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MAGNETO DIELECTRIC MATERIALS AND THEIR USE IN THE
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MAGNETO DIELECTRIC MATERIALS AND THEIR USE IN THE
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I began my study abroad journey seven years ago by going to Armenia for high school. After graduating in May 2019, I came to the United States to continue my higher education. I earned my undergraduate degree in May 2023, graduating with a major in Electrical Engineering, a minor in Computer Science, and a certificate in Engineering Leadership. Now, a little over a year later, here I am in December 2024, graduating once again with a master's in Electrical and Computer Engineering.

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Abstract

Historically, the reliance on materials with a relative permeability of one has limited the potential for antenna miniaturization, leaving magneto-dielectric materials underexplored due to their lack of commercial availability. This limitation becomes critical in applications like NASA's EcoSAR airborne radar, where the large and heavy antenna array, weighing approximately 200 lbs, significantly increases operational costs and limits the advancement of smaller, more efficient aircraft. Addressing this, magneto-dielectric substrates such as MAGTREX 555, with a relative permeability of 6, offer a promising solution by enabling miniaturization and improving antenna performance.

This thesis investigates the capabilities of MAGTREX 555 in designing and fabricating a stacked microstrip patch antenna. The study explores the trade-offs between miniaturization, bandwidth, and radiation efficiency. A $6'' \times 6'' \times 0.12''$ antenna was designed to operate at the EcoSAR radar's target frequency of 435 MHz, achieving a 9.2% bandwidth (40.14 MHz) and cross-polarization isolation greater than 30 dB. While the realized gain was limited to 11.5 dBi due to substrate losses, the design demonstrated a weight reduction of 66.85%, bringing the array weight down to 66.3 lbs from the original 200 lbs. Additionally, the study examines the impact of substrate thickness, coupling coefficients, and ground plane dimensions on antenna performance, showing that increasing the ground plane size from $6'' \times 6''$ to $11'' \times 11''$ improved directivity and realized gain from 3.4 dBi and -6.8 dBi to 5.7 dBi and -1.6 dBi, respectively.

The findings illustrate that while MAGTREX facilitates substantial size and weight

reductions, its limitations in bandwidth and efficiency warrant further optimization. Nevertheless, this research underscores the potential of magneto-dielectric materials in revolutionizing radar systems for airborne and spaceborne applications, paving the way for lighter, more cost-effective, and versatile designs.

Chapter 1

Introduction

1.1 Overview

Antennas have been considered fundamental components for communication systems, enabling long-distance transmission of data, signals, and information. From everyday mobile devices to advanced satellite communication, antennas are the critical link that allows electromagnetic waves to propagate through space, which helps transmit information across vast distances. With the advancement of space exploration in the last decade, the importance of highly efficient and miniaturized antennas is becoming more significant. For agencies like NASA, antennas do not only facilitate communication, but also serve as essential tools in planetary science missions.

One prominent project that underscores the importance of antennas in space exploration is NASA's EcoSAR mission [1]. EcoSAR, a synthetic aperture radar system, is designed to make polarimetric and interferometric measurements of Earth's surface. However, the large size and weight of the antenna system required for such a mission pose a significant challenge, especially for space-bound instruments where weight is directly linked to launch costs.

With NASA's ambition to eventually deploy EcoSAR-like instruments in space, the need to miniaturize these systems has become a priority. Reducing the size of these

antennas without compromising their performance is key to making space missions cost-effective.

As space missions push the boundaries of technology, the motivation to miniaturize antenna systems like EcoSAR becomes evident, laying the foundation for innovative advancement in antenna design.

1.2 Motivation

The increasing demand for smaller, lighter, and more efficient antennas is primarily driven by the need to reduce launching costs paving the way for more efficient space exploration. For NASA's EcoSAR system, which operates in P-band frequency range (300 MHz to 1 GHz), with a center frequency of 435 MHz and four polarization modes, the large, bulky antenna arrays used for airborne missions pose a significant hurdle for future spacecraft deployment. Then, as space exploration expands to include more planetary missions, such as mapping Mars' surface, miniaturizing the EcoSAR antenna becomes essential. Current antennas weigh hundreds of pounds, creating prohibitive costs for space launches, where every pound translates into thousands of dollars.

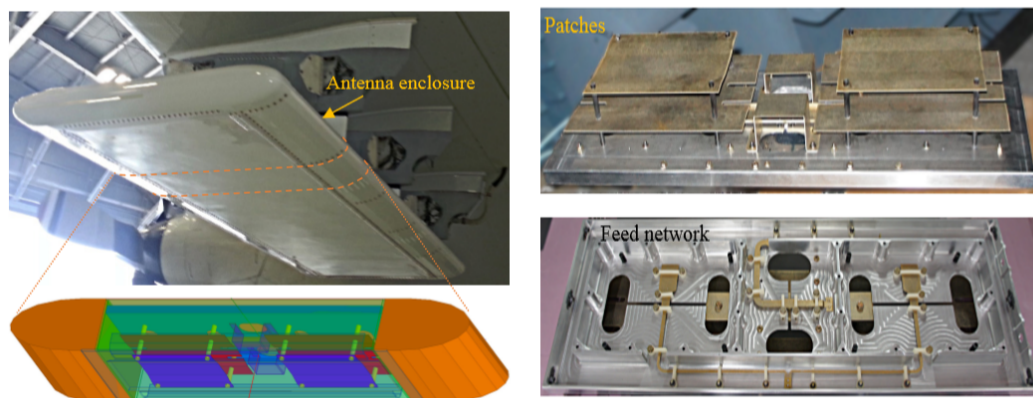


Figure 1.1: EcoSAR enclosure (left) and its stack-up (right) from [1]

By looking at figure 1.1, it can be noticed that these antennas are made of huge copper patches and, thus, are very heavy. Miniaturizing devices like this while keeping the same capabilities has been highly important in research. One of the objectives of EcoSAR is to be sent into space eventually. However, the cost for NASA to send a pound into space using its Space Launch System (SLS) is around \$58,000 per pound due to the high costs associated with the SLS program, designed for deep space missions like Artemis [2]. In addition to reducing costs, the ability to miniaturize antennas is vital for ensuring the flexibility and adaptability of radar systems used in various environments, whether in Earth's atmosphere or deep space. As NASA continues to push the boundaries of exploration, developing compact, high-performance antennas is critical for enabling the next generation of radar instruments that will map planetary surfaces, monitor environmental changes, and contribute to critical scientific discoveries.

The motivation behind this thesis is not only to solve a specific technical challenge but also to contribute to a broader vision of making space exploration more efficient and sustainable through advanced antenna technology. By addressing the miniaturization of antennas like EcoSAR, this research aims to pave the way for future radar systems that are lighter, cheaper, and more capable of supporting NASA's ambitious space missions.

1.3 Problem Statement

Considering the critical importance of minimizing antenna size and weight for space missions, the motivation driving this research is evident.. The challenge now lies in defining the specific problem that must be solved to achieve these goals. It has been established through research that a potential solution to reducing antenna size is to use magneto-dielectric materials [3]. While the use of magneto-dielectric materials presents a promising path toward miniaturization, it is necessary to determine how these materials

can be applied without a major compromise in performance.

To meet this challenge, this thesis focuses on investigating the potential of magneto-dielectric materials. To investigate such materials, a high-impedance laminate provided by ROGERS Corporation is used which is MAGTREX 555. This material helps reduce the size of the microstrip patch antennas (MPAs) used in EcoSAR due to its material properties. MAGTREX is a material known for its unique magnetic and dielectric properties, which provide the potential to create smaller antennas while minimizing the compromise in the antenna's performance in terms of bandwidth, radiation patterns, and polarization isolation [4].

This research aims to design and analyze a stacked microstrip patch antenna that operates at the required resonance frequency of 435 MHz, while adhering to most of NASA's specifications. These specifications include dual polarization with a vertical polarization bandwidth of up to 200 MHz and a horizontal polarization bandwidth of up to 120 MHz, as well as cross-polarization isolation greater than 30 dB. Additionally, NASA's EcoSAR has a gain of 19 dBi and provides a $\pm 35^\circ$ scan angle achieved through 10 subarrays comprising a total of 20 elements, with a total weight of 200 lbs[1]. Figure 4.33 shows the current configuration of the EcoSAR antenna.

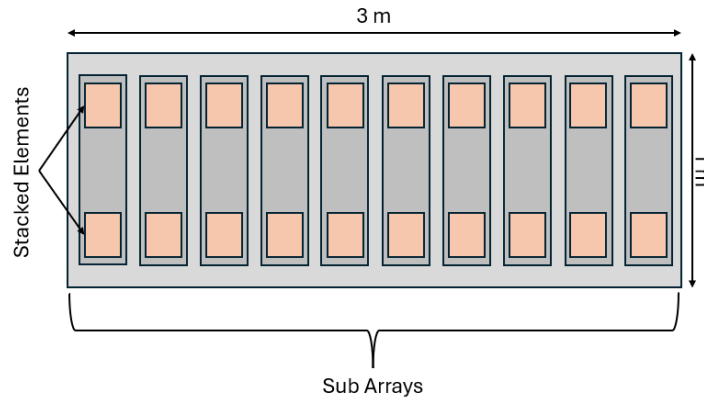


Figure 1.2: Layout of NASA's EcoSAR antenna showing 10 sub-arrays and 20 total elements

The core problem addressed in this thesis is how to maintain the necessary performance parameters of the EcoSAR antenna while minimizing its physical dimensions. By utilizing ROGER's MAGTREX material, this research explores the limitations and benefits of using magneto-dielectrics for antenna miniaturization, contributing to the broader goal of making advanced radar systems more viable for space deployment. The successful miniaturization of these systems will have wide-reaching applications beyond space exploration, potentially influencing fields such as remote sensing, environmental monitoring, and airborne surveillance.

1.4 Outline

With the articulated problem in hand, the structure of this thesis is designed to systematically address the challenges of antenna miniaturization using magneto-dielectric materials. The thesis begins with a detailed background on antenna types and key parameters essential for understanding the challenges addressed in this research in chapter 2. Chapter 3 delves into microstrip patch antennas (MPAs), their functionality, and previous work in stacked antennas, laying the groundwork for the design phase. Chapter 4 focuses on the design process of a stacked microstrip patch antenna using MAGTREX, discussing the initial mathematical formulations and subsequent optimization through simulation tools like Ansys HFSS and MATLAB. The fabrication of the antenna is explained in Chapter 5, detailing each step from material selection to the construction of the prototype. Chapter 6 presents the measured results of the fabricated antennas' performance, comparing them with the simulated models and evaluating the extent to which NASA's specifications are met. Finally, Chapter 7 concludes the thesis by summarizing the findings and suggesting directions for future research in antenna miniaturization for airborne applications.

Chapter 2

Antenna History and Background

Having outlined the motivation for this research, which focuses on the miniaturization of antennas using magneto-dielectric materials, it is crucial to first understand the broader historical and technical context of antenna technology. This foundation helps highlight the innovative aspects of this research and the challenges it aims to overcome.

2.1 Antenna History

Understanding the development of antenna technology is essential to addressing the challenges posed by modern applications like NASA's EcoSAR system. The quest to miniaturize antennas for space missions cannot be fully appreciated without first looking at how antenna technology has evolved. By tracing the history of antennas, a better grasp of the foundational principles that have shaped their current designs and performance can be achieved. These principles are critical to tackling the present-day problem of reducing antenna size without compromising performance.

The origins of antenna technology are rooted in the pioneering work of Michael Faraday, who, in the 1830s, discovered electromagnetic induction, establishing a foundation for understanding electromagnetic waves. This discovery paved the way for James Clerk Maxwell's 1865 theoretical framework, predicting electromagnetic wave

propagation. In 1887, Heinrich Hertz confirmed Maxwell's predictions by experimentally generating and receiving radio waves using rudimentary dipole and loop antennas. Hertz's breakthrough was pivotal, employing an antenna composed of two flat metallic plates attached to 30 cm rod [5], demonstrating the principles of electromagnetic wave transmission and reception which can be shown in figure 2.1 and is further illustrated in figure 2.2.

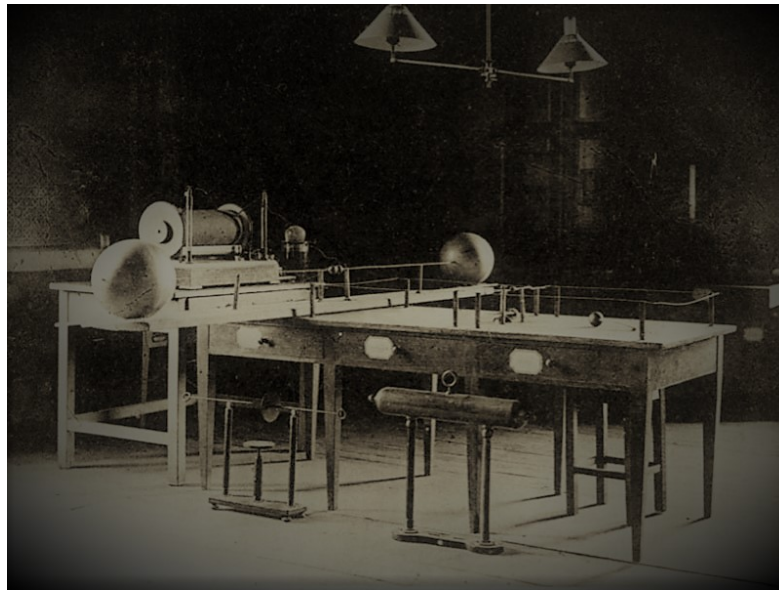


Figure 2.1: The experiment conducted by Heinrich Hertz [5]

As technology advanced, antennas became more diverse and specialized. Marconi's transatlantic communication in 1901 utilized vertical wire antennas, showcasing their potential for long-distance communication. The 20th century saw the introduction of designs like Yagi-Uda antennas in the 1920s, horn antennas in 1939, and microstrip patch antennas in the 1970s. These developments reflected a growing need for efficiency, directionality, and compactness in various applications, from radar to telecommunications [6].

Since then, antenna technology has progressed to meet various demands, especially in military and communication systems. From basic wire antennas to advanced arrays,

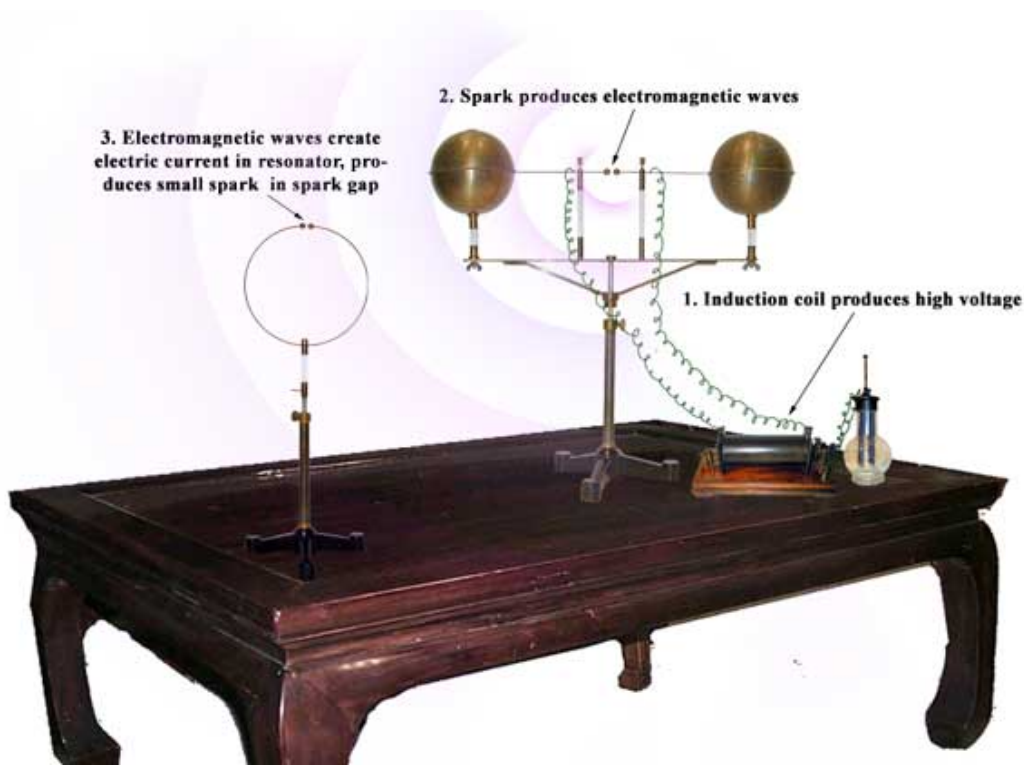


Figure 2.2: Illustrative figure of the experiment conducted by Heinrich Hertz [5]

their ability to transmit and receive signals has made them essential in modern society. Antennas, being passive and reciprocal devices, operate by radiating electromagnetic energy from a transmitter to a receiver, enabling the transfer of information across distance [7].

As antenna technology advanced, so did the variety of designs and types, each tailored to meet specific needs. From point-to-point communication to radar systems and beyond, the antenna's ability to transmit signals directionally and efficiently has been crucial. With this historical context in mind, it becomes clear that understanding antenna evolution is not merely academic; it forms the basis for tackling present-day challenges like miniaturization and optimization, especially for complex systems such as NASA's EcoSAR radar.

2.2 Antenna Performance Metrics

Building on the historical development of antenna technology, it is important to understand the core principles that dictate how antennas function upon which an assessment of their performance can be made. Although the evolution of antennas has led to a range of designs and innovations, the basic working principles remain crucial for any modern application.

At their core, antennas serve as transitional structures that enable communication between transmission lines and free space, converting guided electromagnetic waves into radiated signals and vice versa as shown in figure 2.3. The basic function of an antenna is straightforward: it receives signals from a source through a transmission line, converts them into electromagnetic waves, and transmits them into free space. The process is reversed for receiving antennas, where incoming electromagnetic waves are converted back into signals and passed through a transmission line to a receiving device. This simplicity, however, belies the complexity of antenna performance, which depends on several key parameters.

One of the most important parameters is the *radiation pattern*, $F(\theta, \phi)$, representing the radiation variation with the angle around the antenna while transmitting signals. It is quantified by noting a power density value, S , at a fixed distance, r , from the antenna. The radiation can either be *directive* as shown in figure 2.4, *omnidirectional* (uniform in one plane), or *shaped main beam* [7]. Radiation is a critical parameter for understanding antenna performance, as it directly influences the efficiency and effectiveness of electromagnetic wave propagation.

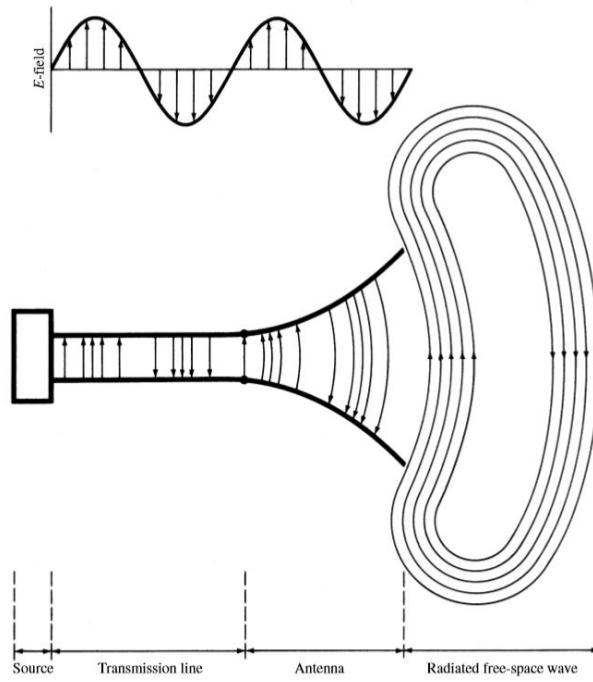


Figure 2.3: Path of propagation of electromagnetic waves from source to free space using an antenna from [8].

