) has no phase and so is just a magnitude and corresponds to the gain given by the VGAs. By only using VGAs for beam steering the cost of the system is reduced, and the design is simplified. The array factor is then multiplied by the radiating field of one antenna to get the total radiation pattern of the array. Changing the VGA outputs changes the primary patterns shape which then steers the secondary pattern.
Since given excitations steer the beam to a given direction, we identify the excitations as beam states. The beam states will be chosen so that they maximize the SNR for various orientations along the path of the orbit. These beam states will be stored and correspond to the given beam orientation to be used in our signal searching algorithm.

The system uses a three point search on the beam states to maximize the SNR. As mentioned previously the power received will correspond to the DC signal received from the power detector. The signal is sampled at the current and two neighboring positions and the system will then steer the beam to the beam state that had the highest SNR. Once it is at this new position the algorithm will repeat. The states must have beams wide enough to cover neighboring state so that as the ESAF is performing the three point search it won’t lose signal when on adjacent states.

### 3.4 Contributions

This system design required contributions from several different individuals in BYU’s SatCom research group. We will now go over the specific contributions made by the author to the overall system. First, will discuss some of the contributions that were made on the hardware design and characterization, such as the power detector and the filter, and then discuss testing which was done with several team members.

#### 3.4.1 Power Detector

A power detector is needed at the IF band. This is essential to having a working ESAF, because the power detector provides the information that is used to make steering decisions. The output of the power detector needs to be variable enough when the antenna points on and off the signal so that the algorithm will be able to differentiate to maximize SNR.

In order to understand the sensitivity needed by the power detector a downlink budget analysis is needed, which requires information about the satellite and receiver. We tested the system with television signals coming from the Galaxy 19 satellite and received in the ku band at 11.7 – 12.2 GHz. A map of the EIRP over different locations in the United States, for this satellite and band, is shown in Figure 3.2. As can be seen on the map, the approximate EIRP at the position of the receiver in Provo, Utah is 46.5 dBW. The distance between the position of the receiver on
the earth to the satellite can be calculated using,

\[ d = \sqrt{a^2 - 2ar\cos(\phi)\cos(\lambda) + r^2}, \]

where \( r \) is the radius of the earth, \( a \) is the radius of the geostationary orbit, \( \phi \) is the latitude of the position on the earth, and \( \lambda \) is the longitude [19]. The distance between the receiver and the satellite is calculated to be 38,028 km. The gain of the receiver as given in the commercial dish data sheet is 39.3 dB. This gives the power at the output of the receive antenna, from Equation 2.12, to be \(-89.8\) dBm. Because of the gain given by the LNAs, the LNB, the IF amp, and the loss from the divider and filter, the power at the input of the power detector is 0.22 dBm. There will be other losses from pointing error, receiver losses, atmospheric absorption losses, and gain from the VGAs, so this just gives an estimate of the power requirements needed for the power detector. The actual power should vary around this value for different settings of the VGAs.

To verify the power estimates for the system, I measured power at a couple of different stages in the commercial as well as ESAF systems, the results of which are shown in Table 3.4.1. The table shows how well the estimates match the measured. Carrier-to-noise ratio (CNR) is used.

---

Figure 3.2: This image is an Intelsat Galaxy 19 Ku-band footprint map of the EIRP [1].
Table 3.1: Compares the power reading after the LNB for both commercial and the ESAF, and each of these either directly or after a power splitter.

<table>
<thead>
<tr>
<th>Configuration</th>
<th>CNR (dB)</th>
<th>Power Measured (dBm)</th>
<th>Power Estimated (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Commercial Direct</td>
<td>14.4</td>
<td>−32.5</td>
<td>−29.8</td>
</tr>
<tr>
<td>Commercial Split</td>
<td>13.9</td>
<td>−39.1</td>
<td>−35.8</td>
</tr>
<tr>
<td>ESAF Direct</td>
<td>7.2</td>
<td>−6.5</td>
<td>−9.8</td>
</tr>
<tr>
<td>ESAF Split</td>
<td>7.1</td>
<td>−12.9</td>
<td>−15.8</td>
</tr>
</tbody>
</table>

in the table and is the SNR before the signal is demodulated. This is used to demonstrate that actual signal is being received while these power measurements are being taken. The estimated power in the table is from the link budget analysis, but only counts gains and losses from the LNB and/or splitter. The ESAF measurements were taken for when the system was optimized for boresight. There are some discrepancies which were expected and can be explained by the losses and gains not accounted for. For example the ESAF measurements have higher power than the estimated: this is most likely due to the gain given by the VGAs, which was not accounted for in the estimate. For various states the gain from the VGAs could be quite different. Taking these measured results from the ESAF the power that would be seen at the input of the power detector, accounting for estimated losses and gains after the splitter, is 3.1 dBm.

A power detector is selected from Mini-Circuits (Fig 3.3) that would be suited for our power ranges. It is designed for the 10 – 8000 MHz band. It is powered by five volts of DC and detects power from −60 dBm to 5 dBm, which is a sensitivity range suited for our purpose.

To test the performance of the power detector, a signal with a 1.355–1.365 GHz band, was put into the power detector and the output voltages were measured. The results are recorded in Figure 3.4 where the voltage is shown to decrease with an increase in power.

3.4.2 IF Passband Filter

When initial tracking was attempted there was not an appreciable difference in the voltage output of the power detector when the beam was steered, which made tracking impossible. This was because the power detector was receiving signal over its entire bandwidth. Power was being received from frequencies that were not associated with the desired signal. By filtering out fre-
The power detector used in the ESAF system. It is a Mini-Circuits power detector and is identified as ZX47-60-S+.

The power measurements were taken with an input signal over the 1.355-1.365 GHz band.

When the frequencies that were not in the desired band the power received by the power detector more directly corresponded with the signal and the power detected changed significantly from on the signal to off the signal, thereby allowing tracking.

The filter was designed in ADS to have the power measured by the power detector to better reflect the signal received. Figure 3.5 is an image of the manufactured filter and Figure 3.6 demonstrates the performance of the filter as measured on a network analyzer. Once this filter was added to the system as shown in Figure 3.1, the system was able to track a moving signal source.
3.4.3 Testing

The system was assembled and tested on the roof of the Clyde building on BYU campus. Several research group members assisted in this process. To compare the performance of the commercial LNBF with our design, we tested both on a dish. The commercial LNBF setup is shown in Figure 3.8 and our system is shown in Figure 3.7.

The ESAF system requires that beam states are pre-selected to be used when maximizing SNR, thus a calibration of states is required. To calculate SNR for each state, the noise power and the signal power need to be measured. The noise power was measured by pointing the dish off of the signal, at the sky and the voltage output of the power detector was measured for each state.

We then pointed the dish at different points along the geostationary orbit, and then adjusted the VGA’s until the approximated SNR was maximized. This was done by taking the voltage
corresponding to the noise power measured for the current state and dividing it by the voltage corresponding to the signal power. Whatever state maximized this value was our state that maximized SNR. The noise power and signal power were inverted since the voltage output goes down for higher power input in the power detector. The state would then be saved and correspond to the angle that the dish was pointing off of the target. This was done for several positions over the opening angle $-3^\circ$ to $3^\circ$ relative to the satellite location, and then stored. We steered to these points, which were equally spaced along the geostationary orbit, using a mechanical motor on the dish.

The ESAF and three point search algorithm were tested with these stored beam states over the same opening angle. The system was able to track the signal as the dish was mechanically steered, but there were some limitations to the speed at which it could move. If the dish was moved too quickly the algorithm was not able to keep up. This is believed to be caused by a delay
in the output of the power detector, which causes the previously measured power to carry over into the next power measurement.

The ESAF was also optimized for each position using the pre-selected beam states over the steering angle mentioned above and then the SNR was measured. We also took the commercial horn and measured the SNR for each orientation. The ESAF system received the satellite signal well enough to watch satellite television for a broader angle than the commercial system. The results of the test of the two systems is shown in Figure 3.9.

We were able to receive a strong enough signal to watch television for over two degrees off of boresight in both directions for the ESAF, while the commercial feed was only able to do so for slightly over one degree. The measurement of the SNR of the ESAF is shown to oscillate. This is because there were only a limited number of beam states selected. When the dish was oriented such that the signal was coming at or near where a pre-selected beam state was optimized, the SNR was at a peak. By selecting more states, the change in SNR due to this effect would be reduced.
Figure 3.9: ESAF SNR performance over the steering angle is compared to the commercial system LNBF. The line labeled horn is for the LNBF.