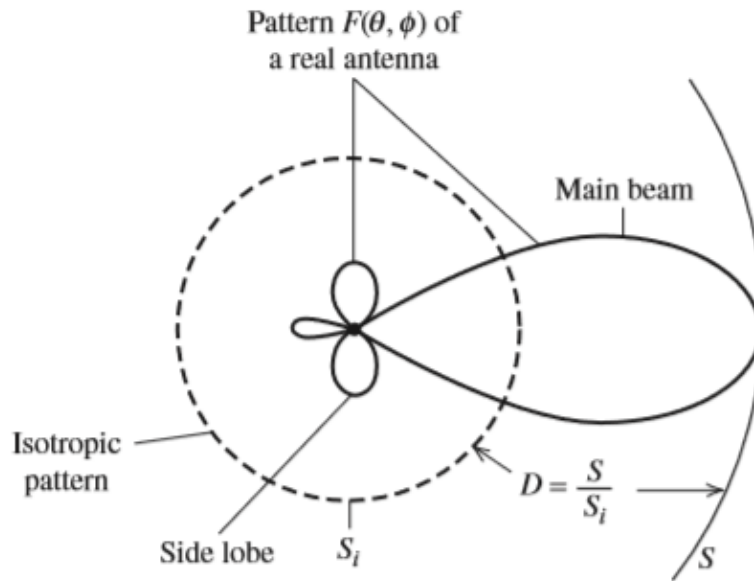


Figure 2.4: Illustration of radiation pattern and directivity from [7].

Radiation Mechanism

To understand how radiation works, a look at single-element antennas shows that radiation is emitted due to the acceleration of electric charges within the antenna structure. When an alternating current is applied to the antenna, it causes the electrons within the antenna to accelerate back and forth, creating changing electric and magnetic fields around the antenna. These fields propagate outward as electromagnetic waves. While monopoles and dipoles are examples of single-element antennas with wide radiation patterns and low directivity, this is not universal for all single-element antennas. Aperture antennas, such as horn antennas and parabolic reflectors, demonstrate that single-element designs can achieve high directivity depending on their structure and operating principle

Antennas with more than a single element, such as arrays, produce fields that add in phase at specific points in space, as shown in figure 2.5. The amount of radiation is determined by how the fields from different elements combine. Consider a loop of wire carrying current. The current distribution around the loop results in elements of current that oppose each other at opposite points of the loop. Let's assume that the current elements are represented by $d\vec{I}_1$ and $d\vec{I}_2$ which are separated by a distance D . Now consider a point A in space, where the distance from $d\vec{I}_1$ is R_1 , and the distance from $d\vec{I}_2$ is R_2 . The radiation from the loop will depend on how these current elements interact at point A , and how the resulting fields from each element combine based on their relative distances and phases.

If $R_2 - R_1 \geq 0.1\lambda \implies D \leq 0.1\lambda$, the current elements add out of phase, producing small radiation. However, if $R_2 - R_1 \leq 0.1\lambda \implies D \geq 0.1\lambda$, then the current elements add in phase, causing an increase in radiation and hence, directivity compared to the single element [9].

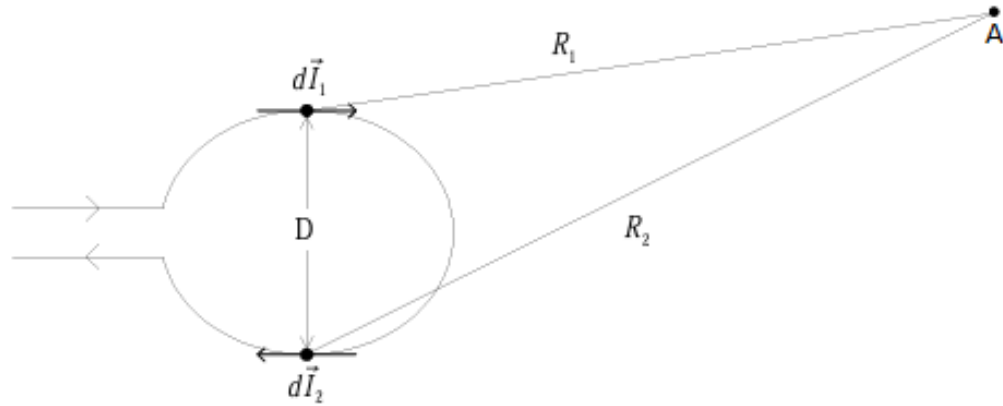


Figure 2.5: Field at a location from two current elements from [9]

Then, after the transmission of electromagnetic waves by antennas, these electromagnetic waves get intercepted by receiving antennas. The receiving antenna generates a small amount of current from these waves, varying depending on the signal strength. This current will deliver power to the antenna load. A typical example is the car's antenna, which picks up FM/AM (frequency-modulated/amplitude-modulated) signals from radio stations. The antenna captures radio waves and converts them into electrical signals in the form of current. This current is then processed by the radio, which includes demodulation to extract the transmitted information. The demodulated signal is subsequently amplified and transformed into an output, such as sound delivered through speakers or data for other applications like communication or broadcasting.

Radiated Power

Following that, to measure the intensity of the radiation, a characterization can be made to see how intense different regions are, and this value will be characterized by power called radiated power.

Radiated power is the power that leaves the vicinity of the source and never returns. The quantity used to describe this power is the instantaneous Poynting vector defined

as:

$$\vec{P} = \frac{1}{2} \left\{ \vec{E} \times \vec{H}^* \right\} \quad (W/m^2) \quad (2.1)$$

Where $\vec{E}$ and $\vec{H}$ are the electric field vector and the magnetic field vector, respectively.

By looking at (2.1), we can observe that the real part represents the average real power density, whereas the imaginary part can be assumed to represent the reactive stored energy associated with the electromagnetic field. It is also known that in the far-field region, the electromagnetic fields of the antenna are predominantly real and can be referred to as *radiation density* [8]. Now, (2.1) can be rewritten to obtain an equation for the radiated power:

$$P_{rad} = P_{av} = \oint\oint_s P \cdot ds = \oint\oint_s P \cdot \hat{n} da = \frac{1}{2} \oint\oint_s \text{Re} \left(\vec{E} \times \vec{H}^* \right) \cdot ds \quad (2.2)$$

This is the general case for radiation. However, the ideal case, known as the isotropic radiator, must be mentioned, although this radiator does not exist in practice.

Isotropic radiators are antennas that radiate equally in all directions. They are mentioned to provide a convenient isotropic reference with which other antennas can be compared. Due to its symmetric radiation, its Poynting vector will only have a radial component. This changes (2.2) according to the following:

$$P_{rad} = \oint_s P_0 \cdot ds = \int_0^{2\pi} \int_0^{2\pi} [\hat{a}_r P_0(r)] \cdot [\hat{a}_r r^2 \sin(\theta) d\theta d\phi] = 4\pi r^2 P_0 \quad (2.3)$$

Where P_0 is the power density that is uniformly distributed over the surface of a sphere of radius r [8] shown in (2.4).

$$P_0 = \hat{a}_r W_0 = \hat{a}_r \frac{P_{rad}}{4\pi r^2} \quad (W/m^2) \quad (2.4)$$

Far-Field Conditions and Field Regions

As electromagnetic waves radiate from the antenna, they go through different regions of which the far-field is the most important one when it comes to calculations. In the far-field of an antenna, the electromagnetic fields have $1/r$ magnetic dependence. To examine the point where the far field begins, figure 2.6 can be referred to.

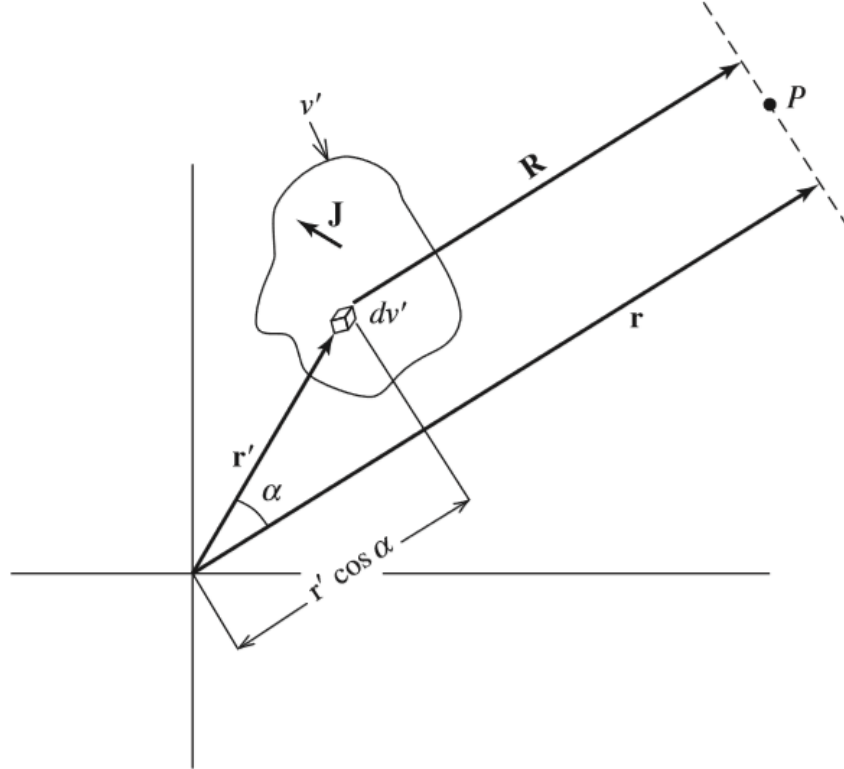


Figure 2.6: Parallel ray approximation for far-field calculations of a general source [7]

It can be seen that R can be determined by relating it to the phase difference by the following geometrical technique:

$$\vec{R} = r - r' \cos(\alpha) = r - r' \frac{\vec{r}' \cdot \vec{r}}{r r'} = r - \vec{r}' \cdot \vec{r} \quad (2.5)$$

Then, the distance where the far field begins is given by:

$$r_{ff} = \frac{2D^2}{\lambda} \quad (2.6)$$

The distance where the far field begins, r_{ff} , is taken to be that value of r for which the path length deviation due to neglecting higher orders of r is a sixteenth of a wavelength. D is the directivity representing the length of the line source that can differ depending on L and λ , L being the length of the line source. r_{ff} is the Rayleigh distance, the far-field distance; hence, the far-field region is at $r \geq r_{ff}$. The condition given by (2.6) will not stand if the antenna size is small compared to a wavelength as the distance is not sufficiently large enough and thus, additional considerations are needed as stated in [7].

Before reaching the far field, the radiation will be in the near field, divided into the reactive near field and the radiating near field. The reactive part is a region where the reactive fields dominate the radiation, whereas the radiating part is where the radiation fields dominate the reactive fields.

The ratio of radiated to reactive power density is $1/(\beta r)^3$. Figure 2.7 will be used to better understand and simplify the rest of the conditions.

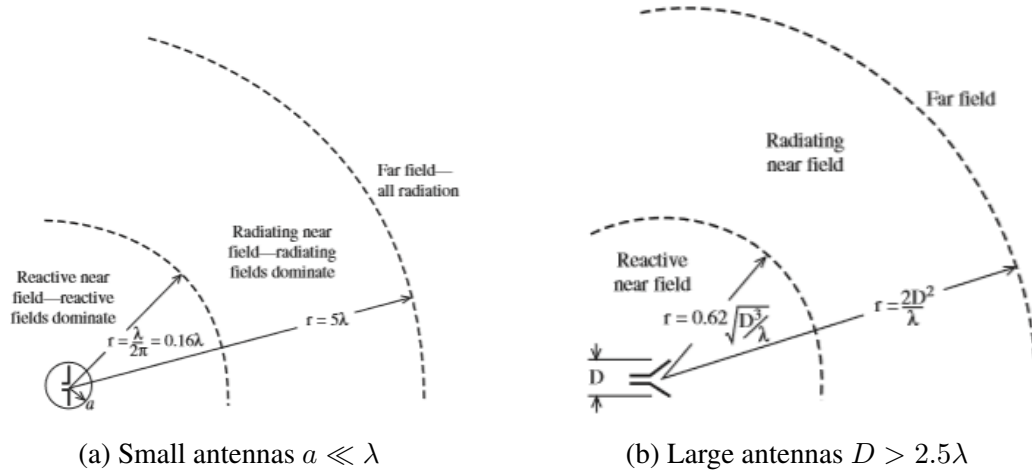


Figure 2.7: A figure that shows the conditions for the far-field region from [7]

From figure 2.7, there exist the following conditions for the far-field region:

$$r > \frac{2D^2}{\lambda} \quad (2.7a)$$

$$r > 5D \quad (2.7b)$$

$$r > 1.6\lambda \quad (2.7c)$$

The far-field distance criterion given by (2.7a) is the limiting condition for antennas operating in the UHF region and above. However, at lower frequencies where the antenna can be small compared to the operating wavelength, the far field distance will have to be greater than $2D^2/\lambda$ to satisfy all conditions in (2.7) [7].

To understand radiation better, *directivity*, D , can be mentioned and dove into as it expresses the ratio of power density in the direction of the pattern peak to the average power density at the same distance from the antenna. Figure 2.4 shows a further illustration of the radiation patterns by showing a visual comparison between $F(\theta, \phi)$ of a real antenna and that of an isotropic spatial distribution [7]. The isotropic pattern, represented by a dashed circle, depicts an ideal antenna radiating uniformly in all directions. In contrast, the real antenna's radiation pattern shows directional characteristics. The real antenna's pattern is characterized by a main beam, which represents the region of maximum radiation intensity and points in the desired direction of energy transmission. Surrounding the main beam are side lobes, which indicate smaller, unintended radiation in other directions. These side lobes are generally undesirable as they can lead to reduced efficiency and unwanted interference.

Gain, G , is another parameter that is very important to consider. It ties to *directivity* as both go hand in hand. The main difference between *gain* and *directivity* is that *gain*

considers the losses of the antenna [7]. Besides, *gain* measures how much input power is concentrated in a specific direction, whereas *directivity* measures how much power an antenna radiates in different directions. It is an important parameter to consider to help account for the efficiency of the antenna in addition to its directional capabilities. Gain is important for enhancing the quantification of the antenna's performance, as directivity is only controlled by the radiation pattern.

Gain is the ratio of intensity, in a given direction, to the radiation intensity that would be obtained if the power accepted by the antenna was radiated isotropically [8]. The reference antenna is often known as a lossless isotropic source with a calculable known gain. Turning this into an equation:

$$G(\theta, \phi) = 4\pi \frac{U(\theta, \phi)}{P_{in}} = e_{cd} \left[4\pi \frac{U(\theta, \phi)}{P_{rad}} \right] = e_{cd} D(\theta, \phi) \quad (\text{dimensionless}) \quad (2.8)$$

Where $U(\theta, \phi)$ is the radiation intensity, P_{in} is the total input (accepted) power, e_{cd} is the antenna radiation efficiency ($0 \leq e_{cd} = \frac{P}{P_{in}} \leq 1$), and $D(\theta, \phi)$ is the directivity of the antenna. If the dimension is not stated, the power gain is usually taken in the direction of maximum radiation [7].

Looking at (2.8), it can be noticed that a maximum value of the gain can be obtained by relating it to a maximum value of directivity, D_0 , through the following general equation:

$$G_0 = G(\theta, \phi) |_{max} = e_{cd} D(\theta, \phi) |_{max} = e_{cd} D_0 \quad (2.9)$$

A partial gain can also be defined for the antenna for a given polarization in a given direction by breaking (2.9) into the following:

$$G_0 = G_\theta + G_\phi \quad (2.10)$$

Where,

$$G_{\theta} = \frac{r\pi U_{\theta}}{P_{in}} \quad (2.11a)$$

$$G_{\phi} = \frac{r\pi U_{\phi}}{P_{in}} \quad (2.11b)$$

Where U_{θ} is the radiation intensity in a given direction contained in the E_{θ} field component, and U_{ϕ} is the radiation intensity in a given direction contained in the E_{ϕ} field component [8].

Some of these equations will be broken down further in this section to better understand how they work. The radiation intensity that is mentioned in (2.8) is the power radiated in a given direction per unit solid angle and has units of watts per square radian (steradian, sr) and is independent of distance [7].

$$U(\theta, \phi) = \frac{1}{2} \text{Re} \left(\vec{E} \times \vec{H}^* \right) \cdot r^2 \hat{r} = S(\theta, \phi) r^2 = U_m |F(\theta, \phi)|^2 \quad (2.12)$$

Where U_m is the maximum radiation intensity ($U_m = U(\theta_{max}, \phi_{max})$), $|F(\theta, \phi)|^2$ is the power pattern normalized to a maximum value of unity in the direction $(\theta_{max}, \phi_{max})$.

For non-isotropic sources, an average power per steradian can be defined as:

$$U_{avg} = \frac{1}{4\pi} \iint U(\theta, \phi) d\Omega = \frac{P}{4\pi} \quad (2.13)$$

Where P is the total power radiated. The average radiation intensity, in this case, U_{avg} , equals the radiation intensity, $U(\theta, \phi)$, that an isotropic source with the same input power, P , would radiate. Besides, (2.14) shows how directivity can be calculated.

$$D = \frac{U_m}{U_{avg}} = \frac{4\pi U_m}{P} = \frac{\frac{1}{2} \max \left[\text{Re} \left(\vec{E} \times \vec{H}^* \right) \cdot \hat{r} \right]}{P/4\pi r^2} = \frac{4\pi}{\Omega_A} \quad (2.14)$$

Where Ω_A is the beam solid angle and is given by:

$$\Omega_A = \iint_{sphere} |F(\theta, \phi)|^2 d\Omega = \frac{P}{U_m} \quad (2.15)$$

A comparison with the directivity of an isotropic source is needed to understand the value of the obtained directivity. The directivity of an isotropic source antenna is $D = 1$ [7]. If the obtained D is less than 1, it is said to be less directive than an isotropic source, indicating a bad radiator. Suppose it is greater than 1 in a certain direction. In that case, the antenna is directive in that direction and not in other directions, making it a suitable radiator for its designed purpose.

Another way to represent gain can be by (2.16) which takes efficiency into account.

$$G = e_p \cdot e_d \cdot e_c \cdot D \quad (2.16)$$

Where e_p is the polarization efficiency, e_d is the efficiency of the dielectric, and e_c is the conductor efficiency. While talking about gain, realized gain has to be mentioned as it is an important parameter for antenna engineers. It differs from that of gain as it takes the efficiency of the matching network, e_m , into account. This adds e_m to (2.16) as the new total efficiency is $e_T = e_m \cdot e_c \cdot e_d \cdot e_p$.

Another main parameter is *polarization*, which describes the vector nature of the electric fields radiating from the antenna. Antenna polarization can either be linear, circular, or elliptical. A dual-polarized antenna enables the doubling of communication capacity by carrying separate information on orthogonal polarizations over the same physical link on the same frequency [7]. This can be illustrated in figure 2.8 which shows the interaction between the electric field polarization of transmitted signals and the receiving antenna's configuration. It demonstrates how the polarization of the transmitted signal (horizontal, vertical, or circular) influences its reception by linearly polarized receiving antennas. The impact of polarization mismatch is particularly evident

in cases where the polarization of the transmitted signal does not align with that of the receiving antenna.

For instance, when a horizontally polarized signal is received by a vertically polarized antenna, the signal cannot be properly captured, resulting in signal degradation or loss. Similarly, circularly polarized signals are shown being transmitted and received by linearly polarized antennas. In these cases, a partial polarization mismatch occurs, leading to reduced signal strength, as the linearly polarized antenna only receives a portion of the circularly polarized wave's energy.