Figure 3.9 shows a third plot of what the SNR would look like if the signal was optimized using all available beam states at each position.

While the ESAF system was shown to maintain the signal at a wider angle than the commercial system, the commercial system had a much better SNR performance. This was expected at least in part because the LNA noise figure is typically 1 dB and can reach as high as 2.3 dB, while the LNBF has a noise figure of 0.3 dB. However the difference is still greater than expected, and further investigation is needed to find the cause of SNR degradation. This degradation could be from poor matching between components on the board with the antennas, LNAs, and VGAs. Also much more gain is coming out of the ESAF system to the receiver than with the LNBF, from the LNAs and VGAs. This means the receiver could be clipping the signal it receives from the ESAF.

3.5 Summary

An ESAF was designed, fabricated, and tested. It was demonstrated that beam steering could be accomplished using a low cost feed. The ESAF also demonstrated a wider steering angle than a traditional LNBF. While there is still needed progress to be made such as reducing the SNR.
penalty, and making the system compact, we were able to successfully prove the concept of beam steering for a SatCom receiver.
CHAPTER 4. LOW PROFILE NATIVE DUAL CP ANTENNAS FOR A SATCOM REFLECTOR FEED

4.1 Introduction

As discussed in the introduction feed antennas for dishes in SatCom are most commonly horns which are bulky and heavy, as well as expensive to make. This motivates making light low profile antennas. In this chapter two low profile dual CP antenna designs are discussed.

The most common dual CP planar antennas use hybrids. These antennas have a quadrature hybrid that is connected to the ports of a dual linear polarized antenna. One output port of the hybrid connects to one port of the antenna and the other connects to the orthogonally polarized port. When a signal is put into one of the input ports of the hybrid it will excite one linear polarization and then the 90° delayed signal will excite the orthogonal polarization. This creates a circularly polarized signal. By exciting the other input port of the quadrature hybrid the orthogonal CP polarization will be radiated, thereby the hybrid enables a dual CP antenna.

These antennas can have several qualities such as high aperture efficiencies and quality CP, but they generally do not have high radiation efficiencies. This is due to the added network which comes from the hybrid. This causes more losses in the antenna. An antenna that operates similarly is discussed in [20]. It has a Wilkinson divider that excites two slots at perpendicular linear polarizations, with a ninety degree delay between the two polarizations. But like the hybrid the added network causes losses.

Another dual CP antenna is discussed in [21]. It involves a microstrip to slotline transition, and then the slotlines form a circle and excite patch antennas. These patch antennas are linear and use sequential rotation to achieve CP. The slotlines are traveling wave feeds and are designed so the patches are excited at the appropriate delay. There are two slot lines which run next to each other but are excited by different microstrip lines, which allows for the excitation of orthogonal circular
polarizations. While the design is clever its involved feeding technique requires transitions, as well as a traveling wave feed, which is expected to increase dielectric and conductor losses.

There are few low profile dual CP designs developed to date, and none address the issue of achieving high radiation efficiencies while decreasing the size, weight, and complexity of the design. The two antenna designs in this chapter are intended to address these open research challenges. The challenge comes in creating an antenna that achieves this without sacrificing antenna performance for other parameters. Many low profile native dual CP antennas have been attempted in BYU’s SatCom group, but the challenge of maintaining high isolation as well as low return loss while doing so has not been achieved.

A summary of design requirements for these antenna designs is to achieve:

- XPI above 20 dB for the secondary pattern.
- Radiation efficiency above 90%.
- Spillover and aperture efficiencies above 70%.
- Isolation between the two ports above 20 dB over the 11.7 – 12.2 GHz bandwidth.
- Return loss for both ports above 10 dB over the 11.7 – 12.2 GHz bandwidth.

### 4.2 The Cross Antenna as a Dual CP Microstrip Antenna

The antenna designed in this section is based on the cross antenna designed by Roederer in [22]. It is a traveling wave antenna as opposed to a standing wave or resonant antenna, such as a rectangular patch antenna. Traveling wave antennas are antennas that radiate as a voltage wave propagates along them [12]. This cross antenna design is a winding wire over foam and underneath the foam is a ground plane. See Figure 4.1 for the form of the wire. The winding wire terminates in the second port. Due to the antenna’s symmetry the orthogonal polarization is achieved when the second port is excited. The simplicity of the feeding system of the antenna helps in achieving low dielectric losses from the network, which makes this a good starting point for a high efficiency dual CP antenna.

The cross antenna uses similar principles to achieve CP as sequential rotation. Sequential rotation is an array design method used to achieve CP by taking linearly polarized antenna elements
and rotating them. Each of the elements is placed and oriented such that they have the proper phase delay to allow the combination of all the antennas linear polarized waves to form a CP wave. For example, if there are $N$ antennas, each antenna element is oriented such that it is rotated $N/2\pi$ around its center from the orientation of the element previous to it. Each antenna is placed on a circle and is separated by an angle of $N/2\pi$ from the adjacent antennas. The feed is placed so that each antenna is excited at a phase delay of $N/2\pi$ from the previous antenna going either in a clockwise order if LHCP, or in a counter clockwise order if RHCP. The cross antenna operates similarly, by radiating a linearly polarized wave first at the first arm, but will be delayed in radiating from the other arms as the wave travels around the antenna. By careful design the arm lengths are designed such that the phase delay will be the appropriate amount to create CP, as in sequential rotation. The theory of the operation of the design is discussed more fully in [22].

There are several design variations to the cross antenna, but the one displayed is chosen due to its symmetry. Some of the variations of the cross antenna have multiple wraps, which makes it difficult to use the other end of the wire for the orthogonal polarization as the radiation pattern would be different. Therefore using the symmetry of the current design allows for identical behavior between the two ports. Preserving this symmetry is beneficial because it simplifies the design process. If one port works the other should work similarly, allowing us to focus on the performance of one port, decreasing simulation time.

The proposed design will be based on the cross antenna and put a similar design on a PCB, make it two port for dual CP, and tune it to have a center frequency at 12 GHz. This antenna has many parameters which creates a rich design space. By putting the cross antenna on a PCB the manufacturing process becomes easier, than forming a wire over foam and ground plane as in the original design. This is because of the standardized manufacturing processes for PCBs. The substrate chosen was RT/duroid 5880 made by Rogers Corporation. The substrate thickness is about 1.57 mm. The antenna is via fed at each port with the connectors directly connected on the underside of the antenna. The center frequency of the design is at 12 GHz, with a wavelength around 25 mm. The diameter is just over a wavelength. Figure 4.1 shows the top layer of the antenna, which would all be placed over a dielectric core and ground plane.
4.2.1 The Simulated Antenna Performance

The model was designed and tuned for the given substrate and center frequency. This was done using the finite difference time domain (FDTD) solver on the commercial package Empire XCcel. The far-fields results simulated from Empire were then imported into in house physical optics (PO) reflector model. The reflector model used these fields to compute the XPI, SNR, and the radiation, spillover, and aperture efficiencies. The PO reflector model was set for a LNB with a noise figure of 0.8 dB and a reflector with an $f/D$ of 0.9. This $f/D$ means that the opening angle of the reflector opens from $-31^\circ$ to $31^\circ$.

The simulated gain elevation pattern cuts of the antenna for both RHCP and LHCP at the azimuthal angles $0^\circ$ and $90^\circ$ are shown in Figure 4.2. The gain of the antenna peaks at boresight at 12.2 dB. This boresight gain is an attractive feature as it does not require an array of antennas to meet gain requirements for a dish. As can be observed in the figure, there is a large sidelobe in the RHCP $90^\circ$ cut. This is caused by an uneven distribution of energy being radiated along the antenna, which is a consequence of it being a traveling wave antenna. As the wave propagates along the antenna it radiates, which means less power is propagating the further the wave travels. This causes more power to be radiated along the first arms of the antenna than the later ones, thereby causing the sidelobe. This is also illustrated by the poor reflector illumination which can be seen from the simulated aperture efficiency of 45.2% and a spillover efficiency of 56.1%. The side lobe causes much of the radiated power to miss the reflector as is illustrated by the spillover efficiency.

The AR for azimuthal angles $0^\circ$ and $90^\circ$ are shown in Figure 4.3. The pattern cuts are less than 3 dB from elevation angles $-20^\circ$ to $15^\circ$. Initially this may not seem to be a wide enough AR to properly illuminate the dish to get quality CP, as the opening angle of the dish is wider than the
angle that the cross antenna radiates with an AR below 3 dB. After using the reflector model an XPI of 27.3 dB is simulated, which is better than our requirement of 20 dB. This is believed to occur because a larger amount of energy is radiated in the quality CP direction. This can be seen by observing how the gain trails off in the same area that the AR increases.

In Figure 4.4 the S-parameters are shown for the antenna. Due to the symmetry of the design, both ports have the same S-parameters. The return loss shows a satisfactory bandwidth around our center frequency. The $S_{21}$ on the other hand needs to be below -20 dB but is only around -12 dB. This causes a couple of problems. First, the ports are not well isolated, causing interference between the two signals. The other issue is that energy put into the system that passes through to the other port is not being radiated, causing the antenna to have a lower radiation efficiency. As simulated this antenna achieved a radiation efficiency of 88.3%.

### 4.2.2 Summary

The microstrip cross antenna has some attractive features such as high gain, quality dual CP behavior, ease of feeding, ease of manufacturing, and a fairly high radiation efficiency. However it has two major flaws. The first is the large sidelobe of the beam pattern, which causes much of the
Figure 4.3: Simulated AR pattern cuts, the AR is less than 3 dB from about -20 to 15 degrees.

Figure 4.4: Simulated S-parameters of the microstrip Cross Antenna.
radiated energy to be wasted. The second is the poor isolation between the two ports, which causes much of the energy that put into one port to be coupled to the other, degrading SNR performance as well as radiation efficiency. To have the antenna be useful in industry these two problems need to be addressed.

4.3 Dipoles over an EBG Surface

Since the design requirements were not met with the microstrip cross antenna design, another direction needed to be pursued. The main challenge with the cross antenna as well as with other designs attempted, is achieving quality dual CP performance while simultaneously having high isolation between the two ports. This issue was intended to be addressed with the next design. This design was based on an antenna with a dipole over electromagnetic band gap (EBG) surface by Yang et al [23].

EBG structures are generally three dimensional periodic structures. One characteristic they have is high impedance for waves in certain frequency bands in all directions [24]. EBG surfaces can be made to have these properties, but in two dimensions, by creating a structure that is periodic in two dimensions.

Since EBG surfaces have a frequency band where surface waves do not propagate, a lattice can therefore be designed to reduce surface waves in a certain band. For antennas this can be beneficial for several reasons, energy loss to the surface waves can be eliminated for the in band signal, smoother beam patterns can be achieved because there won’t be surface waves to radiate out the edges of the material, and mutual coupling can be decreased [25].

The EBG surface can also be designed so that the phase of a reflected wave can be changed from the incident wave’s phase. This change in reflected phase can be anywhere in between $-180^\circ$ to $180^\circ$ [26]. This allows for linearly polarized antennas, such as a dipole antenna, to be placed nearly directly above the surface, if the reflected phase is designed to be in phase with the incident wave. Dipole antennas are usually placed a quarter wavelength above ground planes so that the reflected wave adds constructively with the wave transmitted from the dipole. Using an EBG surface allows the dipole to be placed just above the surface, allowing for a lower profile dipole antenna than would be possible otherwise.
The low profile single dipole antenna over an EBG surface designed by Yang et al. takes advantage of these properties of EBG surfaces. The design consists of a dipole antenna placed closely above an EBG surface. The EBG surface used consists of a two dimensional lattice of rectangular conductive patches above a ground plane, where each element in the lattice is connected to the ground through a via [25]. The lattice and the ground plane are separated by a dielectric material. This design therefore can easily be made on PCB materials with common manufacturing processes. Normally dipoles radiate linear polarization, but by placing it over an EBG surface and tuning the surface, it can also radiate CP. CP is achieved because the reflected wave has a $90^\circ$ phase shift while simultaneously being orthogonally polarized to the incident wave. This orthogonally polarized reflected wave is associated with the orientation of the dipole over the EBG surface. The feed is coaxial where one post of the dipole is connected to the inner pin and the other is connected to the grounded shielding. The theory of operation is explained more fully in [23]. Figure 4.5 shows picture of a model of the antenna designed by Yang et al. which was designed for a center frequency of 3.56 GHz. This model was created in Empire XCcel and simulated. The simulation results agreed with what was reported in the paper.

This antenna was chosen to be a base design for a dual CP antenna because it was believed that high isolation could be achieved between ports of orthogonal polarizations. This is due to two reasons: the radiating mechanism of the antenna and the high impedance surface. The antenna radiates CP by generating a linear polarized wave from the dipole antenna and transforming it into a CP wave by placing it over an EBG surface. Since dipole antennas are often used to form dual linearly polarized waves with high isolation between the ports, it is believed that by placing orthogonal dipoles over an EBG structure, high isolation might be accomplished because the dipole elements still radiate linear polarized waves, but are transformed by the EBG surface. Also the high surface impedance disallows surface waves in a certain frequency band to propagate, decreasing the coupling between the two ports.

An antenna was designed for the 11.7 – 12.2 GHz band, with the Yang et al. antenna as motivation. The second dipole was placed orthogonal to the first. Initially for proof of concept no feeding network was added, and lumped ports were used to excite the antennas. Figure 4.6 shows an image of the antenna. Each dipole length is around half a wavelength at 12 GHz. The antenna
is symmetric therefore the performance for both ports are similar. This was shown by simulation, therefore the performance analysis of the antenna will all be shown for one port.

The gain pattern is shown in the $0^\circ$ and $90^\circ$ cuts in Figure 4.7. The max gain is at boresight and is 3.8 dB. The pattern width in the $90^\circ$ cut is much wider than the pattern width in the $0^\circ$ cut. The pattern is not symmetric but could be addressed by arraying the antennas, which due to the limited gain is required anyways. The CP quality for the antenna as a feed is shown by the AR cuts shown in Figure 4.8. The AR cuts show that at boresight the CP performance for the primary pattern is not below the 3 dB limit. This is not ideal and may be improved with further tuning, but for an initial analysis we can see that the antenna has CP behavior. The radiation efficiency for the element was simulated to be quite high at 95%. The performance of this as the antenna was simulated without a feeding network is shown in Figure 4.6. The simulated S-parameters are shown in Figure 4.9. The $S_{21}$ for this antenna is about -25 dB at the center frequency, which is encouraging since it is below the requirement discussed at the beginning of the chapter.

From this initial analysis the dipole over EBG appeared promising, so a design that made a more symmetric pattern was attempted. We did this by using a technique known as the eleven
Figure 4.6: Dual dipole antenna over an EBG surface.

Figure 4.7: The gain pattern cuts of LHCP and RHCP for the LHCP port for the cross dipole over EBG surface antenna.
Figure 4.8: AR pattern cuts for the dual CP dipole over EBG surface antenna.

Figure 4.9: The S-parameters of the dual CP dipole over EBG surface antenna.
feed, which is done by arraying two dipoles. We arrayed dipoles for both polarizations, and this is shown in Figure 4.10.

The eleven feed design requires two cores and four layers. This is to separate the radiating portion from the feeding network. The top layer is for the radiating portion, the middle layers are for ground, and then the bottom layer is used for a feeding network. This is shown in Figure 4.11. The top core is a 32 mil thick dielectric and the bottom core is a 20 mil thick dielectric. Both cores are made out of Rogers 4003c board, the dielectric of which has a relative permittivity of 3.55. The network for the design is being integrated with the radiating portion and will be added to this thesis when optimized.

The simulated performance, without the feeding network, of the gain pattern, AR and the S-parameters is shown respectively in Figures 4.12, 4.13, and 4.14. The gain pattern of the antenna demonstrates the improvement that the eleven feed design gives the pattern shape as well as the maximum gain. It is much more symmetric than without the eleven feed, and therefore increases its applicability to be used for a dish feed. The gain is still lower than desired. The radiation efficiency is 91%, which meets the stated requirements.
The axial ratio unexpectedly improved, which could be due to a combination of more carefully tuning the design and arraying the antennas. The S-parameters demonstrate quality isolation as well as 10 dB return loss.