The figure emphasizes the importance of polarization alignment between the transmitting and receiving antennas to ensure optimal signal reception. Polarization mismatch can cause power losses proportional to the cosine square of the angle between the polarization vectors, as described by the polarization loss factor (PLF). This highlights the need for proper design and alignment in communication systems to minimize such mismatches and maintain efficient signal transmission. Besides polarization, there is cross-polarization (X-pol) which is the polarization orthogonal to the intended polarization, as antennas are never 100% polarized in a single mode.

Other parameters include *impedance*, Z_A , which is the input impedance at the antenna terminals, and *bandwidth*, BW , which is the range of frequencies over which essential performance parameters are acceptable [7].

These are the six main parameters, but other considerations go into the design process, which sometimes causes some parameters' optimization to be overridden. These considerations can be mechanical (size, weight, aerodynamics, vibration, positioning accuracy, etc...), environmental (wind, rain, temperature, altitude, etc...), and can just be appearance preference.

The mentioned considerations are general; however, special considerations exist for transmitting and receiving antennas separately. For the transmitting side, power han-

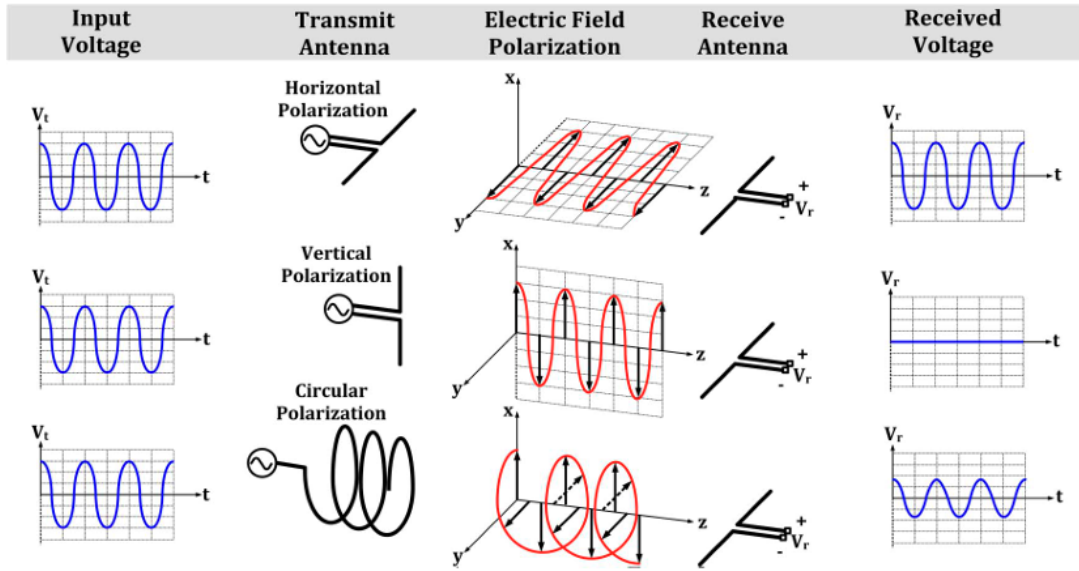


Figure 2.8: Illustration of how a signal is transmitted and received depending on electric field polarization and antenna configuration [10]

ding and radiation hazards have to be taken seriously during the design process, as malfunctions can cause the device to get damaged or fry up some of the connected devices. As for the receiving side, noise is usually the main issue, which, in this context, is the energy that is not the desired signal. This noise is called electromagnetic interference (EMI) or radio frequency interference (RFI). RFI is narrowband noise at specific frequencies, whereas EMI is random broadband noise [11].

2.3 Antenna Types

With the mentioned key parameters that define antenna performance, it becomes evident how these characteristics influence the design and functionality of various antennas. The choice of antenna type plays a crucial role in optimizing these parameters for specific applications, whether the goal is long-range communication, wideband operation, or miniaturization. To appreciate how different antenna types meet diverse performance

needs, it is essential to explore the various designs and configurations that have been developed. Each antenna type brings unique advantages and trade-offs, making it necessary to select the appropriate design based on the intended use.

2.3.1 Resonant Antennas

Resonant antennas are antennas whose length is $\sim$ half a wavelength. The antenna is purely resistive with zero reactance at resonance, meaning a maximum amount of current will flow through it. This current is directly related to the amplitude of the radiated electromagnetic fields. Although standing waves of current are mainly associated with wire antennas such as a dipole, they can also be realized by other resonant structures, such as microstrip patch antennas. In understanding resonant antennas, it is important to note that the radiation characteristics often stem from the combined effects of individual current elements within the structure. This concept can be extended to antenna arrays, where multiple radiating elements (each potentially resonant) combine their electromagnetic contributions to achieve specific radiation patterns. The underlying principle is that an array of current filaments generates electromagnetic components that combine coherently or destructively, depending on their phase relationship. While this section focuses on individual resonant antennas, arrays are introduced here as they leverage similar principles to enhance performance, such as increasing gain or controlling beam directionality. These antennas are chosen for their simplicity, good input impedance, and narrow frequency band. They typically exhibit low to moderate gain, real input impedance, and narrow bandwidth [7], which will be explored in chapter 3. The narrow bandwidth comes from the nature of the antenna, which relies on establishing a standing wave. Figure 2.9 shows examples of resonant antennas.

Half-wave dipoles are also known as a doublet or Hertz antennas, where the current

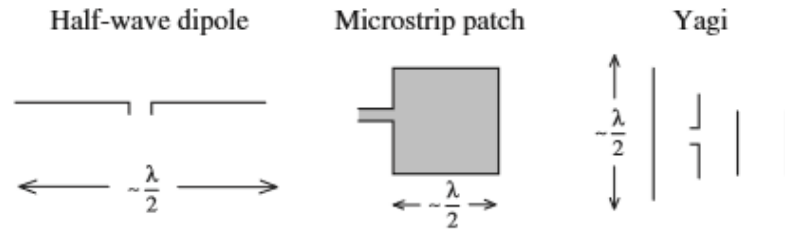


Figure 2.9: Examples of resonant antennas [7]

is maximum, and the voltage is minimum at the center of the antenna and vice versa at the edges. Although they can be used as standalone low-gain antennas, they can also be incorporated as driven elements in antenna designs such as the Yagi antenna. Besides half-wave dipoles, there exists a monopole that comes with a different configuration. Monopole antennas are a specific type of resonant antenna widely used for their simplicity and efficiency. These antennas are typically quarter-wavelength long, $\lambda/4$, and rely on a conductive ground plane to simulate the other half of a dipole antenna. This configuration effectively creates a radiation pattern similar to that of a half-wave dipole while using less physical space. At resonance, a monopole antenna exhibits zero reactance, ensuring its impedance is purely resistive and allowing for optimal power transfer. Its radiation pattern is omnidirectional in the horizontal plane, making it ideal for applications where uniform signal transmission or reception is required in all directions. The typical input impedance of a quarter-wave monopole is approximately 36.5 ohms, half that of a half-wave dipole.

Yagi antennas are more complicated than half-wave dipole antennas. They include a dipole, as stated earlier, but have better directivity. It focuses signal strength in one direction while rejecting signals from other directions by using a series of antenna elements, a driven element being the dipole and directors, to create a waveguide in which radio waves are amplified and re-radiated into the desired direction. Figure 2.10 shows the basic configuration of this antenna type. This antenna was designed by Yagi and Uda

who initially designed the antenna for shortwave communication, and their work was published in 1926. The concept gained international recognition after Yagi presented the design in English-language journals, leading to widespread adoption, particularly during World War II for radar and other communication systems. Today, Yagi antennas are widely used for television reception and other radio frequency applications [12].

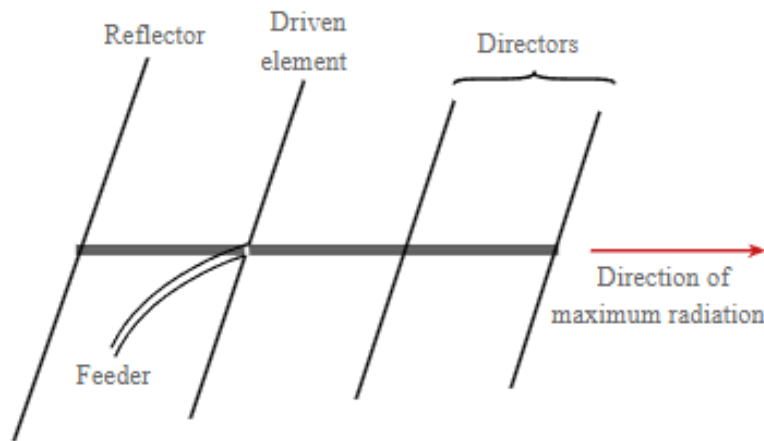


Figure 2.10: Basic concept of Yagi antenna showing reflector 5% longer and directors 5% shorter than the driven element

2.3.2 Electrically Small Antennas

Electrically small antennas are much shorter than a wavelength, $< \lambda/10$. An example is the vertical monopole used for AM reception in cars, approximately 0.003λ long, with an omnidirectional pattern in the horizontal plane. Despite the advantages of electrically small antennas, such as simplicity, they suffer from low input resistance and high input reactance, leading to significant ohmic losses on the structure [7]. Resistance represents the real part of impedance, while reactance corresponds to the imaginary part, which is influenced by the system's inductance and capacitance. Achieving the correct input impedance is crucial for ensuring efficient power transfer and effective radiation.

However, there are instances where this ideal impedance cannot be reached. In such cases, the primary objective becomes optimizing the real part of the impedance. Despite this, excessive reactance can hinder proper impedance matching.

In addition to the low input resistance and high input reactance, electrically small antennas exhibit low radiation efficiency and directivity. The most common structures used in this type of antenna are the short dipole and the small loop, shown in figure 2.11.

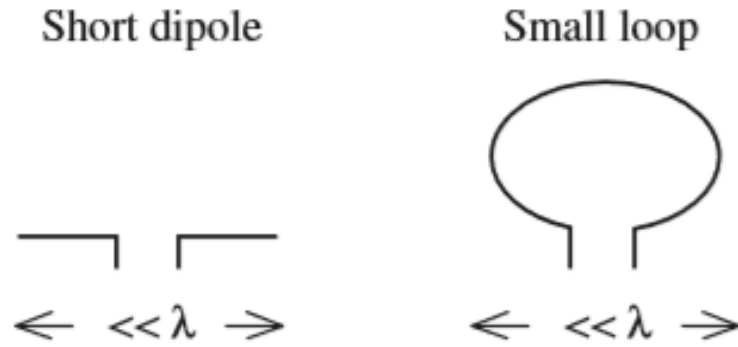


Figure 2.11: Examples of electrically small antennas showing a short dipole (left) and a small loop (right) [7]

Starting with the short dipole example, a $\lambda/10$ dipole with a 1 mm conductor diameter at 100 MHz has an impedance at the center feed point of $1.96 - j1758\Omega$ according to [13] using NEC2 numerical modeling. The high reactance and low resistance indicate a large impedance transformation is needed to match this antenna to a typical 50Ω system. Although this antenna is small compared to a half-wave dipole, the efficiency gets reduced by the matching system rather than the antenna itself, as the antenna's gain is only 0.37 dB less than that of a half-wave dipole.

As for the small loop, at a diameter of $\lambda/10$, the radiation resistance is calculated to be 1.92Ω according to [13]. However, The feed point impedance is affected by factors such as the resistive losses of the conductor and the inductance of the loop, which results in an impedance value of approximately $\sim 3 + j800\Omega$. As a consequence, the radiation

pattern and gain of the antenna will resemble that of a short dipole with a length around $\lambda/10$, characterized by a nearly uniform current distribution.

2.3.3 Aperture Antennas

Aperture antennas have an aperture at its end that can come in many forms, such as square, cone, rectangle, or circle. The aperture is the area around the antenna where power can be derived from electromagnetic field effects. In this type of antenna, the gain for a fixed wavelength is proportional to the effective aperture and inversely proportional to the square of the wavelength. Aperture antennas have the unique feature of propagating electromagnetic waves through their openings. The aperture is usually several wavelengths long in one or more dimensions. They have higher radiation than the two-wire transmission line, omnidirectional radiation, better performance, and are simple to make. On the other hand, their VSWR is higher. VSWR is the voltage standing wave ratio that measures how efficiently radio frequency gets transmitted from the power source to the load. It is calculated as the ratio between the maximum and minimum amplitudes of the corresponding field components on a transmission line. It is often expressed as a ratio with 1:1 being the perfect impedance match [7].

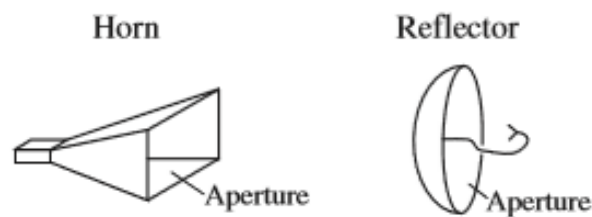


Figure 2.12: Examples of aperture antennas showing a horn antenna (left) and a reflector (right) [7]

One of the main types of aperture antennas is the horn antenna shown in figure 2.12. Horn antennas have a directional radiation pattern with a high antenna gain that can

get to 25 dBi in some cases. They have wide impedance bandwidth, meaning the input impedance slowly varies over a wide frequency range. The directivity is roughly equal to the gain due to having very little loss. The horn antenna is typically fed by a section of a waveguide that is fed with a short dipole. Waveguides are hollow metal cavities that are used to guide electromagnetic energy from one place to another [8].

Another heavily used aperture antenna is the parabolic reflector antenna from figure 2.12, commonly known as the satellite dish antenna. They have very high gain and low cross-polarization. The reflector antenna consists of a feed antenna pointed towards a parabolic reflector, with the feed antenna often being a horn antenna with a circular aperture. The reflecting dish must be at least several wavelengths in diameter, with the diameter reaching an order of 100 wavelengths for very high-gain dishes [14].

2.3.4 Broadband Antennas

Broadband antennas are the types that are designed to operate across a wide range of frequencies. They are characterized by their ability to transmit and receive signals over a broad spectrum of frequencies without significantly losing performance or efficiency. They are defined by angles instead of lengths, meaning that they are frequency-independent, as angles are independent of wavelength for any frequency. They come in various designs tailored to different applications, such as circularly polarized broadband antennas like the spiral antenna in figure 2.13 [7].

Spiral antennas are a famous type of broadband antenna due to their substantial bandwidth, as the fractional bandwidth can be as high as 30:1. If the lower frequency is 1 GHz, the antenna can be efficient up to 30 GHz. They are also known to be circularly polarized with a peak radiation direction perpendicular to the plane of the spiral. They are heavily used in the defense industry. The spiral antenna design started by wrapping

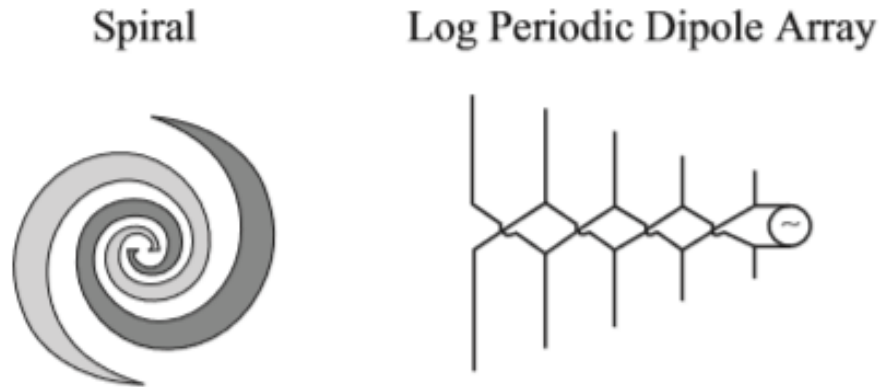


Figure 2.13: Examples of broadband antennas showing a circularly polarized antenna (left) and a log-periodic antenna (right) [7]

the arms of the dipole for investigation purposes by Edwin Turner in 1954 [15].

Another heavily used broadband antenna is the log periodic dipole array antenna shown in figure 2.13 (right). Like spiral antennas, they are designed to have an extensive bandwidth, with an infinite theoretical achievable bandwidth. The larger the structure, the lower the frequency limit in the bandwidth, and the finer the features on the antenna, the higher the upper frequency limit. In this type of antenna, an expansion factor is usually used, which means that each dipole is a certain percentage longer than the one before it, with each element being fed out of phase to the element on either side. This antenna has both active and inactive regions. For example, if the radiation mechanism is looked at at 500 MHz, the bulk of radiation will come from the dipoles with lengths near $\lambda/2$ at 500 MHz [16]. Although this might be similar to the Yagi-Uda antenna used domestically, log-periodic antennas are used commercially as they allow tuning over a range of frequencies.

In summary, the choice of antenna type implementation is influenced by many factors, of which gain is one of them. The following are some examples of antennas organized according to their corresponding gain [7]:

- ***For 5 dBi gain or less***
 - Electrically small antennas
 - Loops
 - Dipoles/Monopoles
- ***For 5 dBi to 8 dBi***
 - Microstrip patches
- ***For 8 dBi to 15 dBi***
 - Yagi-Uda
 - Helix/Spiral (axial mode)
 - Log periodic dipole array
- ***For 15 dBi and more***
 - Aperture Antennas
 - * Horns
 - * Reflectors

Following the discussion of antenna history and the various types, including resonant and directive designs such as Yagi antennas, the focus now shifts to a more contemporary class of antennas: microstrip patch antennas. As antenna technology advanced to accommodate the growing demand for compact, lightweight, and cost-efficient designs, particularly in mobile communication and satellite systems, microstrip patch antennas gained popularity due to their low-profile nature and ease of fabrication. These antennas have become an essential component in modern wireless communication due to their efficient design, adaptability, and integration with integrated circuits and systems.

Chapter 3

Microstrip Patch Antennas

Having explored the various types of antennas and their unique characteristics, it becomes evident that each design serves specific applications depending on the desired performance metrics such as size, gain, and bandwidth. Among these, microstrip patch antennas (MPAs) have gained significant popularity due to their low profile, ease of fabrication, and adaptability to various communication systems. Their compact design makes them ideal for applications requiring lightweight yet efficient solutions where these antennas are integral to achieving precise radar measurements while minimizing size and weight.

3.1 Overview

MPAs are among the most widely used types of antennas due to their lightweight, compact size, low cost, low profile, ease of fabrication, and conformity to various surfaces. In addition, they offer significant advantages, including the ability to provide dual and circular polarization, dual-frequency operation, frequency agility, broad bandwidth with customized designs, flexible feedline configurations, and beam-scanning capabilities with omnidirectional patterns [17]. The basic structure of a microstrip patch antenna, as depicted in Figure 3.1, includes a copper ground layer, a dielectric substrate,

and a copper patch layer. An additional critical component not shown in the figure is the feeding network, which can be implemented in a variety of ways, depending on the design and operational requirements of the antenna system. Besides, the patch can take different shapes: square, rectangular, circular, elliptical, triangular, disc sector, circular ring, and ring sector as shown in figure 3.2.

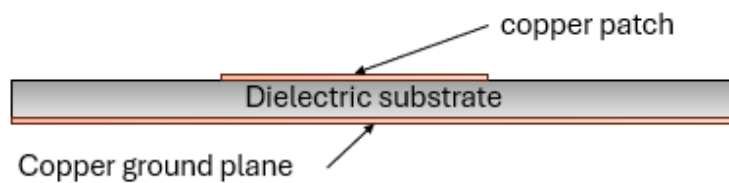


Figure 3.1: Sideview structure of a microstrip patch antenna

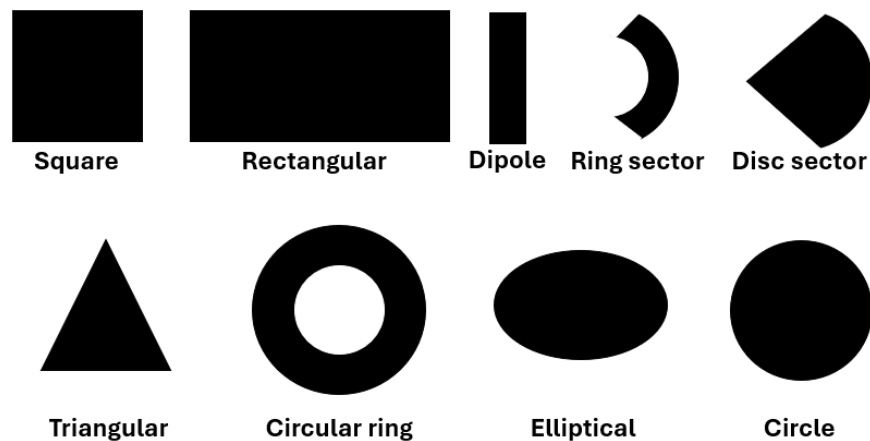


Figure 3.2: Various shapes of patches

This type of antenna has numerous applications, ranging from GPS to RF identification to mobile and satellite communication. This thesis focuses on radar applications

where the objective is to miniaturize the size to have a lighter-weight antenna with the same specifications.

The microstrip antenna is usually less than 0.05λ thick, where λ is the free space wavelength, with main constraints caused by physics rather than technology [7]. Due to its resonance properties, the bandwidth is a narrowband, which presents a challenge in getting the broad bandwidth needed for wireless and radio applications. In addition to the advantages of microstrip patch antennas mentioned at the beginning of this section, a mention of the disadvantages need to be highlighted, such as feed spurious radiation, poor cross polarization purity, limited power handling, and adjustment difficulty after fabrication [7].

When the length of the patch is approximately half of the wavelength in the dielectric λ_d , the antenna becomes resonant, meaning it exhibits zero input reactance. At this point, the antenna is referred to as a half-wave patch antenna [7].

The simplest formula for estimating patch size is the following:

$$L = 0.49\lambda_d = 0.49\frac{\lambda}{\sqrt{\epsilon_r}} \quad \text{half-wave patch} \quad (3.1)$$

Where λ is the free space wavelength at resonance and λ_d is the wavelength of a wave in an infinitely large block of dielectric with ϵ_r . This formula is a starting point for simulation and experimental hardware model [7]. The use of 0.49λ instead of 0.5λ here is a simplified approach to account for the fringing effects at the edges of the patch. These fringing effects cause the effective length of the antenna to be slightly greater than its physical length, necessitating a reduction in the actual patch length for accurate resonance. This adjustment compensates for the influence of the dielectric substrate and the edge-induced field distribution, resulting in more efficient and accurate antenna performance

The main problem with (3.1) is that accuracy degrades for large thicknesses, $t >$

0.02λ , and for high dielectric constants. Thus, the following equations (3.2) from [7] can be used for a more accurate resonant patch length.

$$L = 0.5 \frac{\lambda}{\sqrt{\epsilon_r}} - 2\Delta L \quad (3.2a)$$

$$\Delta L = 0.412 \frac{(\epsilon_{re} + 0.3) \left(\frac{W}{t} + 0.264 \right)}{(\epsilon_{re} - 0.258) \left(\frac{W}{t} + 0.8 \right)} t \quad (3.2b)$$

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{10t}{W} \right)^{-0.5} \quad (3.2c)$$

Where ΔL is the fringing length, t is the thickness of the substrate, W is the width of the antenna which is used to adjust the input impedance, and ϵ_{re} is the effective dielectric constant. ΔL increases with substrate thickness linearly as the electric fields extend farther and “fringe” more as the thickness increases. This means that ΔL is the amount removed from each edge to get the actual physical dimensions of the patch length.

3.2 Substrate Effect

While the fundamental structure of microstrip patch antennas, with their patch, dielectric substrate, ground plane, and feed line, plays a crucial role in determining their overall performance, the characteristics of the substrate itself are equally important. The substrate directly affects key parameters like the antenna’s bandwidth, radiation efficiency, and overall size. Understanding how material properties—such as permittivity and permeability—impact the performance of microstrip patch antennas is essential, especially when miniaturization techniques are being studied.

Substrate materials, which sit between the antenna’s radiating elements and the ground plane, can significantly influence how electromagnetic waves propagate through

and around the antenna. As antenna designs evolve to meet the demands of modern communication systems, the role of substrates has become a critical area of research, particularly when attempting to miniaturize antennas without compromising performance.

Simple antennas, such as monopoles and dipoles as shown in figure 3.3, traditionally rely on air or basic dielectric materials as substrates. However, as antenna applications have expanded into advanced radar systems and space missions, more complex designs, such as microstrip patch antennas, have emerged. These designs use substrates with varying permittivities and permeabilities to precisely control the antenna's electrical and radiative properties.

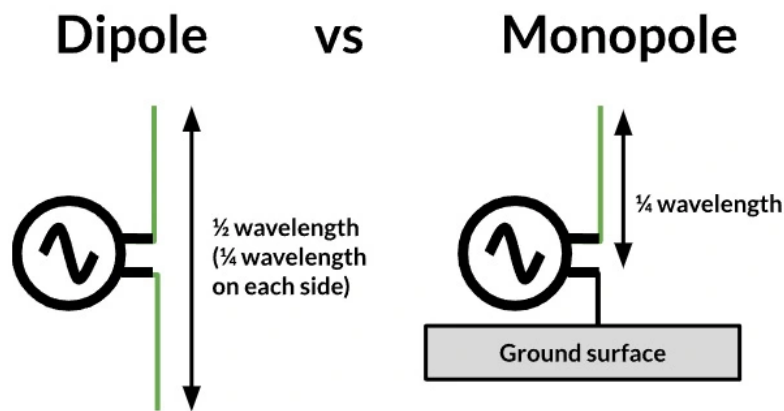


Figure 3.3: Dipole (left) vs. monopole (right) antenna diagrams

3.2.1 Permittivity

Permittivity is a fundamental property of dielectric materials that affects how electromagnetic waves propagate through the antenna by being a measure of the opposition offered against the formation of an electric field. Higher permittivity materials restrict wave propagation by confining the waves within the substrate, slowing down their velocity. This effect is crucial in applications where controlling the resonant frequency of

an antenna is important, such as the case with microstrip patch antennas. The resonant frequency, defined as the frequency at which the antenna naturally vibrates at its highest amplitude, is a key design parameter.

The standard formula for frequency in free space is the following:

$$f = \frac{c}{\lambda} \quad (3.3)$$

Where f is the center frequency, c is the speed of light (3×10^8 m/s), and λ is the wavelength of the wave, which is the distance between identical points in the adjacent cycles of a waveform signal propagating in a medium. The standard formula for the speed of light is the following:

$$c = \frac{1}{\sqrt{\epsilon_0 \mu_0}} \quad (3.4)$$

Once a substrate is introduced, electromagnetic waves no longer travel at the speed of light; however, they start traveling at lower speeds denoted by v . Then, with v , the frequency equation can be adjusted from (3.3) to (3.6) with $\mu_r = 1$ for nonmagnetic substrates.

$$v = \frac{1}{\sqrt{\epsilon_0 \mu_0}} = \frac{1}{\sqrt{\epsilon_0 \mu_0} \sqrt{\epsilon_r \mu_r}} = \frac{c}{\sqrt{\epsilon_r \mu_r}} = \frac{c}{\sqrt{\epsilon_r}} \quad (3.5)$$

$$f = \frac{c}{\lambda \sqrt{\epsilon_r}} \quad (3.6)$$

Equation (3.6) shows that as the substrate's relative permittivity, ϵ_r , increases, the desired resonant frequency decreases. However, high permittivity degrades the radiation efficiency and impedance matching of the implementable antenna [3].

3.2.2 Permeability

While permittivity has traditionally been the focus in antenna design, the role of permeability—how easily a material can support the formation of a magnetic field—has gained increased attention, especially in the development of magneto-dielectric materials. By adjusting both permittivity and permeability, designers can fine-tune the speed of wave propagation and improve antenna performance as the frequency formula changes from (3.6) to (3.7). Lowering the permittivity while increasing the permeability can help keep the same value of the desired resonant frequency.

$$f = \frac{c}{\lambda\sqrt{\epsilon_r\mu_r}} \quad (3.7)$$

Using materials with both high permittivity and high permeability, known as magneto-dielectrics, allows for further control over the antenna's electrical properties, particularly in terms of miniaturization. This dual control enables the design of smaller antennas that still meet specific performance requirements. By optimizing the balance between permittivity and permeability, it is possible to maintain the desired resonant frequency and functionality while reducing the overall size of the antenna [8].

It has been established that having a high permittivity substrate leads to having a narrower bandwidth with a low radiation efficiency. So, switching to a substrate with $\mu_r > 1$ helps keep the usage of a low permittivity substrate to compensate for the low radiation efficiency and also increase the percentage bandwidth while decreasing the patch size of the antenna according to [18]. This is possible because the non-unitary μ_r will offer resonance stability due to the confined magnetic fields which enhances the quality factor balance. Besides, researchers have emphasized investigating magneto-dielectric materials with approximately equal permittivity and permeability. This helps match the impedance of the antenna's substrate to free space which broadens the bandwidth and

makes it easier to match which also reduces reflections, distortions, and VSWR and maximizes efficiency [19] and [20]. The reason for this matching is that the free space impedance, $\sqrt{\frac{\mu_0}{\epsilon_0}}$, is equal to the substrate impedance, $\sqrt{\frac{\mu_0}{\epsilon_0} \frac{\mu_r}{\epsilon_r}}$, because $\mu_r \simeq \epsilon_r$.

By incorporating these materials into antenna design, researchers have the potential to create smaller, lighter antennas that can be used in future space missions, where every ounce of weight saved translates into significant cost savings. This thesis further explores how the use of magneto-dielectric materials can overcome the limitations of traditional substrates, enabling the miniaturization of antennas without a huge compromise in performance.

3.3 Feeding Techniques

Having discussed how the substrate's permittivity and permeability influence the performance of microstrip patch antennas, particularly in terms of bandwidth, efficiency, and size, it is important to understand how a microstrip patch antenna can be fed as that also affects some of its key performance parameters such as realized gain. The choice of feeding technique not only affects impedance matching but also plays a significant role in determining the antenna's radiation characteristics and overall efficiency.

There are many ways to feed a microstrip antenna, shown in figure 3.1, such as coaxial probe feeding, microstrip line feeding, aperture coupling, and proximity coupling.

3.3.1 Microstrip Line Feeding

The microstrip line feeding technique is a conducting strip with a much smaller width than the patch. This feeding technique is easy to fabricate and simple to match and model. As for the physics behind it, the edge nearer to the source has infinite impedance because of zero current at the open circuit edge. The impedance is zero at the center

of the patch because of the maximum current at the patch center. Moving closer to the patch center, a $50\ \Omega$ input impedance feed point is selected. The microstrip-fed patch antenna offers good gain but is inefficient because of spurious radiation from the feed line and patch junction [21]. Thus, its main disadvantage appears with increasing the thickness of the substrate as the surface waves and spurious feed radiation will increase, limiting the line's ability to provide a good bandwidth.

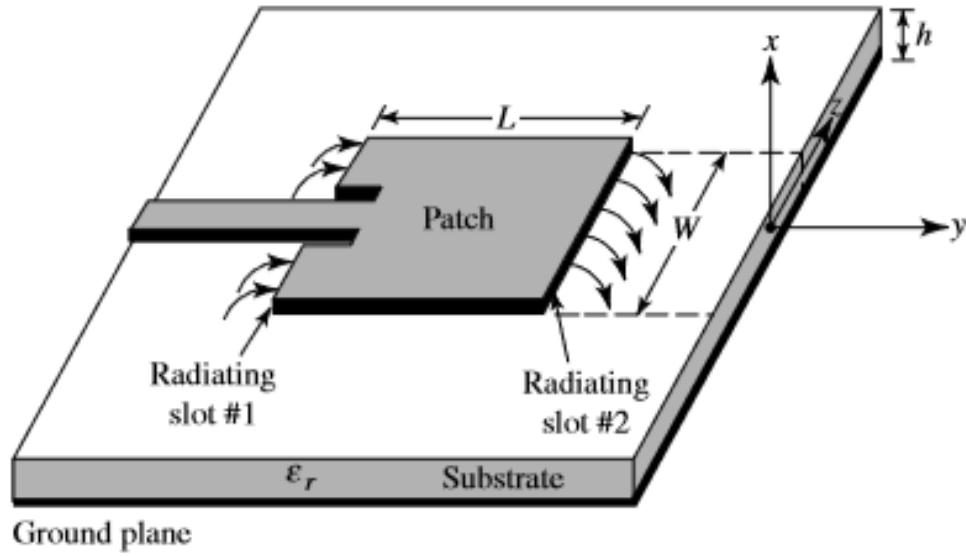


Figure 3.4: Microstrip line feeding into a rectangular patch from [8]

Figure 3.4 shows the microstrip line feeding configuration for a microstrip patch antenna. The microstrip feed line is inset into the patch by a specific distance, designed to achieve impedance matching between the feed line and the patch antenna. The inset depth is determined based on the patch's dimensions $L \times W$, as well as the substrate's dielectric constant ϵ_r and thickness h and is usually chosen to ensure a characteristic impedance of $50\ \Omega$ at the feed point. The inset feed plays a crucial role in ensuring efficient power transfer and minimizing reflection losses. By precisely tuning the feed position, the input impedance of the patch is matched to the feed line's impedance, re-

sulting in improved radiation efficiency and bandwidth. Additionally, this configuration impacts the excitation of the two radiating slots, ensuring uniform radiation and a symmetric radiation pattern.

3.3.2 Proximity Feeding

The proximity feeding technique works by placing a feed line between two dielectric substrates, as shown in figure 3.5. This presents a problem of having difficulty in choosing a $50\ \Omega$ feeding point as the impedance at the edges will be very high. To overcome this problem, the feed line is moved to a lower level below the patch separated by a dielectric substrate of relative permittivity, ϵ_{r1} , as shown in figure 3.5 with its edge located at a point where the antenna's input impedance is $50\ \Omega$. The rest of the design is still the same as figure 3.1 of having the top part separated from a ground plane by a dielectric substrate of relative permittivity, ϵ_{r2} , in this case. Power transfer takes place through electromagnetic field coupling. This technique reduces feed line radiation and allows planar feeding while improving bandwidth [21].

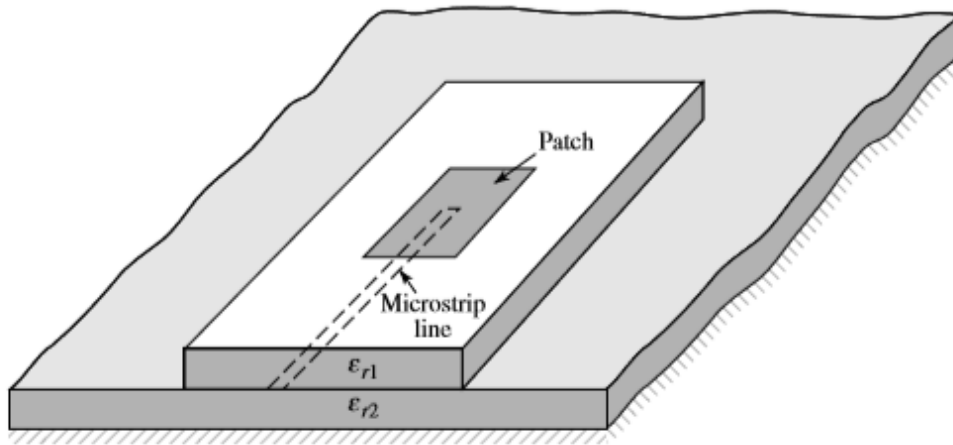


Figure 3.5: Proximity feed technique into a rectangular patch [7]

3.3.3 Aperture Couple Feeding

The aperture coupling feeding technique is slightly complicated compared to the other ones, as it consists of two dielectric substrates: the antenna patch substrate and the feed substrate. Figure 3.6 shows how this feeding configuration is established where both substrates are separated by a ground plane with a slot in the center. Then, the metal patch is placed on the top of the antenna substrate, whereas the ground is placed on the other side of the antenna substrate. On the other side of the ground plane, the feed dielectric and feed line are placed perpendicularly to the ground plane slot to provide isolation. The feed line causes magnetic polarization in the ground plane slot, which excites the metal patch. The center of the patch and the slot must coincide for maximum coupling. The termination point of the feed line is crucial because it determines how efficiently power is transferred from the feed line to the patch. Proper termination minimizes reflection losses and ensures strong coupling. The feed line length is carefully designed to maintain the correct impedance and phase relationship between the feed and the patch. Any mismatch or improper termination can lead to reduced efficiency or distortion of the radiation pattern. This technique is suitable for independently optimizing the antenna and feed line substrates. The isolation feature prevents spurious radiation from corrupting the radiation pattern [21]. However, it has a disadvantage similar to the proximity feed technique, which has a multilayer fabrication process that will add more complexity to the stacked design.

3.3.4 Coaxial Probe Feeding

Coaxial probe feeding is a non-planar feeding technique in which a coaxial cable feeds the patch. The inner conductor of the coax penetrates through the dielectric layer to be attached to the patch at a $50\ \Omega$ input impedance point, whereas the outer connector

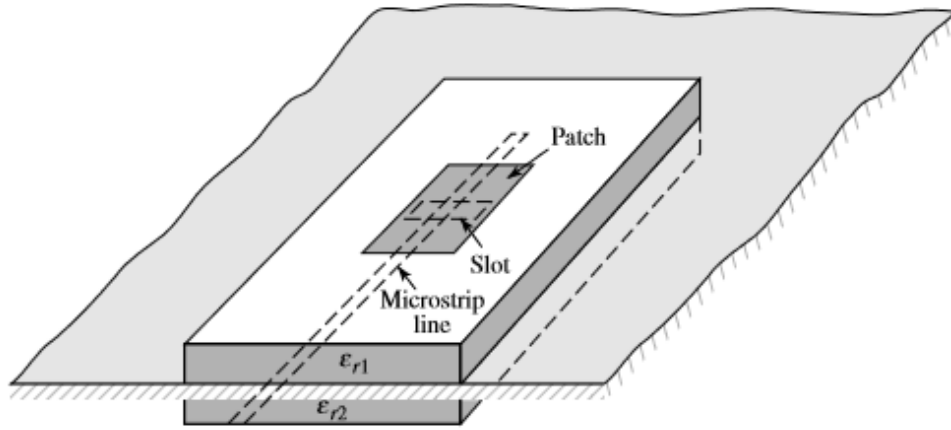


Figure 3.6: Aperture feed technique into a rectangular Patch [7]

is connected to the ground plane. Similar to microstrip line feeding, this is easy to fabricate and match while exhibiting low spurious radiation. The probe can be placed at any desired location to match the input impedance, which makes matching, in this case, easier than the microstrip line feeding technique. The probe generates high cross-polarized fields if electrically thick substrates are used.