Each dipole element consists of two posts, where one is connected to the signal and the other post is grounded. As mentioned in Chapter 2, when feeding dipole antennas a balun is required so that an even amount of current is pulled from both posts. There are many different designs to these baluns. At first glance the dipole over EBG antenna seems as if there is no balun, however the EBG surface itself performs this function. The EBG surface has a high impedance for the in band signal, and so does not allow surface waves to propagate in that band. This high impedance for the in band signal causes the EBG to act as a balun for the dipole. Current cannot be pulled from the ground because the EBG blocks the in band surface waves, thereby causing currents to only come from the post. This causes the currents on both posts of the dipole to be equal and the radiation pattern of the dipole is preserved.

A microstrip feed design is needed to be integrated with the radiating portion to finish the design. The feed will combine each of the like polarizations, and is shown in Figure 4.15, this is the bottom layer of the antenna. Each polarization will be fed with an SMA surface mount connector.

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Figure 4.12: Gain pattern of dipoles over EBG in eleven feed arrangement.
Figure 4.13: AR of dipoles over EBG in eleven feed arrangement.

Figure 4.14: S-parameters of dipoles over EBG in eleven feed arrangement.
Figure 4.15: Feeding network on bottom layer for the dipole over EBG in the eleven feed arrangement.

These will be centered over the circles shown in the figure and oriented so that the feet sit on the pads which have vias connecting them to ground. The more central surface mount connector pad is feeding the simplest port feed design, which consists of transformers to match the 50Ω dipole feed to one of the 100Ω ports on the combiner.

The surface mount connector pad to the top left of the figure is a similar design with transmission lines that moves the pad away from the center so that the needed separation between ports can be achieved. Models of the SMA connectors were added to the total feed model so that simulated results would be accurate.

The microstrip feed design was integrated and tuned with the radiating portion. The simulated AR performance for the LHCP port is shown in Fig 4.16 and the radiation pattern for the same port is shown in Fig 4.17. As can be seen radiation performance is maintained with the added feed.
Using the physical optics reflector model, for a \( f \) over \( d \) of 0.6, we simulated an aperture efficiency of 50% and a spillover efficiency of 65%, also for the LHCP port. These are not to specifications therefore ways to improve the radiation pattern need to be investigated further. This could be studied by experimenting with the dipoles spacing and how it effects the beam pattern. The radiation efficiency was at 91% which showed no degradation from the simulations without a microstrip feed network. The XPI for the LHCP port is 21 dB and so meets specifications.

The S-parameters for the LHCP port are shown in Fig 4.18. High isolation is demonstrated as well as a \( S_{11} \) that meets specifications. We need the S-parameters of the other port to understand if the antenna fully meeting the specifications for the S-parameters.

The RHCP port was also simulated but the design is slightly different than what was simulated for the LHCP port due to time constraints. The effect should be minimal due to the parts of the design that were altered. The simulated AR performance for the RHCP port is shown in Fig 4.19 and the radiation pattern for the same port is shown in Fig 4.20.

Using the physical optics reflector model we simulated an aperture efficiency of 50% and a spillover efficiency of 63%, also for the RHCP port. The radiation efficiency was at 89% which showed no degradation from the simulations without a microstrip feed network. The XPI for the RHCP port is 19.5 dB just below specifications. The S-parameters for the RHCP port are shown in Fig 4.21. High isolation is demonstrated. The \( S_{22} \) is shifted but could be tuned to the correct frequency.

**Coax Feed Design**

Since the microstrip feed design requires multiple layers the cost of fabrication is quite expensive. The design has global vias and blind vias on both cores, which increases the complexity of the manufacturing process. This motivates a simpler design to test for proof of concept for the radiating portion of the antenna. We did this by using one core, and feeding each dipole by a coax feed, and combining the like polarizations with a splitter. This simplifies the antenna design dramatically and allows for a much lower fabrication cost, allowing for proof of concept for the radiation portion of the design at a more practical expense.
Figure 4.16: The simulated AR performance for the LHCP port for the dipole over EBG in Eleven arrangement with integrated microstrip feed.

Figure 4.17: The simulated radiation pattern performance for the LHCP port the dipole over EBG in Eleven arrangement with integrated microstrip feed.
Figure 4.18: The simulated S-parameters for the LHCP port the dipole over EBG in Eleven arrangement with integrated microstrip feed.

Figure 4.19: The simulated AR performance for the RHCP port for the dipole over EBG in Eleven arrangement with integrated microstrip feed.
Figure 4.20: The simulated radiation pattern performance for the RHCP port the dipole over EBG in Eleven arrangement with integrated microstrip feed.

Figure 4.21: The simulated S-parameters for the RHCP port the dipole over EBG in Eleven arrangement with integrated microstrip feed.
Figure 4.22: Dipoles over EBG in the eleven feed arrangement with ports fed by coaxial cables.

Figure 4.23: Side view of dipoles over EBG in the eleven feed arrangement with ports designed to be fed by coaxial cables.

The design was tuned again since the dipole thickness changed because it was determined by the inner conductor of the coaxial cable. The top view is shown in Figure 4.22, and the side view is shown in Figure 4.23 which shows the coax feeds that were part of the model.

The simulated beam pattern for this antenna is shown in Figure 4.24, and achieves similar gain as does the multilayer design with the microstrip feed. The pattern is also well shaped with low side lobes. The axial ratio in Figure 4.25, demonstrates the CP behavior of the antenna. The S-parameters are shown in Figure 4.26, which shows the insertion loss to achieve the 10 dB requirement and the coupling is less than 20 dB. The S-parameters were computed by taking the
simulated S-parameters from HFSS and importing them into an object in ADS and simulating a feeding network that includes the splitters.

**Fabrication and Measurements of Coax Feed Design**

The coax feed design of the dipoles over EBG in eleven arrangement antenna was fabricated and is shown in Figures 4.27 and 4.28. The PCB was manufactured by a PCB fabrication company. Coaxial cables where cut and the exterior layers were stripped on the ends. These ends were used to penetrate the EBG surface and to be used for the dipoles. The outer conductor was then soldered to the ground on the back of the EBG surface. The inner conductor was then inserted through holes in the PCB and bent for initial measurements, as shown in the figures just mentioned. After initial measurements some modifications were made on the dipole rods. Because bending the posts was not as close to the simulated model, posts were soldered perpendicular to the penetrating posts so that a right angle could more accurately be approximated. This was quite challenging and the process could still be improved due to the size of the posts as well as the design requiring the horizontal rods to be very close to the EBG surface. This challenge brought about several issues with the fabricated antenna. The solder balls add conductive parts that were not in the simulated
Figure 4.25: AR for dipoles over EBG in eleven arrangement with coax feed.

Figure 4.26: S-parameters for dipoles over EBG in eleven arrangement with coax feed.
model. If the soldering is not good enough then the radiation behavior could be changed from what is desired. Also aligning the dipoles properly was challenging and so they may not be situated perfectly.

Another challenging part of fabricating this design was making sure that the coaxial cables leading from the antennas to the tee splitters were the same length. If they are not the same length then the signals will be out of phase when being combined which would decrease signal quality. In spite of these challenges the antenna as fabricated and tested.

The S-parameters of the antennas were measured using a network analyzer. The AR measurements were more difficult. A commercial dual CP horn antenna was used as a reference antenna. The horn antenna and dipole over EBG antennas were pointed at each other, and where far enough apart to be in each others far-fields. The dipole over EBG antenna was used as a receiver and the signal received was viewed in a spectrum analyzer. A signal generator was connected to the horn antenna. For a single receiver polarization the power received when a RHCP and a LHCP signal were recorded. The ratio of these powers was calculated which gave the XPI for the port of the receiver. This was done for both receiver ports. From these XPI values the AR was computed. It was challenging to point the antennas at boresight, so the measured values shown in the following plots are just approximations of the boresight AR. It would have been desirable to characterize an antenna with known AR performance but only one dual CP antenna was available. This would help verify the legitimacy of the test procedures.