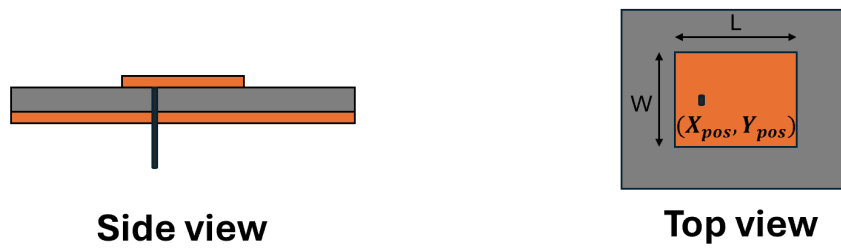


Figure 3.7: Coaxial probe feeding into a rectangular patch

To have an initial theoretical coordinate point of where the location of the feed will be implemented, figure 3.7 can be referred to where the point of contact will be defined

by a coordinate point (x-coordinate, y-coordinate). The following equations show how to obtain these coordinates:

$$X_{pos} = \frac{L}{\sqrt{\epsilon_{re}}} \quad (3.8a)$$

$$Y_{pos} = \frac{W}{2} \quad (3.8b)$$

3.4 Analysis Methods

The feeding technique plays a vital role in determining how effectively the microstrip patch antenna operates by influencing impedance matching and radiation characteristics. However, understanding the performance of these antennas requires more than just selecting the right feed. Rigorous analysis methods are essential to accurately predict antenna behavior, optimize designs, and ensure the desired performance is achieved.

When it comes to analysis, there are two main analytical methods to predict how a design can behave or to control how it will behave. These two main analytical methods are the transmission-line and cavity models, which will be elaborated on in subsections 3.4.1 and 3.4.2.

3.4.1 Transmission Line Model

The transmission line model is essentially a one-dimensional model that assumes uniform fields in the y-direction in the patch and treats the patch as two parallel radiating slots, each having an equivalent parallel slot admittance. The viewpoint of these radiating slots helps find the radiation pattern of a rectangular patch, with the region between the patch and the ground acting as a half-wavelength transmission line that is

open-circuited at the end [7].

Thus, this model is more accessible than the cavity model discussed in subsection 3.4.2 but is less accurate and lacks versatility. The fields at the patch's edges undergo fringing because of the finite dimensions of the patch along the length and width. To design a patch considering fringing effects, (3.2) from section 3.1 can be used. One of the main equations in patch design is the one for the dominant TM_{010} mode where the resonant frequency is a function of its length, which can be given by the following equation:

$$(f_r)_{010} = \frac{1}{2L\sqrt{\epsilon_r\mu_r}\sqrt{\epsilon_0\mu_0}} = \frac{c}{2L\sqrt{\epsilon_r\mu_r}} \quad (3.9)$$

However, (3.9) does not account for fringing. Thus, it does not fit the transmission line model. It can be modified to account for fringing according to the following equation:

$$(f_{rc})_{010} = \frac{1}{2L_{eff}\sqrt{\epsilon_{re}\mu_{re}}\sqrt{\epsilon_0\mu_0}} = \frac{c}{2(L + 2\Delta L)\sqrt{\epsilon_{re}\mu_{re}}} = q\frac{c}{2L\sqrt{\epsilon_{re}\mu_{re}}} \quad (3.10)$$

Where q is the fringing factor that represents the length reduction factor $q = \frac{(f_{rc})_{010}}{(f_r)_{010}}$. When the substrate height increases, the fringing will also increase, leading to a larger separation between radiating edges and lower resonant frequencies [8].

Equation (3.10) can be rearranged to solve for the length of the patch. As for the patch's width, the transmission line model in [8] provides an equation where $\mu_r = 1$ that can be seen in (3.11).

$$W = \frac{c}{2f_r}\sqrt{\frac{2}{\epsilon_r + 1}} \quad (3.11)$$

Figure 3.8 shows how the transmission line model of the patch antenna works where the patch acts as two parallel radiating slots. Each slot is represented by a parallel

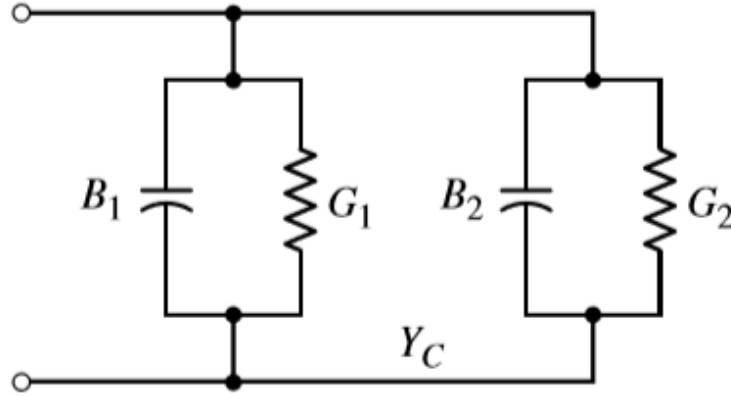


Figure 3.8: Transmission model equivalent of a microstrip patch antenna [8]

equivalent admittance Y_c (with conductance G and susceptance B). The equivalent admittance represents how the patch antenna interacts with the electromagnetic fields by summing the conductance and susceptance. It helps determine the input admittance of the antenna at the resonant frequency, which ensures maximum power transfer and minimal reflections. Conductance represents the real part of the admittance and is related to the power radiated by the antenna. A higher conductance indicates more efficient radiation because it quantifies the portion of the input power that is converted into radiated electromagnetic waves. On the other hand, susceptance represents the imaginary part of the admittance and is related to the stored reactive energy in the antenna's near field. Proper tuning of the susceptance ensures efficient operation at the desired frequency, and a well-matched susceptance can lead to a broader bandwidth by reducing reactive losses.

3.4.2 Cavity Model

The cavity model is more generally applicable than the transmission line model due to its approximation method. It includes all internal variations of the patch through a

summation of modal fields, allowing for a detailed representation of the electromagnetic behavior within the patch [7]. This means that the normalized fields within the dielectric substrate between the patch and the ground plane can be found more accurately by treating the region as a cavity bounded by electric conductors (ground and patch) and by magnetic walls to simulate an open circuit along the patch's perimeter. This means that the fields inside the patch can be described using a summation of different modes, allowing the model to account for the complex interactions of these modes.

Although this is an approximation method, it is considered an acceptable approach as computed pattern, input admittance, and resonant frequencies often compare well with measurements. When the microstrip patch is energized, a charge distribution is established on the upper and lower surfaces of the patch as well as the surface of the ground plane. This charge distribution is controlled by two mechanisms: attractive and repulsive. The attractive mechanism is between opposite charges on the bottom of the patch and the ground plane, whereas the repulsive mechanism is between the like charges on the bottom of the patch, which pushes charges from the bottom to the edge and top of the patch. The movement of these charges creates current densities at the bottom and top surface of the patch.

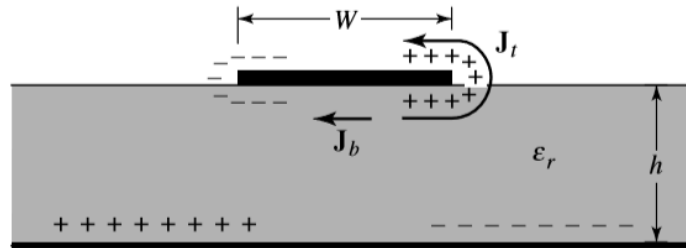


Figure 3.9: Charge distribution and current density creation [8]

If the microstrip antenna is only considered a cavity, the absolute amplitudes of the electrical and magnetic fields would be impossible to find. Thus, by treating the walls

of the cavity and the substrate as lossless, the cavity would not radiate, and its input impedance would be purely reactive, with a function representing impedance that only has real poles. A loss mechanism has to be introduced to account for radiation by taking the radiation resistance R_r and the loss resistance R_L into account. For the cavity model, introducing an effective loss tangent, $\tan(\delta_{eff})$ will help take care of the loss [8]. This will make the cavity model behave like an antenna by having a complex input impedance with a function that has complex poles, $\tan(\delta_{eff}) = \frac{1}{Q}$.

Having a relatively small thick substrate introduces many reflections for the waves between the patch and the ground before reaching the edge, which means that only a small fraction of the incident energy is radiated, making the antenna very inefficient. Because of this and the small fringing fields compared to the normal electric field distribution along the surface of the patch, only TM^x field configuration will be considered within the cavity. While the top and bottom walls of the cavity are perfect electric conductors (PECs), the four side walls will be modeled as perfect magnetic conductors (PMCs).

The following equation gives the resonant frequencies for the cavity:

$$(f_r)_{mnp} = \frac{1}{2\pi\sqrt{\epsilon\mu}} \sqrt{\left(\frac{m\pi}{t}\right)^2 + \left(\frac{n\pi}{L}\right)^2 + \left(\frac{p\pi}{W}\right)^2} \quad (3.12)$$

Where m , n , and p represent the number of half-cycle field variations along the x, y, and z directions, respectively.

However, this is the general equation. The dominant mode is determined by looking at the mode with the lowest-order resonant frequency. For all Microstrip antennas, $h \ll L$ and $t \ll W$, which means that $L > W > t$, this helps place the resonant frequencies in ascending order to help determine the mode of operation. Thus, the first dominant mode will be TM_{010}^x given by (3.13a).

$$(f_r)_{010} = \frac{1}{2L\sqrt{\epsilon\mu}} = \frac{c}{2L\sqrt{\epsilon_r\mu_r}} \quad (3.13a)$$

$$(f_r)_{001} = \frac{1}{2W\sqrt{\epsilon\mu}} = \frac{c}{2W\sqrt{\epsilon_r\mu_r}} \quad (3.13b)$$

$$(f_r)_{020} = \frac{1}{L\sqrt{\epsilon\mu}} = \frac{c}{L\sqrt{\epsilon_r\mu_r}} \quad (3.13c)$$

If in addition $L > W > L/2 > h$, the next higher order mode will be given by (3.13b). But, if $L > L/2 > W > t$, the second order mode will be given by (3.13c).

If $W > L > t$, the dominant mode would be given by (3.13b) while if $W > W/2 > L > t$, the second order mode will be given by TM_{002}^x whose equation can be obtained by plugging 002 for mnp in (3.12).

Thus, the distribution of the tangential electric field along the side walls of the cavity for the TM_{010}^x , TM_{001}^x , TM_{020}^x , and TM_{002}^x is shown in figure 3.10.

The cavity model does not fully consider the fringing effects compared to other models, such as the transmission line model, which makes it slightly less accurate. However, this is a good approximation for most practical applications, balancing simplicity and sufficient accuracy. It provides good approximation due to the trade-off between model complexity and computational efficiency and the simplified analysis that comes with treating the patch as a resonant cavity, making it easier to derive and understand its key parameters. It also does not lead to significant errors by not accounting for fringing effects, which is where the balance between simplicity and accuracy comes in.

3.5 Phased Array Antennas

After discussing the analysis methods used to evaluate the performance of individual microstrip patch antennas, it's important to expand the focus to more complex configu-

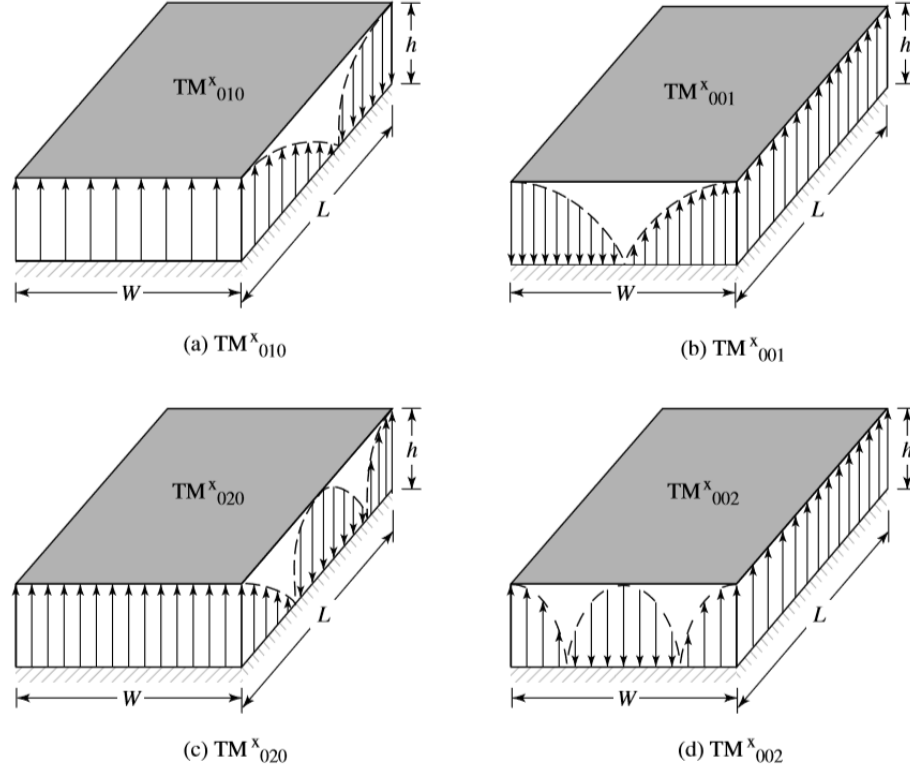


Figure 3.10: Field configurations (modes) for rectangular microstrip patch antenna [8]

rations. One such configuration, the phased array, leverages multiple antennas working together to achieve enhanced directionality, beam steering, and increased gain as is the case with NASA's EcoSAR system.

3.5.1 Basics and Historical Development

Phased array antennas are advanced antenna systems composed of multiple individual radiating elements, each of which can be controlled with varying phase shifts to steer the overall beam direction without physically moving any of the antennas. This process will help control the direction and the shape of the radiation pattern, allowing for rapid beam steering and enhanced signal strength in desired directions [8].

The concept of phased array antennas emerged during the Second World War and

was driven by the need for advanced radar systems. This concept led to the development of a rudimentary phased array system. Figure 3.11 shows one of the world's first phased array antennas, the FuMG 41/42 Mammut, built by Germany. It consisted of 6 to 8 Freya antenna arrays that are switched together and coupled to two Freya devices. Its beam can be electronically steered on a 100-degree arc in front and behind the antenna, making it able to detect targets flying at an altitude of 8 km at a range of 300 km [22].

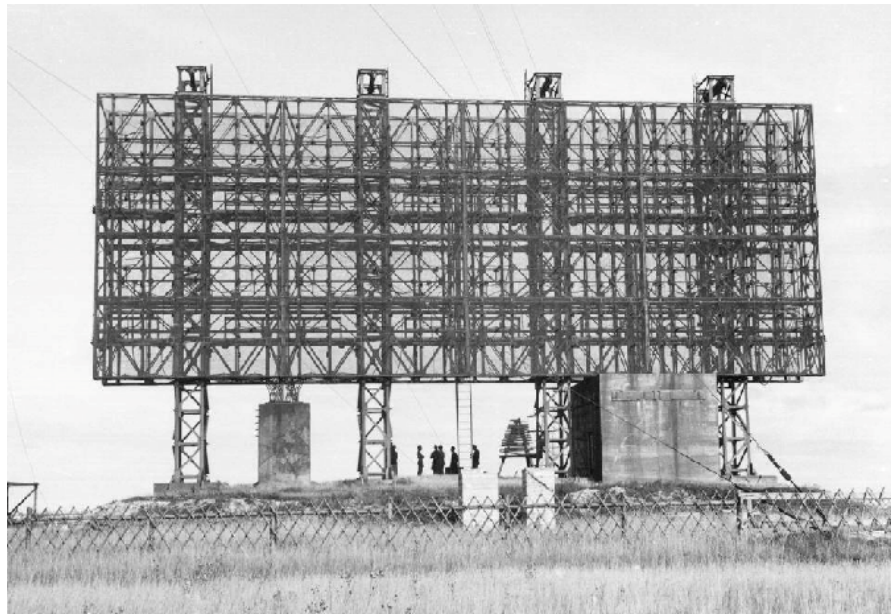


Figure 3.11: The MAMMUT phased array radar

Each radiating element on the phased array is fed with a signal that can be independently adjusted in phase and amplitude. Building on this fundamental concept, the array's overall functionality can be outlined in four key components:

- Element array: the array comprises numerous individual radiating elements arranged in a regular pattern such as linear, planar, or circular. In the case of NASA's EcoSAR array, the radiating elements are patches.
- Phase shifters: each element is connected to a phase shifter that adjusts the signal's

phase, feeding the element that allows for controlling the radiation pattern's first lobe direction.

- **Constructive and destructive interference:** when the signals from each element are in phase in a specific direction, they combine constructively, resulting in an intense radiation pattern in that direction.
- **Beamforming:** the process of adjusting the phase and amplitude of the signals to form a desired radiation. It can be static when the desired pattern is fixed or dynamic when it needs to be rapidly adjusted to track moving objects.

3.5.2 Design Principles

Designing phased array antennas involves several vital equations essential for determining performance characteristics. These equations provide a basic understanding of how these arrays operate and get optimized.

One of the main concepts to analyze and predict the behavior of the radiation pattern of an antenna array is the array factor (AF). For a linear array with N elements spaced d and a progressive phase shift β , the array factor is given by:

$$AF = \sum_{n=1}^N e^{j(n-1)\psi} \quad \text{where} \quad \psi = kd \cos(\theta) + \beta \quad (3.14)$$

Where n is the position of the n -th element, $k = \frac{2\pi}{\lambda}$ is the wavenumber, ψ is the progressive phase, θ is the angle of observation, and β_n is the phase shift of the n -th element [8].

After accounting for constructive and destructive interference of signals using the array factor, calculations for the degree of steering can be done using the following

equation:

$$\theta_0 = \cos^{-1} \left(\frac{\beta\lambda}{2\pi d} \right) \quad (3.15)$$

Where θ_0 is the main beam direction, β is the progressive phase shift between adjacent elements, and d is the spacing between the elements.

Then, to avoid grating lobes and make sure that the antenna will not produce unwanted side lobes that would interfere with the desired signal pattern, the element spacing should be less than or equal to half the wavelength ($d \leq \frac{\lambda}{2}$).

Another important equation that is needed is the phase shift for beam steering, which is fundamental for dynamically adjusting the beam direction in response to changing conditions or target movement [20]. It is given by (3.16).

$$\beta_n = -nkd \cos(\theta_0) \quad (3.16)$$

Moreover, the designed phased array using MAGTREN will be a denser array with elements spaced less than $0.5\lambda_0$ apart, which will bring mutual coupling into account. Mutual coupling is an important concept that refers to the interaction between neighboring elements of the array, where the electromagnetic field of one element affects the performance of adjacent elements. This effect can influence the radiation pattern, impedance matching, and overall array performance.

3.6 Relevant Work

3.6.1 Magneto-dielectric Materials

While it has been established that a microstrip antenna with a coaxial probe feeding has a narrow bandwidth, altering the design to broaden the bandwidth is desirable for

many applications. One way is to use a substrate with magneto-dielectric properties, $\mu_r > 1$. The design in [23] is mainly theoretical due to the unavailability of a magneto-dielectric substrate to achieve the desired outcome in L-band, 1 GHz to 2 GHz, at that time. So, it followed a genetic algorithm (GA) to achieve the required outcome. The GA was binary coded with parameters being quantized into 16 bits to account for the large number of design parameters involved. The purpose of the theoretical antenna was met after a long run of the algorithm, significant miniaturization of up to 60% was achieved while providing a large operating bandwidth, 20%.

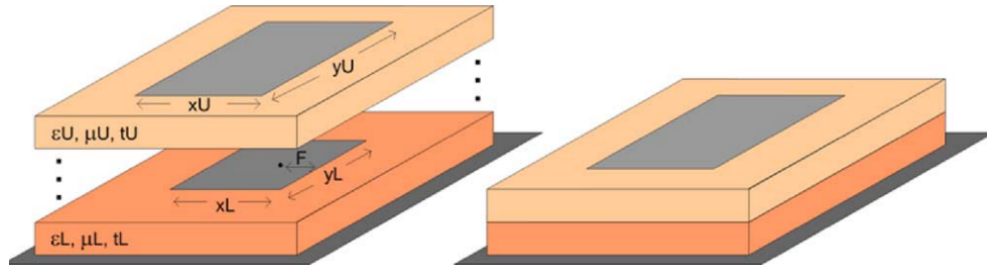


Figure 3.12: Expanded and collapsed view of a stacked-patch antenna [23]

Unlike the theoretical work in [23], a MAGTREX substrate provided by Rogers was used in [24] to operate in L-band. MAGTREX2000 experimental enjoys high permeability, low permittivity, low magnetic loss, and a low dielectric loss over the GNSS band. A different MAGTREX material is used in this thesis which is explored in chapter 4. For the work in [24], an approximated bandwidth for the antenna design was obtained according to the following:

$$BW = \frac{96 \sqrt{\frac{\mu_r}{\epsilon_r}} \frac{t}{\lambda_0}}{\sqrt{2} [4 + 17 \sqrt{\mu_r \epsilon_r}]} \quad (3.17)$$

where λ_0 is the free-space wavelength and t is the thickness of the substrate.