The S-parameters of the measured and simulated performance of the antenna are shown in Figure 4.29 and 4.30. The multiple resonant points as shown in the measured version of $S_{11}$ are believed to be from the standing waves on the coaxial cables. The $S_{11}$ resonance is also shifted below the center frequency of 12 GHz. The isolation is shown to be fairly good over the interested band. It is below $-20$ dB over most of the band but it does get a little higher than that over the lower part of the bandwidth. While these results are not perfect they are encouraging. The isolation is demonstrated to perform near specifications while the fabricated antenna has some clear flaws. With improved fabrication techniques, which would require a little more experimenting, the performance of the antenna would be improved. Also the previously proposed design simulated with better performance for isolation and so might perform better when fabricated as well. The
Figure 4.27: Front view of fabricated coax feed design of dipoles over EBG in eleven arrangement.

Figure 4.28: Coax feed design which shows coax cables coming off of the back with t-splitters to combine ports with like polarizations.
measured S-parameters demonstrate the validity of the design as a dual CP antenna that can achieve high isolation between ports.

In Figure 4.31 shows measured versus simulated performance of the CP quality of both ports of the antenna. While the measured performance did not achieve the same quality of CP as
was simulated, it demonstrated that over some of the frequencies the antenna did achieve quality CP performance.

Some further modifications were made to the antenna, such as re-aligning and re-soldering the dipoles, as well as shortening some of their lengths. The measured results for these modifications are shown in Figures 4.32 and 4.33. The S-parameters still show the isolation performance, but it is shifted from before. The measured results in the AR plots are averaged from two different measurements. It shows a slight improvement in AR performance for the RHCP but a degradation for LHCP. The difference in performance could be explained by fabrication errors due to difficulties explained earlier.

4.3.1 Summary

This antenna has achieved many of the goals that it was designed for, such as quality CP, high radiation efficiency, a well shaped beam pattern, high isolation, high return loss, as well as size and weight requirements. These were demonstrated in a full wave model simulation of the design. Also some of the results such as isolation and CP quality were verified in the fabrication of the coaxial fed design. The gain obtained is lower than desired and could be solved by adding more elements in an array. The results obtained demonstrate promise for future designs and applications.
Figure 4.32: Simulated and measured $S_{12}$ for dipoles over EBG in eleven arrangement with coax feed, modified from previous version.

Figure 4.33: Simulated and measured AR for dipoles over EBG in eleven arrangement with coax feed, modified from previous version.
CHAPTER 5. TESTING GPS DENIED RADAR WITH A PHYSICAL OPTICS SCATTERING MODEL

5.1 Introduction

A MIMO radar is being developed to perform self localization. The motivation for a GPS denied positioning system was discussed in Chapter 1. Outdoor positioning systems used without GPS generally use external structures. These hardware structures are varied and can include towers and electromagnetic coils [27, p. 3]. The MIMO self localization method discussed in this chapter is a self contained radar used as a positioning system. This chapter will discuss how the radar works, the physical optics scattering model, and an analysis of the MIMO radar using the physical optics scattering model.

The physical optics scattering model provides a method for verifying performance and developing the radar algorithm, allowing the radar to be better prepared for testing. Insights can also be gained about the navigation algorithm and the MIMO radar. With these benefits it is intended that the radar will be developed quicker.

5.2 The MIMO Radar

The MIMO radar consists of $M$ receive and $N$ transmit antennas. These antennas will be directed so that their main beam will be downward. As the UAV flies it will transmit signals and then receive the back scattered waves. From these signals the channel matrix for each position can be obtained, and from a series of channel matrices a correlation matrix can be formed. By comparing the correlation matrices that the radar actively acquires to the stored correlation matrices, the UAV can get a prediction of its position. The accuracy of this position is dependent on the richness of the multipath environment. This richness can be determined by calculating the rank of the channel matrix.
5.2.1 Calculating and Comparing Correlation Matrices

A correlation matrix $C$ uses a window of consecutive channel matrices. We will signify the window size by $W$. We form $C$ by reshaping a series of channel matrices, $H_{p:p+W}$, where $p:p+W$ are the indices of the channel matrices obtained along the flight path. Each column of this matrix is a vector formed by the concatenation of each of the columns of a single channel matrix. The dimension of this matrix is $MN \times W$. From the matrix $H_{p:p+W}$ we form the correlation matrix $C$ by

$$C = \frac{1}{W} H_{p:p+W} H_{p:p+W}^T,$$

which will be $MN \times MN$.

As a UAV flies over a path a correlation matrix can be formed for each position from the current channel matrix and the previous $W$ channel matrices. When passing over this area again to find the current position the UAV can use the correlation matrix that is being obtained and compare it to all the stored correlation matrices by [28]

$$e_p = 1 - \frac{tr(C^0C^p)}{||C^0||_F||C^p||_F},$$

where $C^0$ is the reference correlation matrix just obtained and $C^p$ is the stored correlation matrix at position $p$. This is a correlation matrix distance and therefore $e_p$ is always between zero and one. This metric is discussed more fully in [28]. It was demonstrated to be a useful metric to give information about spatial characteristic changes. The stored correlation matrix that minimizes this error corresponds to the position that we would predict the UAV to be.

5.3 The Physical Optics Scattering Model

Radar generates electromagnetic waves that propagate to the underlying terrain and induce scattered fields, as illustrated in Figure 5.1. The scattered fields that reflect back to the radar will
be picked up by the receive antennas. This is the physical phenomenon that will be simulated in the physical optics model.

![Figure 5.1: An incident wave on a surface with the resulting scattered fields.](image)

The physical optics approximation is used because it is faster and uses less memory than solving the problem using numerical methods obtained directly from Maxwell's equations. The approximations made will allow for an accurate representation of reality while lessening computing time and memory usage.

For convenience the following symbols will be defined.

$\vec{E}_i$: The electric field incident on the surface from the transmit antenna.

$\vec{H}_t$: The magnetic field incident on the surface from the transmit antenna.

$\vec{E}_s$: The electric far field scattered from the surface.

$\vec{J}$: The current density that creates an equivalent scattered field.

$\vec{r}$: The position in the far field.

$\vec{r'}$: The position over the radiating surface.

$\omega$: The frequency of the incident electromagnetic wave.

$k$: The magnitude of the spatial frequency of the incident wave.

$\mu$: The permeability of free space.

$\phi$: The azimuthal angle as measured from the x-axis.

$\theta$: The elevation angle as measured from the z-axis toward the azimuthal position.

The incident field generated by the antenna will be approximated as a plane wave, where the magnitude and polarization are determined by the antenna being modeled. By using the physical
optics approximation [29, p. 340],

\[ \vec{J} = 2\hat{n} \times \vec{H}, \]  

(5.3)

we calculate an equivalent current density on the surface. This current density will be used to find the scattered field from the surface. This equation comes from the current density generated from an incident wave on a flat perfect electrical conductor (PEC) plane. Therefore, this approximation only works when the scatterer, the surface in this case, is approximately flat on the order of the wavelength [29, p. 339].

Once the surface currents are computed they are used in the far field radiation integral [30, p. 234] to compute the scattered field, which is

\[ \vec{E}_s(\vec{r}) = -j \omega \mu (1 - \hat{r} \hat{r}) \cdot \frac{e^{-jkr}}{4\pi r} \int e^{jk\hat{r} \cdot \vec{r}'} \vec{J}(\vec{r}') d\vec{r}' . \]  

(5.4)

This calculates the electric field at the receive antenna by using the equivalent current density, \( \vec{J} \), on the surface.

### 5.3.1 Random Rough Surface Generator

A terrain profile generator will be needed as part of the model. Terrain profiles will be used to test the accuracy of the physical optics model as well as to test radar performance.

Some rough surface terrain types can be modeled by random processes representing terrain height. Surface characteristics can be manipulated by controlling the spatial frequency content and the variance of the random processes. The following derivations show how this is done.