Equations for wavelength and impedance in the patch substrate material were derived in [25] according to the following:

$$\lambda = \frac{\lambda_0}{\sqrt{\epsilon_r \mu_r}} \quad (3.18a)$$

$$Z = Z_0 \sqrt{\frac{\mu_r}{\epsilon_r}} \quad (3.18b)$$

In addition to these two equations, equations in section 3.4 have been adjusted to account for permeability according to the following:

$$W = \frac{c}{2f\sqrt{\epsilon_r \mu_r}} \quad (3.19)$$

$$L = \frac{c}{2f\sqrt{\epsilon_{re} \mu_{re}}} - 2\Delta L \quad (3.20)$$

$$\text{where } L = 0.412t \cdot \frac{\epsilon_{re} \mu_{re} + 0.3}{\epsilon_{re} \mu_{re} - 0.258} \cdot \frac{\frac{W}{t} + 0.262}{\frac{W}{t} + 0.813} \quad (3.21)$$

Another equation for the position of the feed in the X_{pos} from section 3.3.4 was also obtained according to the following:

$$X_{pos} = \frac{L}{\pi} \sin^{-1} \sqrt{\frac{R_i}{R_e}} \quad (3.22)$$

where R_i is the input resistance, and R_e is the resistance at the edge, whose equation is the following:

$$R_e = \frac{\eta \lambda_0}{\pi W \left[1 - \frac{(kt)^2}{24} \right]} \quad (3.23)$$

where η is the radiation efficiency and k represents the free-space wavenumber, $2\pi/\lambda_0$.

Then, the actual substrate sizes that account for the fringing effects will be the following:

$$W_s = W + 12t \quad (3.24a)$$

$$L_s = L + 12t \quad (3.24b)$$

Besides, the most relevant work to this thesis is presented by [26] and [27] where MAGTREX 555 is used. However, this work does not go over a stacked patch configuration. Upon reviewing them, it was noticeable that the patch antenna has a negative gain, meaning that it is less than one compared to an isotropic antenna.

3.6.2 Stacked Microstrip Patch Antennas

While phased array antennas demonstrate the power of multiple elements working in unison to enhance performance, further advancements can be made by optimizing individual antenna elements. One such approach is the stacked microstrip patch antenna, which addresses limitations such as narrow bandwidth and low gain by layering multiple patches on top of one another.

A solution suggested by [7] is to design a stacked microstrip patch antenna. Besides, [3] indicates that a broader bandwidth of the antenna can be achieved while simultaneously miniaturizing its size using magneto-dielectric material. Having a stacked patch antenna will not only improve the bandwidth but also lower the cross-polarization and enhance the antenna gain and efficiency. There have not been many equations that show how a stacked patch antenna needs to be designed, especially in the presence of a substrate with non-unitary μ_r . For that, this part focuses on stacking patches made with non-magneto-dielectric materials.

The work in [28] shows how a stacked microstrip patch antenna was designed and fed with a coax to achieve dual resonance. This helps by having antennas with different functionalities depending on the resonance locations excited by the patches. So, the stacking configuration was implemented to achieve dual resonances far from each other

instead of focusing on a single resonance whose bandwidth is to be broadened, which is explored in this thesis.

Another relevant work to achieve a wider bandwidth is shown in [29]. However, the author did not try miniaturizing the antenna as the substrate had no magnetic properties. He got the highest possible bandwidth he aimed for by using a direct coaxial feed into the first patch. The top patch was then coupled with the lower patch. His design is shown in figure 3.13 which resulted in a measured axial ratio of less than 2.5 dB over a 13% bandwidth.

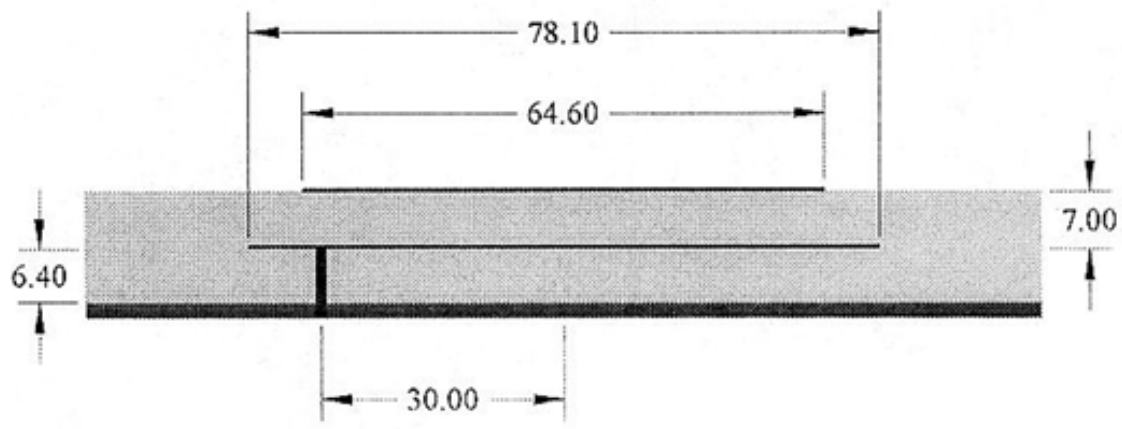


Figure 3.13: Figure showing the dimensions of the designed antenna from [29]

On the other hand, [30] and [31] suggested a similar approach of having three substrates for a stacked antenna of two patches, unlike [29]. The proposed third substrate is to have a lower permittivity so that effective permittivity can be controlled without having to adjust the layers with the patch. The location of the feed points between both works is different, and so is the order of the layers. Both antennas in figure 3.14 operate in C-band, 4 GHz to 8 GHz.

Another design in [32] was made to operate at three different bandwidths with circular polarization. The feed line is designed to get different vias to excite different patches at different times. This is different from the research focus of this thesis as the patches

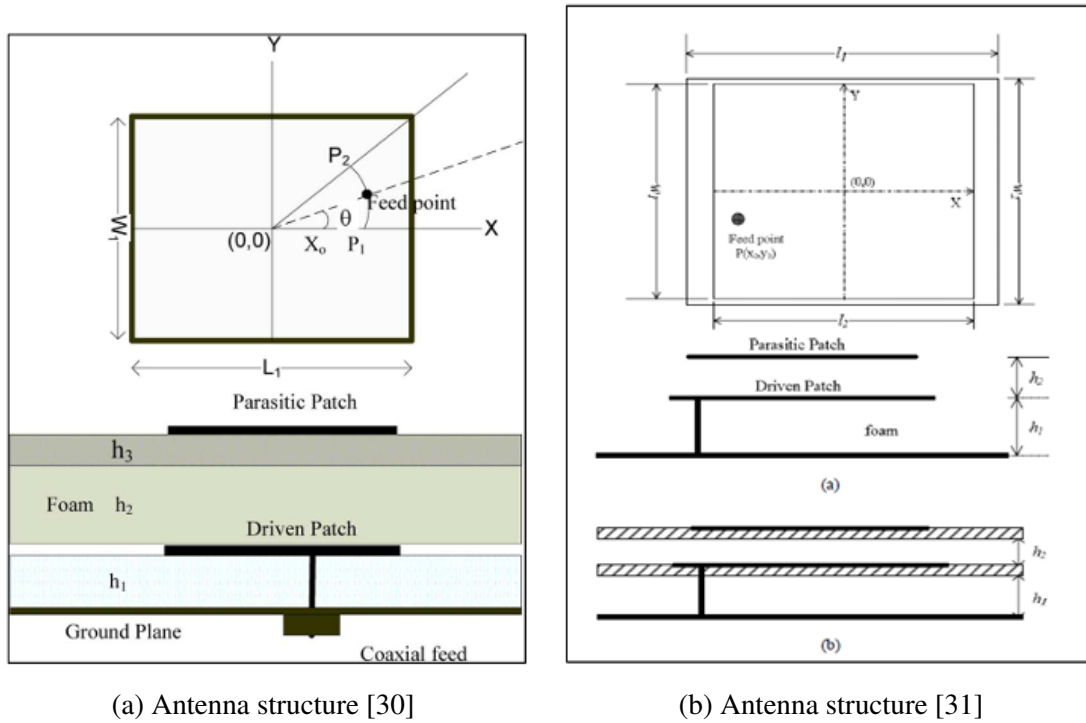


Figure 3.14: Antenna structure of two different design layouts

here are designed to couple and work together rather than separately. The non-excited patches in figure 3.15 are intended to act as ground planes when not used.

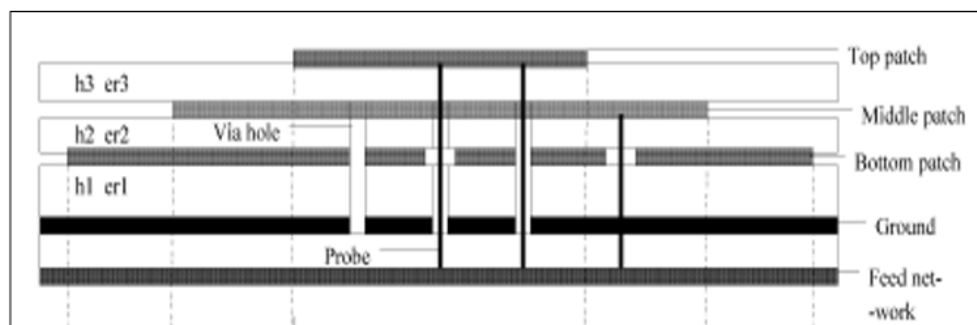


Figure 3.15: Multiple patches antenna from [32]

After reviewing all the previous work, it is clear that there are no in-depth studies of equations related to stacked Microstrip patch antennas, especially with magneto-dielectric materials. Due to the topic's novelty, there is very little information related

to stacked microstrip patch antennas made of magneto-dielectric substrates. The next chapter discusses and presents the implemented design of the stacked patch of this thesis. It is based on equations discussed thus far in this document and then designed and optimized using simulator software such as Ansys HFSS. This thesis differs from the presented relevant work because it implements magneto-dielectric material in a stacked patch configuration. It also presents solutions to the negative gain shown in [26] and [27].

Chapter 4

Design and Findings

With a comprehensive understanding of microstrip patch antennas in chapter 3, including their structure, substrate effects, feeding techniques, and analysis methods, now it is time to move into the design phase. In this chapter, the foundational concepts to develop a stacked microstrip patch antenna tailored to meet the specifications of NASA's EcoSAR system are applied. Given what has been established in the previous chapters, a microstrip patch antenna is designed to match EcoSAR's antenna elements in order to examine the limitation of magneto-dielectric materials for patch antenna realization. Coaxial probe feeding was chosen to be implemented from the feeding methods to avoid any errors that might arise from asymmetric fields caused by the other techniques. Besides, previous trials to work with such materials have been shown in [26] and [27] to use coaxial probe feeding due to the presence of some equations to support its design. This design follows equations from the cavity model as it is more applicable than the transmission line model because it includes all variations through a summation of modal fields [7].

With this being said, analytical calculations are set up and followed through before using HFSS and MATLAB to implement the design and tune it further to get a working prototype, plot the findings, and explore the limitations. A single patch antenna is investigated first to explore the MAGTREX 555 material. Then, to see the material's

miniaturization level, the designed antenna is compared to a reference design that uses RO3006. After exploring the limitations to obtaining an achievable microstrip patch antenna using the MAGTREX material, the bandwidth is broadened by exploring a stacked configuration for the antenna. Once the bandwidth limitation is explored, this chapter moves to explore the radiation patterns of the designed antenna.

4.1 Single Microstrip Patch Antenna

4.1.1 Magneto-dielectric Material

A square patch antenna has been implemented due to its ability to achieve vertical, horizontal, and circular polarization. The design of the MPA needed for this thesis focuses on linear polarization. The required resonant frequency is 435 MHz, and the desired bandwidth is 120 MHz, measured at the 10 dB return loss line to meet the specifications of the EcoSAR antenna by NASA.

A half-wave dipole antenna operates close to a resonant condition of its input impedance, and thus, its length is kept equal to $\lambda_0/2$. The edges of a square patch antenna are treated as half-wave dipoles, so they have to be kept at less than 344.8 mm ($\sim 13.6''$). For this design, a square ground plane whose dimensions are 6"x6" were used to keep the antenna at a good miniaturized level.

To get the length of the patch, the cavity model from section 3.4.2 was used as its equations account for permeability. MAGTREX 555 comes with the following parameters: $\epsilon_r \sim 6$, $\mu_r = 6$, dielectric loss tangent of 0.01, and a magnetic loss tangent of less than 0.05. A magnetic loss tangent of 0.05 is selected for the worst gain case; however, the work in [26] can be referred to for varying magnetic loss tangent analysis.

Single MPA: 60-mil-thick substrate

Equation (3.13a) was rearranged and used with the following inputs: $c = 3 \times 10^8$ m/s, $(f_r)_{010} = 435$ MHz, $\epsilon_r = 6$, and $\mu_r = 6$. A square patch means $L = W = 57.471$ mm. After designing in HFSS, the length was optimized to reach the resonant frequency of 435 MHz by changing it to 61.341 mm. The extra ~ 4 mm is due to the effective length not being accounted for in the cavity model. The equation in section 3.1 does not account for permeability.

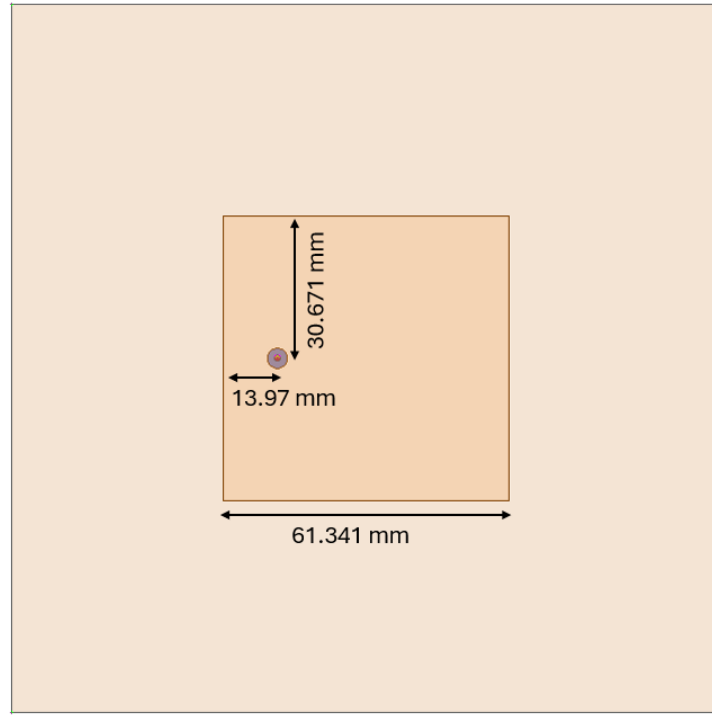
As for the position of the feed, (3.8) was used. We have the following parameters: $L = W = 57.471$ mm, a calculated ϵ_{re} of 5.7226 according to (3.2c), and a measured μ_{re} of 6 assuming that $\mu_{re} = \mu_r$ due to the unavailability of μ_{re} equation. Then, (3.8a) was modified to the following:

$$X_{pos} = \frac{L}{\sqrt{\epsilon_{re}\mu_{re}}} \quad (4.1)$$

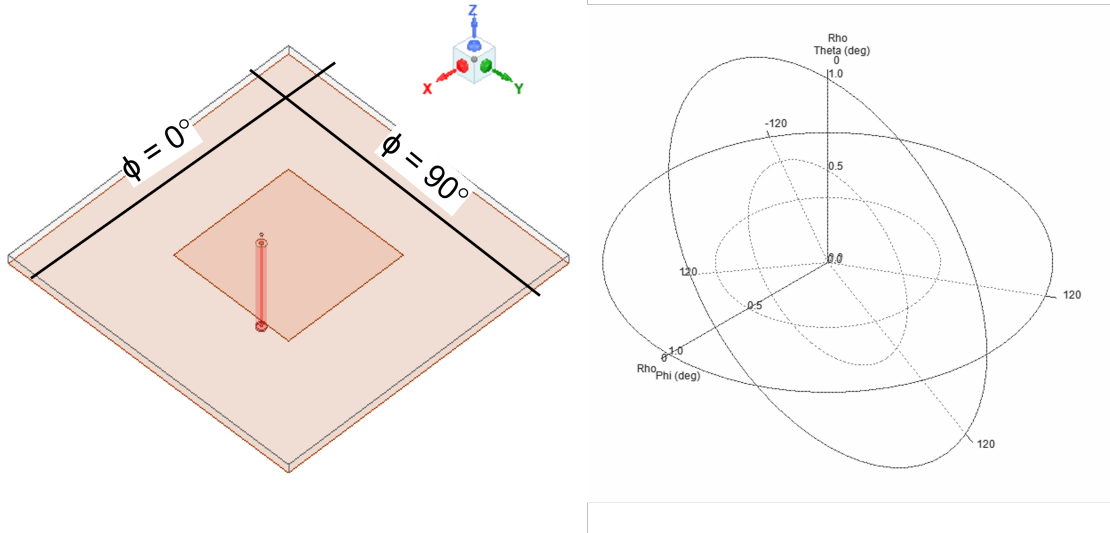
Then, the feeding coordinates were obtained according to the following: $Y_{pos} = W/2 = 28.736$ mm, whereas $X_{pos} = 9.579$ mm (10.468 mm if optimized L from HFSS is used). After optimization in HFSS, the actual Y_{pos} is found to be at 30.671 mm and X_{pos} to be 13.97 mm. The reason behind this discrepancy is the change in the actual patch size and the absence of a μ_{re} equation.

The microstrip patch antenna in figure 4.1 comprises three layers: a ground copper layer, a 60-mil-thick MAGTREX substrate, and a patch copper layer. An $|S_{11}|$ response is obtained by running a simulation of the design in HFSS to see how this antenna is behaving.

Figure 4.2 shows a resonance frequency of 435 MHz at -27.77 dB, meaning that the antenna is well matched, along with a bandwidth of 17.29 MHz from 426.02 MHz to 443.31 MHz, a fractional bandwidth of 3.97%. Keeping in mind that maximizing the



(a) Top View



(b) Isometric view

Figure 4.1: 60-mil-thick single MPA with $L = W = 61.341$ mm, $Y_{pos} = 13.97$ mm, and $X_{pos} = 30.671$ mm

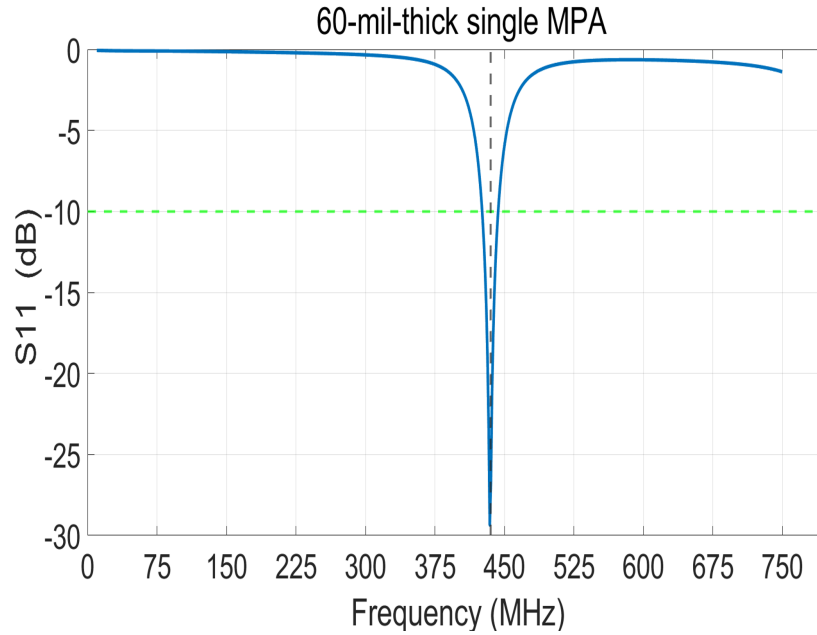


Figure 4.2: S_{11} in dB plotted as a function of frequency (MHz)

bandwidth with a stretch goal of getting close to 120 MHz is desired, further modifications are needed.