A random surface height, \( z'(x, y) \), is represented by a realization of a random process and will be specified by a surface height variance \( \sigma^2 \) and a power spectral density (PSD). We will start with a realization of a white band limited Gaussian process, \( z(x, y) \), the height of an initial surface, which we will manipulate mathematically to have the desired statistical properties of the random process of the surface height, \( z'(x, y) \). The noise will be band limited, because when the model is implemented in code the surface will only be sampled at points separated by \( \Delta x \) and \( \Delta y \) for the respective dimensions. The realization of the random process of the initial surface, \( z(x, y) \), will
have a PSD, $S_i(k_x,k_y)$, where

$$S_i(k_x,k_y) = \begin{cases} \frac{N_0}{2}, & |k_x| \leq k_{x,max} \land |k_y| \leq k_{y,max} \\ 0, & \text{otherwise} \end{cases}, \quad (5.5)$$

where $k_{x,max}$ and $k_{y,max}$ are respectively the maximum spatial frequencies in the $x$ and $y$ dimensions. These spatial frequencies will depend on the distance between surface samples in their respective dimensions and using Nyquist they are

$$k_{x,max} = \frac{\pi}{\Delta x}, \quad (5.6)$$

and

$$k_{y,max} = \frac{\pi}{\Delta y}. \quad (5.7)$$

From this PSD we can find the variance of the random process of the initial surface $z(x,y)$, which we will identify as $h_i^2$. Using the fact that the variance and the integral of the PSD are equal to the average power of the initial surface, we have

$$h_i^2 = 2k_{x,max}k_{y,max}N_0. \quad (5.8)$$

From this variance we can define our initial surface random process which results in the realization $z(x,y)$. It is from this initial surface realization that we will manipulate into our desired surface height, $z'(x,y)$.

First we take the Fourier transform (FT) of $z(x,y)$,

$$Z(k_x,k_y) = \mathcal{F}\{z(x,y)\}. \quad (5.9)$$

For the next step we introduce a filter, $H(k_x,k_y)$ that is determined by the PSD of the desired surface, which in turn will depend on the surface height variance, $h^2$ of the same surface. The
derivation of this filter will be shown shortly. We can filter \(Z(k_x, k_y)\) by \(H(k_x, k_y)\), to obtain

\[
Z'(k_x, k_y) = H(k_x, k_y)Z(k_x, k_y).
\]  (5.10)

By filtering with \(H(k_x, k_y)\), the surface height random process will have the desired PSD and the variance \(h^2\).

Finally we take an inverse FT of \(Z'(k_x, k_y)\) that results in the generated surface height

\[
z'(x, y) = \mathcal{F}^{-1}\{Z'(k_x, k_y)\}.
\]  (5.11)

This process can be thought of as simply taking the initial surface height random process and putting it into the frequency domain, selecting the frequency content that is desired, and transforming it back into the spatial domain. The following section will show how the frequency content is selected.

![Random 2-D Surface](image)

Figure 5.2: A randomly generated 2-dimensional surface, with a surface height variance of 6.3 mm.
The Power Spectral Density of the Surface Height

The PSD will be needed as mentioned earlier to help determine the filter \( H(k_x,k_y) \) in \( k \)-space. The following derivation will show this relationship mathematically. For this derivation the PSD of \( z'(x,y) \) will be defined as \( S(k_x,k_y) \).

First, the power spectral density is equal to the Fourier transform of the autocorrelation function \( R_{zz'}(x,y) \). Using this definition the relationship between the PSD and the filter \( H(k_x,k_y) \) is derived,

\[
S(k_x,k_y) = \mathcal{F}\{R_{zz'}(x,y)\} = \mathcal{F}\{z'(x,y)\}\mathcal{F}\{z'(x,y)\}^* = H(k_x,k_y)\mathcal{F}\{z(x,y)\}\mathcal{F}\{z(x,y)\}^* = |H(k_x,k_y)|^2|\mathcal{F}\{z(x,y)\}|^2 = |H(k_x,k_y)|^2S_i(k_x,k_y).
\]

By constraining \( S_i(k_x,k_y) \) to be equal to one over its bandwidth, and assuming that \( H(k_x,k_y) \) is real only, solving for the filter gives

\[
H(k_x,k_y) = S(k_x,k_y)^{1/2}, \tag{5.12}
\]

which shows that to generate our surface we need to define our PSD.

Once again using the relationship that the average power in the surface is equal to \( h^2 \) and the integral of the PSD, we have the following result,

\[
h^2 = \int_{-\infty}^{\infty} \int_0^{2\pi} S(k_x,k_y)k \, d\phi \, dk, \tag{5.13}
\]

where \( k = \sqrt{k_x^2 + k_y^2} \).

The physical optics approximation is made for surfaces which are approximately flat on the order of the incident wave length. Spatial frequencies with wavelengths that are longer than the surface in one of its dimensions should also not be included in the generated surface. Due to these requirements we will make the shape of the PSD annular to select only the desired fre-
quantities. This will be done by picking an inner radius, \( k_{\text{min}} \) which will be determined by the size of the terrain, and an outer radius, \( k_{\text{max}} \) which will be chosen so that the incident wavelength is approximately flat, for an annulus in k-space.

Only the amount of each spatial frequency in the PSD is left to be determined. Assuming a constant PSD, \( A \), across the chosen frequency band we have the following expression,

\[
S = \begin{cases} 
A, & k_{\text{min}} \leq k \leq k_{\text{max}} \\
0, & \text{otherwise}
\end{cases}
\]  

(5.14)

Then using Equation 5.13 and the PSD we have

\[
h^2 = \int_{k_{\text{min}}}^{k_{\text{max}}} \int_0^{2\pi} A k \, d\phi \, dk.
\]  

(5.15)

Integrating and solving this expression for \( A \) results in the following,

\[
A = \frac{h^2}{\pi(k_{\text{max}}^2 - k_{\text{min}}^2)}.
\]  

(5.16)

Now the PSD of the desired surface is defined. It depends on the surface variance and the limits chosen for the magnitude of the spatial frequency, both of which will be used as the controlling parameters to create the random rough surfaces.

**Calculating the Fourier Transform using the Fast Fourier Transform**

As shown previously, in generating the rough surface model, the Fourier transform (FT) and the inverse FT are used. Since our model will be using discrete points for the surface, a discrete approximation of these integrals needs to be made to generate the surface. The Fast Fourier transform algorithm will be used to make these approximations. The derivation of the approximation of the FT using discrete points will be done in one dimension, and then will be extended for the two dimensional surface.
The single dimension FT and inverse FT are

\[ Z(k) = \int_{-\infty}^{\infty} z(x) e^{-j k x} dx, \quad (5.17) \]
\[ z(x) = \frac{1}{2\pi} \int_{-\infty}^{\infty} Z(k) e^{j k x} dk. \quad (5.18) \]

Numerical approximations for the FT and inverse FT are as follows,

\[ Z(m \Delta k) \approx \sum_{n=1}^{N} z(n \Delta x) e^{-j (m-1) \Delta k (n-1) \Delta x}, \quad (5.19) \]
\[ z(n \Delta x) \approx \frac{1}{2\pi} \sum_{m=1}^{N} Z(m \Delta k) e^{j (m-1) \Delta k (n-1) \Delta x \Delta k}. \quad (5.20) \]

The discrete Fourier transform (DFT) and the inverse DFT can be computed with a Fast Fourier transform function. The DFT and the inverse DFT are

\[ Z[m] = \sum_{n=1}^{N} z[n] e^{-j 2\pi (m-1)(n-1) / N}, \quad (5.21) \]
\[ z[n] = \frac{1}{N} \sum_{m=1}^{N} Z[m] e^{j 2\pi (m-1)(n-1) / N}. \quad (5.22) \]

These can be used to calculate an approximation of the FT. By equating our numerical approximations with the DFT and the inverse DFT we find that we can use the DFT multiplied by the appropriate constants to approximate the FT. The following equation falls out from this relationship,

\[ \Delta k = \frac{2\pi}{N \Delta x}. \quad (5.23) \]

This needs to be extended for a two dimensional surface, so using the appropriate constants and the DFT we have

\[ Z(m \Delta k_x, n \Delta k_y) \approx Z[m, n] \Delta x \Delta y, \quad (5.24) \]
\[ z(m \Delta x, n \Delta y) \approx \frac{NM}{(2\pi)^2} z[n] \Delta k_x \Delta k_y. \quad (5.25) \]
Now that we can calculate approximations of the FT and inverse FT, we have all the tools needed
to create a model for a random rough surface generator.

5.3.2 Verifying Model Performance

The model was tested by comparing its performance to two analytical solutions. The first of
which was an analytical solution for a flat PEC plane and the second was for a rough surface. The
following sections show the simulated radar cross section (RCS) for these type of configurations
compared to the analytical solutions.

Flat PEC Surface

To test the model for a flat PEC surface in the x-y plane, a plane wave was incident on the
surface with the wave vector in the x-z plane, and the backscatter was measured from all angles
above the surface in the x-z plane. The backscatter from the model was compared to the analytical
physical optics approximation for a flat PEC surface at a few different incident wave angles, a
couple of which are shown in Figures 5.3 and 5.4. [29, p. 598]

We compared the RCS results from the numerical and analytical methods. From the phys-
ical optics model described above we get a scattered far-field, $E_s$, which is generated from an
incident electric field, $E_i$. We can calculate the RCS from the scattered field using [31, p. 95],

$$\sigma(\theta^s, \phi^s; \theta^i, \phi^i) = \lim_{r \to \infty} 4\pi r^2 \frac{|\vec{E}_s|^2}{|\vec{E}_i|^2}. \quad (5.26)$$

As can be seen in Figures 5.3 and 5.4, the numerical result matches the analytical quite
well, particularly near the specular reflection, where the strongest backscattered signal is found.
The results shown are for normal incidence and for an incident angle at 45°.

Rough Surface Scattering Test

To test the physical optics model for rough surface scattering, it will be compared with
an analytical result from geometrical optics limit [30, p. 421]. This can be derived by taking the
radiated field in the physical optics approximation and evaluating it analytically as the wavelength
Figure 5.3: RCS of PEC plane with a TE polarized incident plane wave at $\theta = 45^\circ$.

Figure 5.4: RCS of PEC plane with an incident plane wave at normal incidence, $\theta = 0^\circ$, and with TE polarization.

approaches zero. This is why it is called the geometrical optics limit, because in geometrical optics we assume a vanishing wavelength or infinite frequency.

In using the geometrical optics limit the integrated slope variance of the surface spectrum is needed. This can be shown to be the Fourier transform of the second derivative of the autocorrelation function of the surface. To make sure that the calculations of this value were correct in the two dimensional case, an analytical solution for a one dimensional rough surface is compared
with a method of moments (MOM) solution [31, p. 151]. Monte Carlo averaging was used for the MOM solution. The ripples occur because this is an average of several simulated realizations of a random process and is for a surface of limited size. The larger the surface and the more realizations that are averaged, the closer the simulations agree with the analytical results.

![Monostatic Scattering Width](image)

Figure 5.5: MOM compared to the 1D geometrical optics limit of the physical optics approximation for a surface height variance of $\lambda/5$. This demonstrates the accuracy of our calculated value of the integrated slope variance, $S_c^2$.

We calculate this one dimensional solution for the scattering width with the following equation,

$$\sigma_\theta = \frac{k}{\cos^2 \theta} \sqrt{\frac{\pi}{2k \cos^2 \theta S_c^2}} e^{-\frac{\sin \theta}{4k \cos^2 \theta S_c^2}},$$

(5.27)

where $S_c^2$ is the integrated slope variance. The autocorrelation function can be calculated from the integrated slope variance. In one dimension the integrated slope variance is

$$S_c^2 = \int_{-\infty}^{\infty} k^2 S(k) \, dk,$$

(5.28)

where $S(k)$ is the one dimensional surface height PSD, and $k$ is the one dimensional spatial frequency.

As can be seen in Figure 5.5, the analytical method and the averaged MOM match well. From this we verify that our numerical calculation of the integrated slope variance is accurate.
Next we calculate the backscattering from a two dimensional rough surface. We use the geometrical optics backscattering coefficient equation [30, p. 549],

$$\sigma_o = \frac{1}{\cos^4(\theta)} S_c^2 e^{-\tan^2(\theta) \frac{S_c}{S^2}}, \quad (5.29)$$

to calculate the RCS of the surface. The angle $\theta$ in this equation represents the angle from horizontal at which the backscattering coefficient is calculated. The integrated slope variance,

$$S_c^2 = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} k^2 S(k_x, k_y) dk_x dk_y, \quad (5.30)$$

is now for a two dimensional surface, where

$$k = \sqrt{k_x^2 + k_y^2}, \quad (5.31)$$

is the wavenumber for the surface in the spatial Fourier domain and $S(k_x, k_y)$ is the PSD of the random process, $z(x, y)$, representing the height of the surface. Now that we can calculate the integrated slope variance for the backscattering coefficient, we can compare the geometrical optics limit approximation with the simulated results of the RCS from the physical optics model.

We calculate the RCS both analytically and numerically while sweeping over the incident angle. This is done for several surface height variances of which two are shown in Figures 5.6 and 5.7. There is a variance around the analytical results from the physical optics model. Since the analytical method is based on random process behavior of the surface they will not match perfectly. With more averages and a larger surface the simulated and analytic results match more closely. Therefore the variance is to be expected and we can say that the physical optics model matches the analytical method satisfactorily.

### 5.4 Performance of MIMO Radar

The physical optics model is now ready to be used to test the MIMO radar. We simulated a MIMO system with 4 transmit and 4 receive antennas. The four receive antennas are in a line.
orthogonal to direction of travel, and are separated by half a wavelength. The transmit antennas are set up in the same way. The simulated flight path is a straight line for about 20 meters. The UAV is simulated to be at an elevation of 50 meters. We computed the backscattered signal at 6000 different positions at 3.3 mm increments. The beam pattern of the antenna is a sinc pattern where it falls off to zero from the first beam around 10 meters.

The surface simulates a mountainous terrain with large variations in surface height. This was accomplished by placing paraboloids with random sizes throughout the surface, and adding a rough surface, using the rough surface generator, to the total terrain. The range of elevation
included in this terrain is from -10 meters to 10 meters. An example of a portion of a mountainous surface is shown in Figure 5.8.

Figure 5.8: A randomly generated 2-dimensional surface, with elevations between -10 to 10 meters.

5.4.1 SNR for MIMO Radar

SNR was defined previously for a single signal. Here for a MIMO signal we have a channel matrix which has $N_{el}$ elements where

$$N_{el} = MN.$$  \hfill (5.32)

This multi channel signal necessitates us to use an average SNR. For our model we will use an average SNR for the MIMO system defined as,

$$\text{SNR}_{\text{MIMO}} = \frac{|H|^2_{frw}}{\sigma_w N_{el}},$$  \hfill (5.33)

where $\sigma_w$ is the standard deviation of the white noise.

For the model we input a desired SNR and then computed the noise variance, $\sigma_w$. With this standard deviation for the white noise we randomly generate noise to add to our signals in the channel matrices.
5.4.2 MIMO Radar Simulation Results

Using the physical optics model a channel matrix at all 6000 positions was obtained. A window size of 1000 channel matrices was used for the correlation matrices, which means that the correlation path size was about 3.3 m long. With the correlation matrices we measured how well the algorithm could determine the position of the UAV. An example of how the radar worked for various SNRs is shown in Figure 5.9. This is a plot of the correlation matrix distance between the correlation matrix just obtained and the correlation matrices at each of the positions along the path. It demonstrates that, for at least the position that it was simulated at, a 10 dB SNR gives sufficient data for position discrimination. Observe that at the point where position recognition occurs there is a wide gap. This is because we are comparing the correlation matrices instead of the channel matrices for position identification. Since correlation matrices are made up of a series of channel matrices, correlation matrices that are located closely are formed with many of the same channel matrices. This causes correlation matrices that are closer together on the same path to have a smaller \( e_p \), which is the correlation matrix distance introduced in the beginning of the chapter.

Since not all positions are accurately predicted, we need to measure the degree of accuracy for the predicted position. We take the absolute value of the difference of the predicted and actual

![Figure 5.9: This plot shows the correlation matrix distance \( e_p \) between the correlation matrix at the current position and all the stored correlation matrices corresponding to different positions in the map.](image)
Table 5.1: Average error between estimated position and actual position for various SNRs.

<table>
<thead>
<tr>
<th>SNR (dB)</th>
<th>Average Error (meters)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>.28</td>
</tr>
<tr>
<td>15</td>
<td>.12</td>
</tr>
<tr>
<td>20</td>
<td>.02</td>
</tr>
<tr>
<td>25</td>
<td>.01</td>
</tr>
</tbody>
</table>

positions, and this will be the error. Figure 5.10 shows the error in position discrimination as the UAV flies along the designated path. It has an average error of 0.16 meters and was measured for an SNR of 10 dB. We find this measurement for the same radar data for a given path and generated terrain for various SNR’s. We compare the algorithms ability to estimate its position for different SNR performances in Table 5.1.