Figure 4.3 shows the radiation pattern of the antenna from which it can be seen that it exhibits a maximum gain of -10.4 dB with significant backlobe radiation ($\simeq$ -11.75 dB). Although this gain shows that the material is lossy, looking at the realized gain cut planes to see the cross-polarization in the antenna helps give more information about the antenna's efficiency.

From figure 4.4, the maximum elevation cross-polarization for the cut plane at $\phi = 0^\circ$ is -45.4 dBi and that of co-polarization is -10.4 dBi. Moreover, the maximum elevation cross-polarization for the cut plane at $\phi = 90^\circ$ is -49.9 dBi, and that of co-polarization is -10.4 dBi. This means that the cross-polarization is 35 dB below the desired main beam for $\phi = 0^\circ$ and 39.5 dB below the desired main beam for $\phi = 90^\circ$. Hence, the antenna effectively maintains the desired polarization while minimizing cross-polarization.

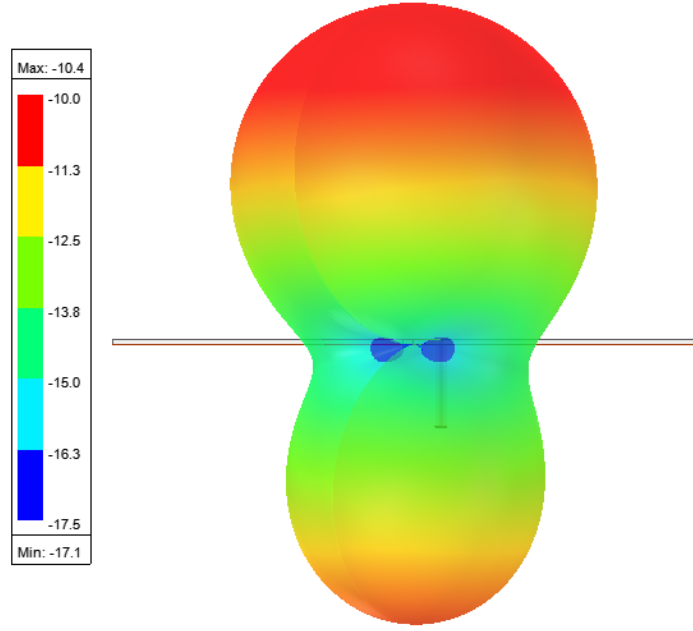


Figure 4.3: 60-mil-thick single MPA side view with radiation pattern and gain scale

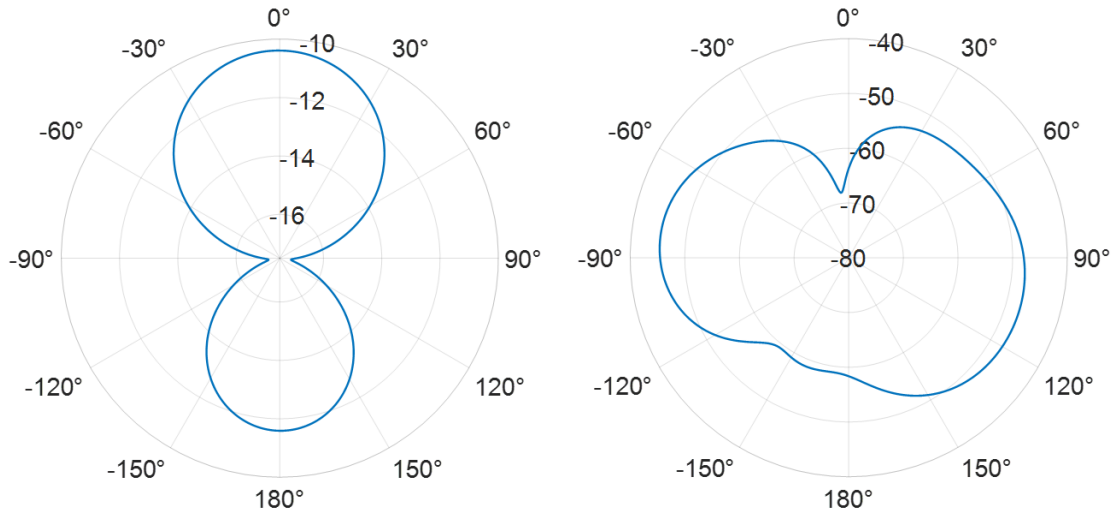
Single MPA: 120-mil-thick substrate

Comparing the 60-mil-thick design to the stacked configuration in section 3.6 is an unfair comparison given that the overall thickness of the stacked configuration is 120 mil. Thus, a 120-mil-thick single MPA is designed for a fairer comparison.

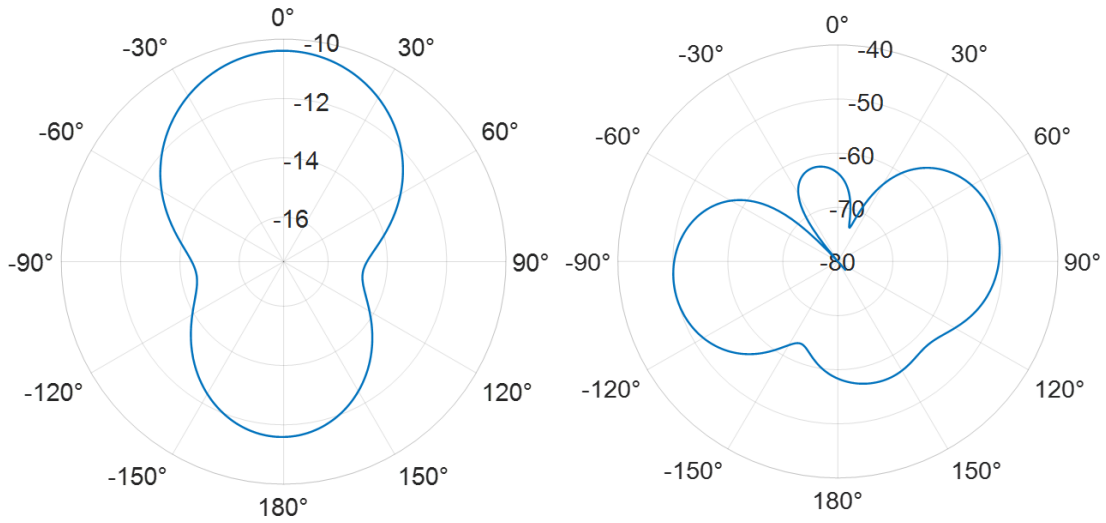
With the change in substrate thickness, the patch size and the feed location had to be tuned to get the resonance at 435 MHz. The new X_{pos} can be obtained using (4.1); however, it needs the new length which can be obtained using (3.13a).

After following the same approach shown for calculating the patch length in the 60-mil-thick substrate design and with further tuning, the patch length and width were found to be 62.484 mm, whereas the tuned feed position was found to be at (16.51 mm, 31.242 mm).

The antenna structure is still the same as in figure 4.1 but with twice the thickness. It can be seen that the gain improved by 3.6 dB, and the bandwidth broadened slightly by



(a) $\phi = 0$ degrees (H-plane)



(b) $\phi = 90$ degrees (E-plane)

Figure 4.4: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a 60-mil-thick single MPA

1.5 MHz, whereas the $|S_{11}|$ response is not as well matched as that of the 60-mil-thick substrate.

As can be seen from figures 4.5 and 4.6, the gain is now at -6.8 dBi with a resonance frequency of 435 MHz at -18.94 dB along with a bandwidth of 18.8 MHz

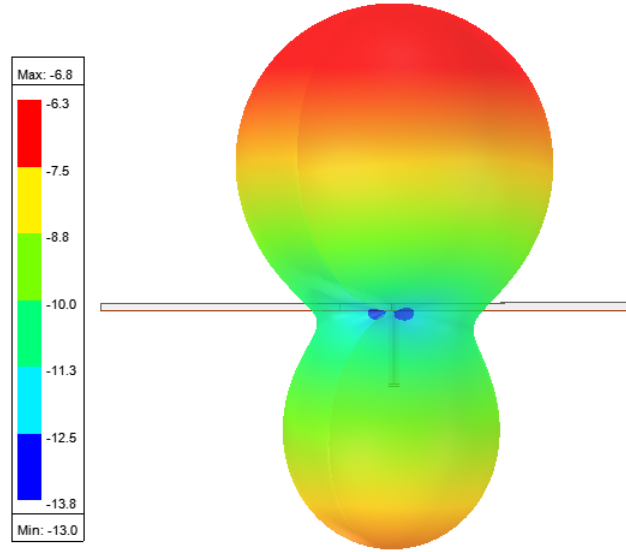


Figure 4.5: 120-mil-thick single MPA side view with radiation pattern and gain scale

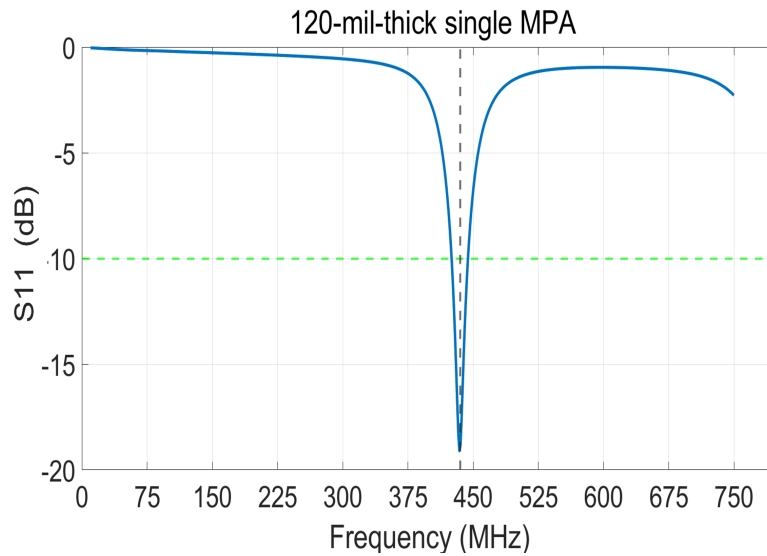


Figure 4.6: $|S_{11}|$ in dB plotted as a function of frequency (MHz)

from 425.31 MHz to 444.11 MHz, a fractional bandwidth of 4.32%. The increased thickness causes a slight bandwidth improvement which can still be maximized using a stacked configuration. Thus, further modifications are needed to explore the limitations of MAGTRES in terms of bandwidth as shown in section 4.2.

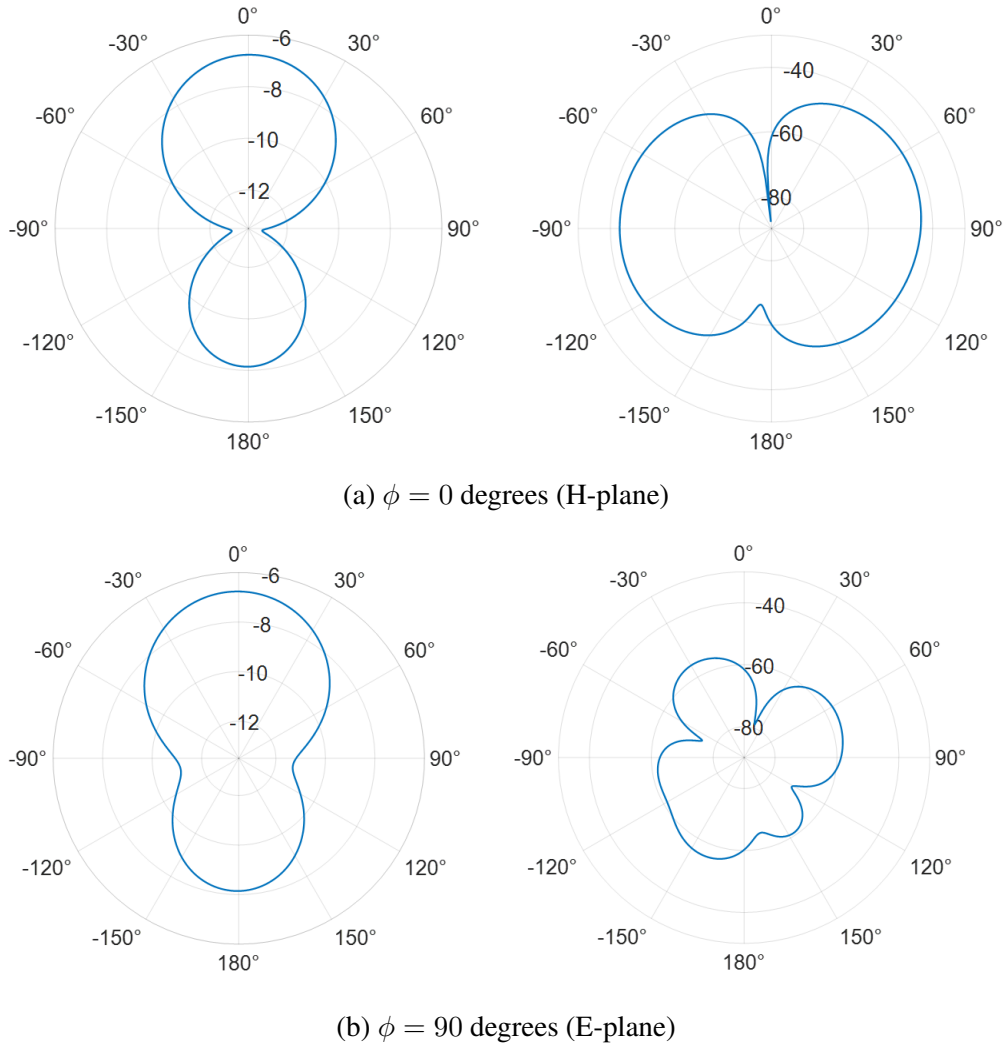


Figure 4.7: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a 120-mil-thick single MPA

From figure 4.7, the maximum elevation cross-polarization for the cut planes at $\phi = 0^\circ$ is -43.2 dBi and that of co-polarization is -6.8 dBi. Moreover, the maximum elevation cross-polarization for the cut planes at $\phi = 90^\circ$ is -55.8 dBi, and that of co-polarization is -6.8 dBi. This means that the cross-polarization is 36.4 dB below the desired main beam for $\phi = 0^\circ$ and 49 dB below the desired main beam for $\phi = 90^\circ$. Hence, the antenna effectively maintains the desired polarization while minimizing cross-polarization.

For a better comparison between the 60 and 120 mil substrates, figure 4.8 can be referred to. It can be seen that the bandwidth increased by 1.5 MHz. A larger patch of 1.143 mm is needed with doubled thickness to operate at the same resonance frequency. From this point forward, only the 120-mil-thick single MPA is used for future analysis and comparison.

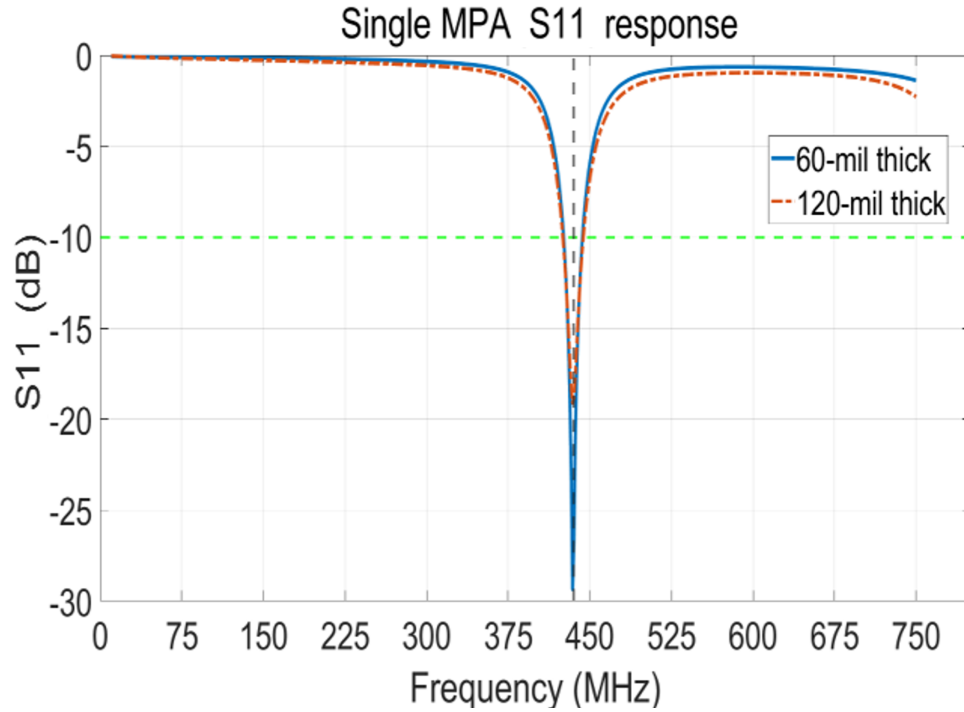


Figure 4.8: $|S_{11}|$ in dB plotted for two single MPAs of different thicknesses as a function of frequency (MHz)

4.1.2 Reference design

To study the limitations of the miniaturization provided by the MAGTREX material, a reference design is used upon which comparisons will take place. RO3006 by Rogers is used as a suitable reference design due to its dielectric constant of ~ 6 . The process to design this MPA is straightforward due to the availability of all the needed equations because $\mu_r = 1$ for this material.

Using (3.13a) with $c = 3 \times 10^8$ m/s, $f_{r010} = 435$ MHz, $\epsilon_r = 6$, and $\mu_r = 1$, a patch length of equal size to the patch width is obtained at 137.25 mm. As for the position of the feed, (3.8) was used. Given the following inputs: $L = W = 137.25$ mm and a calculated ϵ_{re} of 6.22 according to (3.2c). Feeding coordinates are obtained to be the following: $Y_{pos} = W/2 = 68.625$ mm whereas $X_{pos} = 50$ mm. The substrate dimensions and thickness are kept at $6'' \times 6''$ and 120 mil respectively.