Figure 5.10: A plot of the error between the predicted position and the actual. The error is generally pretty low, however at some positions the error is large.

It is important to note that the SNR used in these simulations is an average SNR. The SNR is not constant for the UAV in its different positions. This is because where the surface is close to the aircraft the return signal will be much stronger than when the surface is farther away. Since the SNR is composed from an average of the signal that is received from each antenna at each of the
6000 positions, the instantaneous SNR will vary with the position. A decrease in accuracy still can be seen for a decrease in SNR, however as we would expect.

The MIMO radar simulated in the physical optics model demonstrated that it can estimate its position fairly accurately when the UAV is flying along the same path. A UAV often may need to fly along a different path than the path used to gather the stored channel matrices. To test this the UAV was simulated to fly perpendicular to the paths that were flown when gathering the data. The simulation was over a 2 by 2 meter square region, where there were 100 samples in each dimension, making a 100 by 100 grid of channel matrices. The channel matrices were gathered while the UAV was simulated to fly in the y direction over all the points in the 2 by 2 meter region. The UAV was then simulated to fly over the region in the x direction, orthogonal to the orientation of the UAV when the channel matrices were gathered for the mapping of the 2 by 2 meter region.

The MIMO algorithm was then used to determine if the UAV could predict its position as it flew along this path over the region. It took the channel matrices over the window size, and computed correlation matrices. These correlation matrices were then compared to the correlation matrices formed by channel matrices at each of the points on the 100 by 100 grid. Whatever position corresponded to the stored correlation matrix that minimized the correlation matrix distance was the estimated position of the UAV. Various window sizes were simulated from 0.2 to 3.0 meters.

From the simulated data the UAV was not able to identify its position. The error results are shown in Figure 5.11. The error in this plot shows that the predicted position is not accurate, especially for being in a region of 2 by 2 meters. A plot of the correlation matrix distance for for each position is shown in Figure 5.12. It shows the minimum, maximum and average correlation matrix distance at each of the position. It also shows the correlation matrix distance for the correlation matrix at the position of the UAV. This shows that the actual positions correlation matrix distance is around the average correlation matrix distance for each position. This means that the distance is not even a local minimum, and so it seems that there really is no similar information between the correlation matrices centered at the same spot but generated from different paths. This is probably due to the fact that the correlation matrix is composed of channel matrices gathered along the flight path. If these flight paths are orthogonal then there are only a few similar channel matrices used to form the same correlation matrix.
The antenna arrays were perpendicular to the flight path, therefore they are perpendicular to each other when the flight paths are perpendicular. Further simulation and testing needs to be done to figure out better ways to both gather the channel matrices, such as alternative ways to place antennas in arrays, as well as to use the data in a position recognition algorithm.

![Error vs Position](image1)

**Figure 5.11:** Prediction error for the UAV flying perpendicular to paths flown to gather data.

![Correlation Matrix Distances](image2)

**Figure 5.12:** This plot shows the correlation matrix distance $e_p$ between the correlation matrix at the current position and all the stored correlation matrices.
5.5 Summary

A physical optics model was developed and tested for accuracy. It was able to test and give insight to the performance of the MIMO radar. It was simulated to be able to tell its position when flying along a path that it flew before. However, when flying orthogonally to a previous path the algorithm used does not provide an accurate prediction of position. Further simulations need to be performed to more fully understand the performance of the MIMO radar. One of the challenges with this model was the length of time it took to simulate. To simulate the performance of the radar for 6000 positions it took three days. In order to more fully test the radar, methods to speed up the model need to be explored. Furthermore the physical optics scattering model takes too long to try a sweep of frequencies. This limits radar technologies that can be tested with the model.
CHAPTER 6. CONCLUSION

Beam steering for an earth based system for SatCom has been demonstrated for a low cost receiver. The degree to which the beam was able to steer and still receive the signal was wide enough given the constraint of orbital slots for satellites in the Geostationary orbit. There were weaknesses in the system as well. The SNR performance was much lower than it needed to be, to be considered practical for the SatCom industry. The ability of the system to steer rapidly was also not fast enough for certain tracking conditions such as in a windy environment. However by demonstrating beam tracking the results of this research shows the utility that ESAFs potentially have in industry. Many of the misalignment issues with SatCom could be softened with this technology, allowing for more reliable communication.

The eleven arrangement of the dipole over EBG antenna demonstrated a low profile dual CP antenna that simultaneously achieved quality dual CP performance and high isolation between the two ports. This makes this antenna a promising candidate for future SatCom dish feeds. Before it can become practical however a design that achieves the gain necessary for a dish feed while not compromising performance elsewhere is needed. This antenna did demonstrate the ability of compact low profile dual CP antennas to achieve high isolation between ports, which was an open research problem. This antenna is therefore shown to be promising candidate to replace current commercial horn feeds. It would reduce the cost, weight and size of dual CP receiver antennas used in industry. This antenna could also be used in an ESAF design due to its compact nature, to be used to electronically track CP signals.

While the microstrip cross antenna demonstrated quality CP as well as high gain, its large side lobe in the beam pattern and low isolation between ports are large challenges that need to be addressed before the design can be pursued as a valid option for a reflector feed.

The MIMO radar demonstrated localization capabilities when flying along a path that was previously mapped, when simulated in the physical optics model, by using correlation matrix dis-
stances. However when the stored data was gathered when the UAV was flying orthogonal to its active flight path the algorithm does not work. This needs to be explored further to fully understand the capabilities of the radar discussed. For example whether the radar algorithm is dependent on path or position needs to be more completely explored. The radar developed does demonstrate that a self contained radar can be used to gather information about the position of the UAV, but to understand its full potential further simulations are needed to fully test the MIMO radar.

The physical optics model was designed and verified to perform properly by using analytical models for a PEC plane and rough surfaces. It also demonstrated its ability to be used as a tool in analysis of radar performance. This tool can be developed further and used to test various radar technologies in the future.

6.1 Future Work

The major issue with the ESAF was the SNR performance. An investigation for the cause of the poor SNR performance needs to be done. This includes testing the impedance matching between components and the integrated performance of the various components. The filter could be miniaturized, and its performance improved. The system could also be made more compact, by integrating the filter and the power detector on the same board as the antennas, LNAs, and VGAs. This would require power detector and filter designs that would be compact enough to put on one board. In order to make the system more practical it also needs to have full steering capabilities by adding another dimension to the antenna array. The steering speed also needs to be increased. This requires a more complete investigation into a power detector that can more rapidly predict the power of the incoming signal.

The dipole over EBG antenna needs to be optimized to achieve the gain needed for a dish feed. This could be done by various array layouts. To do this a study of the effect on isolation and polarization performance of placing the dipole in various positions over the EBG needs to be studied. Also, making the design planar by placing microstrip dipole antenna over an EBG surface needs to be investigated. This would be done so that the design could be made by common PCB manufacturing processes. The multilayer design with a microstrip feed could be fabricated and tested, to verify simulation results and demonstrate the performance of the design.
More simulation tests can be done using the current model. For example determining if the radar can determine the UAV’s position when the flight paths cross between 0 and 90 degrees. The physical optics model could be optimized to increase simulation speed. This could be done using fast physical optics algorithms, vectorizing the code further, or building the model in something other than matlab. The model would also be improved by adding simulation over a band of frequencies. Once the model is faster, the performance of the radar could be tested more fully. Different antenna patterns and antenna array arrangements could be simulated to study their effect on localization abilities.

One major restriction for this method of localization is that the correlation matrices for each position in a given region need to be obtained previously. This could possibly be done by creating a radar modeler that would allow certain maps of terrains to be imported and used to generate the correlation matrices, but this would depend on how accurate the model was for the given map.

By using other instruments that are common on UAVs, inertial and elevation information can be utilized, in addition to the correlation matrices obtained, to take a path into an unmapped region and follow the same path out. For example the speed of the UAV could be used to determine what the possible positions are for the UAV after a certain time of travel. Elevation could also be used so that the UAV flies at the same elevation that was used to gather the stored data.
REFERENCES


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