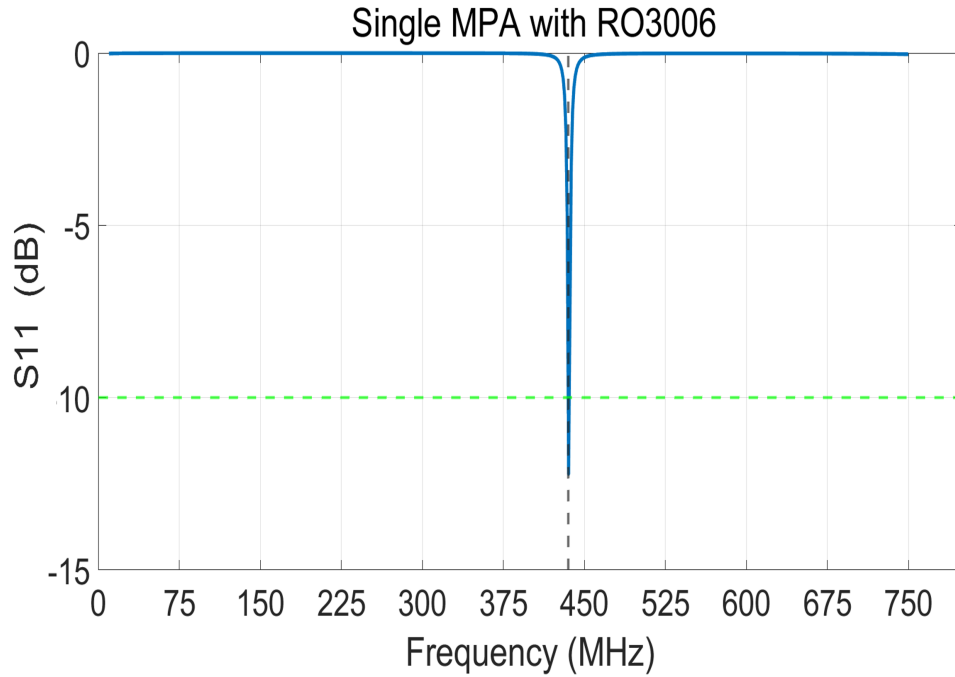


Figure 4.9: $|S_{11}|$ in dB plotted as a function of frequency (MHz)

Figure 4.9 shows a resonance frequency of 435 MHz at -12 dB along with a narrow bandwidth of 0.97 MHz from 435.13 MHz to 436.10 MHz, a fractional bandwidth of 0.22%. The bandwidth in this reference design is much narrower than that of the 120-mil-thick single MPA with a magneto-dielectric material and is not well-matched in comparison as seen in figure 4.10. Thus, compared to the reference design in a single MPA configuration, the magneto-dielectric material achieved 79.5% miniaturization, 1,844% increase in bandwidth, and provided better matching.

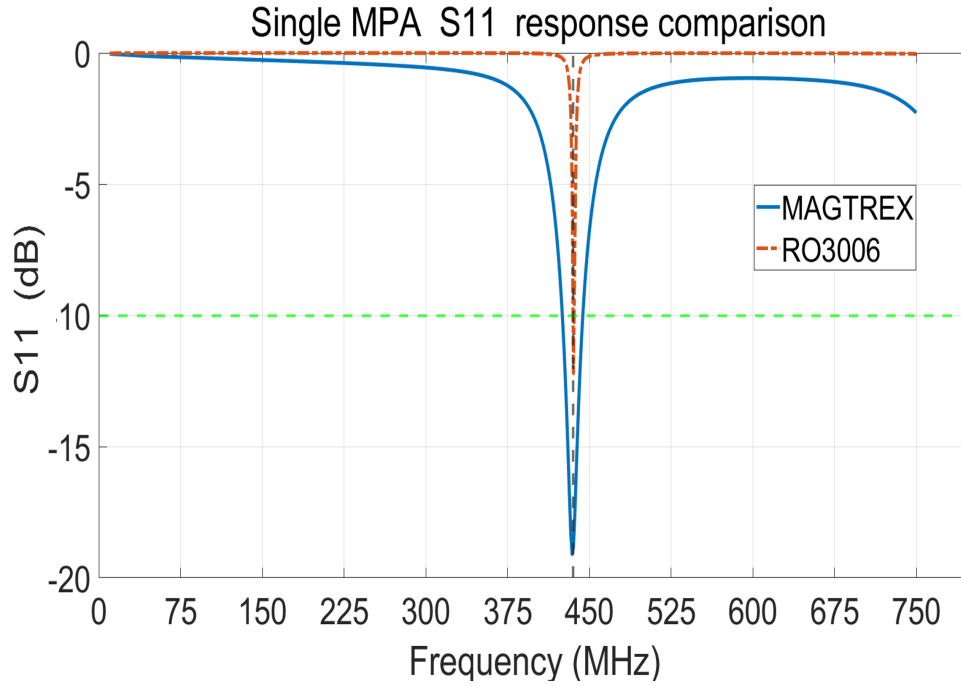


Figure 4.10: $|S_{11}|$ in dB plotted for two single MPAs, MAGTREX and RO3006, as a function of frequency (MHz)

Figure 4.11 shows how directive the antenna is through the radiation pattern and having a maximum gain of -0.7 dBi with significant back lobe radiation. This gain is yet to be radiated due to the finite ground plane that is chosen to show good comparison with the designed antenna. However, the non-magnetic nature of the substrate makes the magnetic loss tangent to be zero, which shows a significant increase in gain from -6.8 dBi to -0.7 dBi. Similar to the previous subsection, looking at the realized gain cut planes to see the cross-polarization in the antenna can help give more information about the antenna's efficiency.

From figure 4.12, the maximum elevation cross-polarization for the cut plane at $\phi = 0^\circ$ is -43.9 dBi and that of co-polarization is -0.7 dBi. Moreover, the maximum elevation cross-polarization for the cut plane at $\phi = 90^\circ$ is -52.7 dBi, and that of co-polarization is -0.7 dBi. This means that the cross-polarization is 43.2 dB below

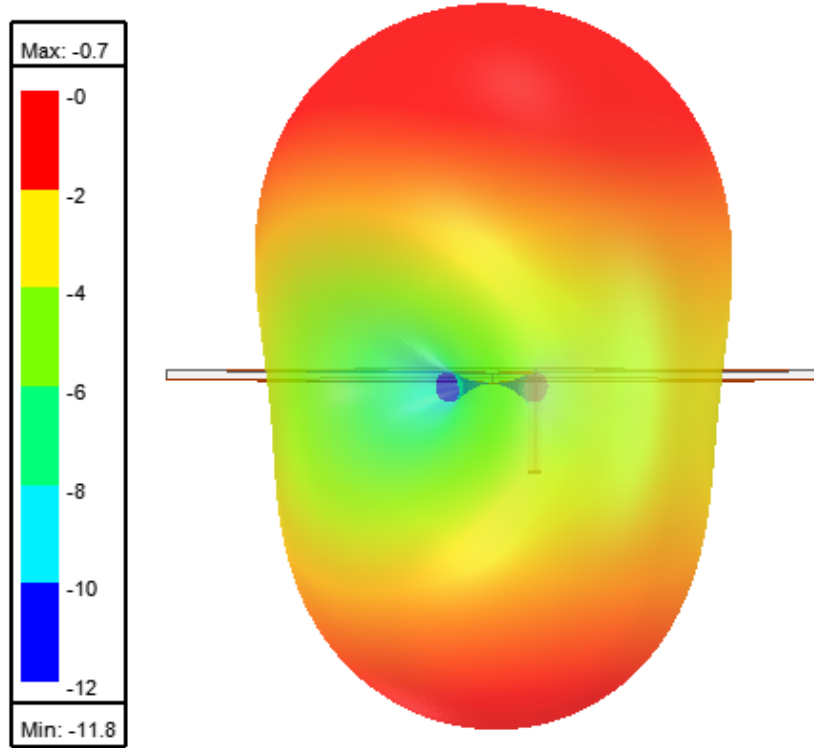


Figure 4.11: 120-mil-thick single MPA using RO3006 side view with radiation pattern and gain scale

the desired main beam for $\phi = 0^\circ$ and 52 dB below the desired main beam for $\phi = 90^\circ$. Hence, the antenna effectively maintains the desired polarization while minimizing cross-polarization.

Although this part shows a positive miniaturization result when the MAGTREX material is compared to that of a reference design, it is still not considered a good enough comparison due to the presence of ways to make patches smaller without the need for magneto-dielectric materials. Such ways need to be included in this analysis, which can be achieved using the Planar Inverted F-Antenna (PIFA).

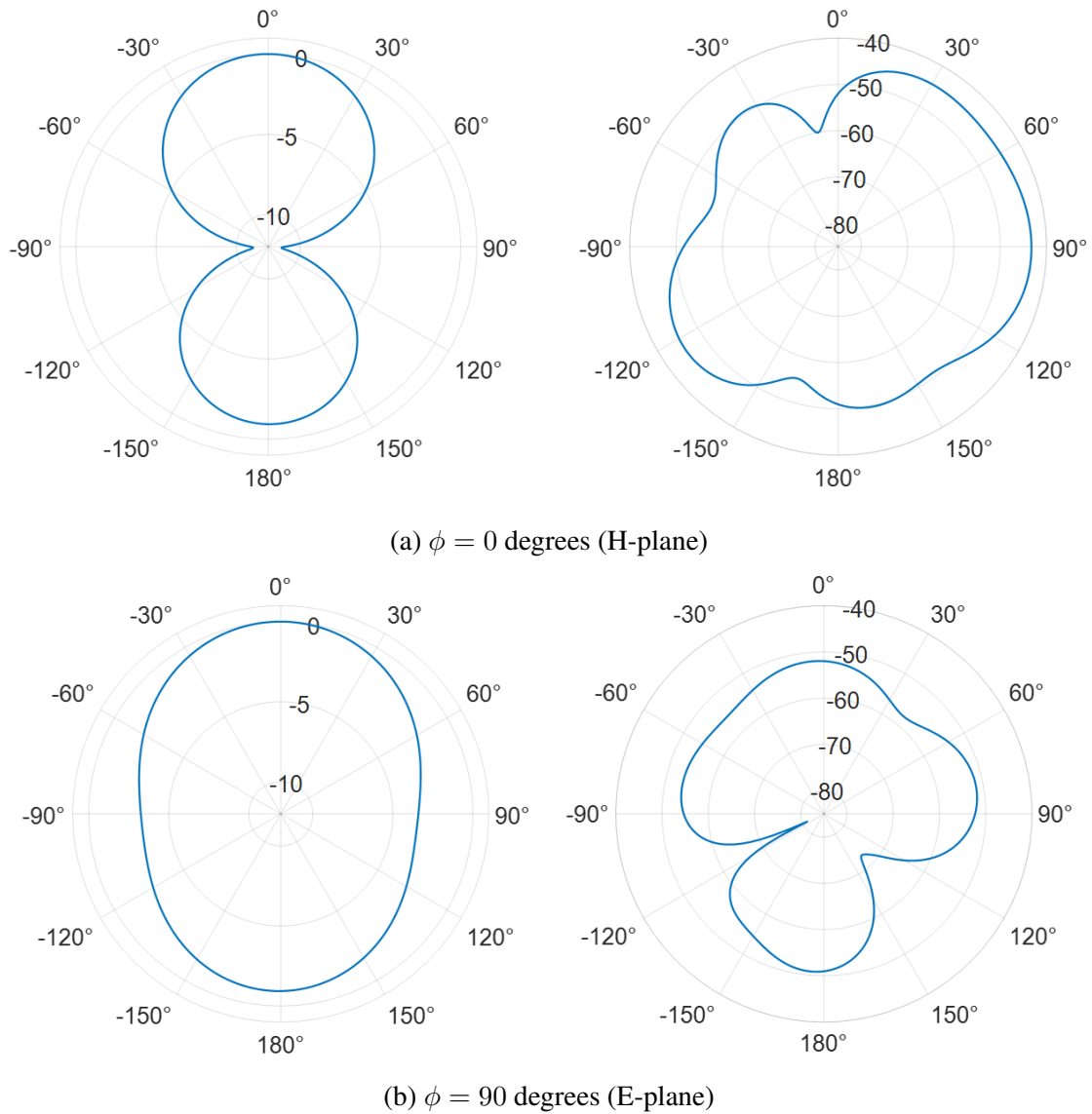


Figure 4.12: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a 120-mil-thick single MPA using RO3006

PIFA

PIFA is a type of antenna that is increasingly being used in the mobile phone market. It is resonant at a quarter-wavelength due to the shorting pin at the end. With the pin connecting the patch to the ground, the current at the edge of the patch is no longer forced to be zero, shifting the current-voltage distribution to be the same as a half-wave

patch antenna. This reduces the size by almost 50% while maintaining basic properties with the trade-off of a reduced gain [33]. Two cases are explored here: the first case is having a resonant PIFA with maximum radiation efficiency when the shorting pin in a plane of width $W = L1$. The second case is when W is just a pin with a width close to 0 where the resonance here occurs when the path length to the pin is $L1 + L2$. The second case cannot be designed while maintaining the basic properties, so the new second case is designed with $W = L1/2$ to have more miniaturization.

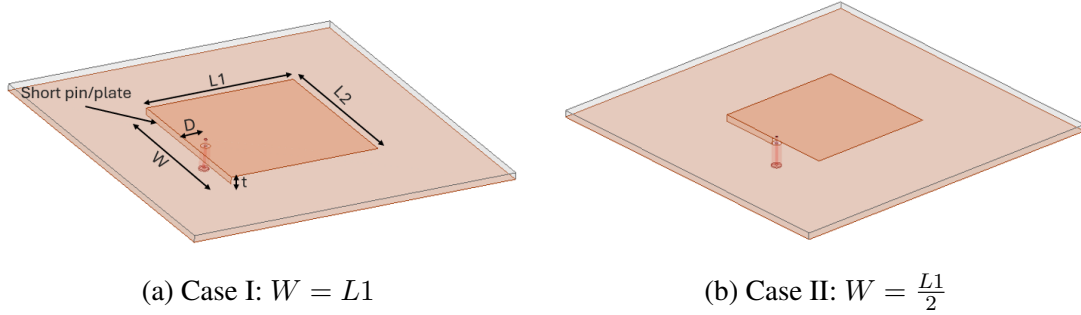


Figure 4.13: The planar inverted-f antenna (PIFA) layout

$$L1 + L2 - W = \frac{\lambda}{4} = \frac{c}{4f_r\sqrt{\epsilon_r}} \quad (4.2)$$

where $L1$ and $L2$ are the dimensions of the patch and W is the width of the shorting pin/plate. Figure 4.13 can be referred to for the PIFA's layout.

The design of both cases uses the following parameters: substrate: RO3006, $t = 120$ mil, $f_r = 435$ MHz, $c = 3 \times 10^8$ m/s, and $\epsilon_r = 6$. For the first case, $W = L1 = L2$ which after substituting in (4.2) gives $W = L1 = L2 = 70.39$ mm which becomes 69.5 mm after tuning. As for the distance of the feeding point, D , the closer it is to the shorting pin/plate, the smaller the impedance, and vice versa. The input impedance of the PIFA can be tuned by this parameter. After tuning, $X_{pos} = \frac{L1}{2} = 34.75$ mm and $Y_{pos} = D = 8$ mm. Following the same approach with the second case with $W = L1/2$

and $L1 = L2$, the new design parameters are the following: $L1 = L2 = 57.5$ mm, $W = 28.75$ mm, $X_{pos} = 28.75$ mm, and $Y_{pos} = D = 2$ mm.

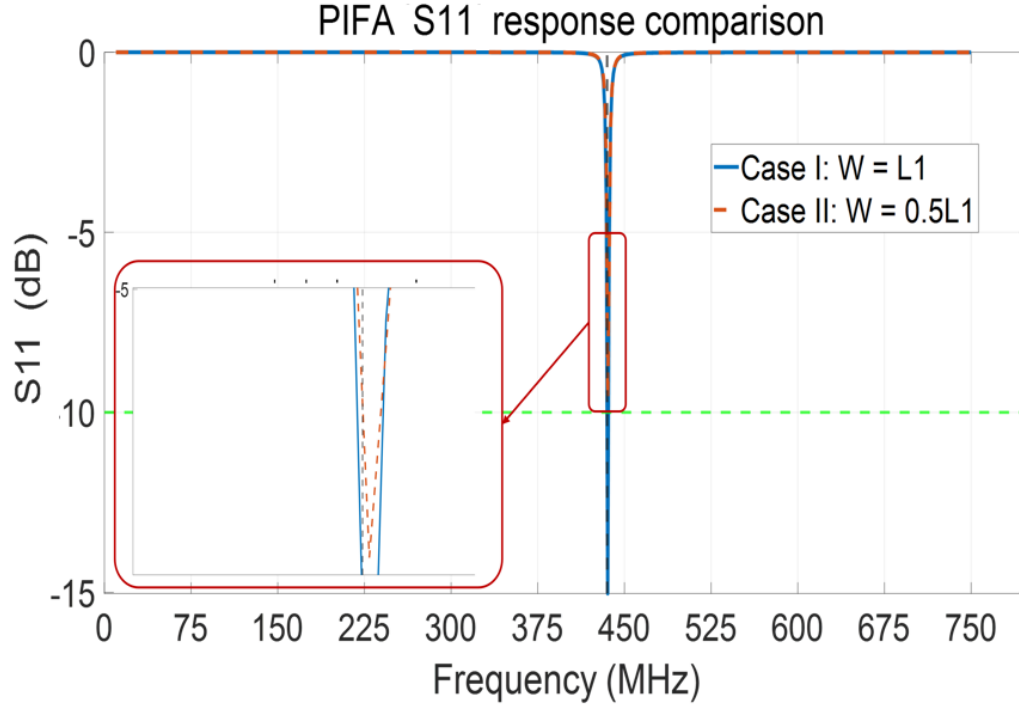


Figure 4.14: $|S_{11}|$ in dB plotted for a PIFA as a function of frequency (MHz)

Figure 4.14 shows a resonance frequency of 435 MHz for both cases. However, case II cannot be matched below $|S_{11}|$ of -10 dB whereas case I is at -14.99 dB with a 1.22 MHz BW from 434.93 MHz to 436.15 MHz, a fractional bandwidth of 0.28%. Both the matching and the bandwidth are worse with the PIFA compared to that of a MAGTRES design as shown in figure 4.15. With this being said, only the first case of PIFA will be considered.

Similar to the previous designs, there is significant back lobe radiation with the $6'' \times 6''$ ground plane but looking at the realized gain cut planes in figure 4.16 to see the cross-polarization in the antenna can help give more information about its efficiency. It can be seen that for case I, the maximum elevation cross-polarization for the cut planes at

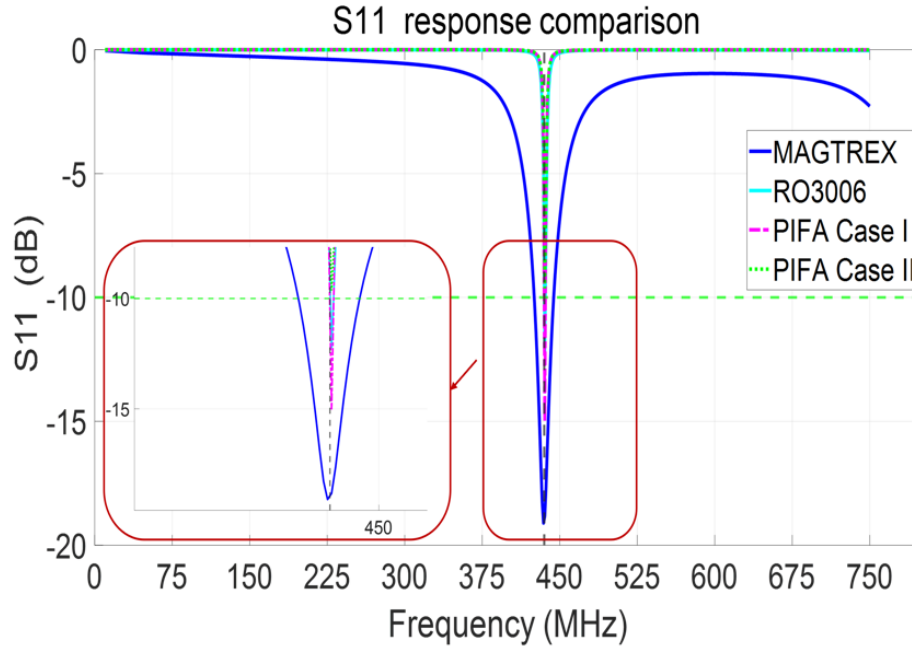
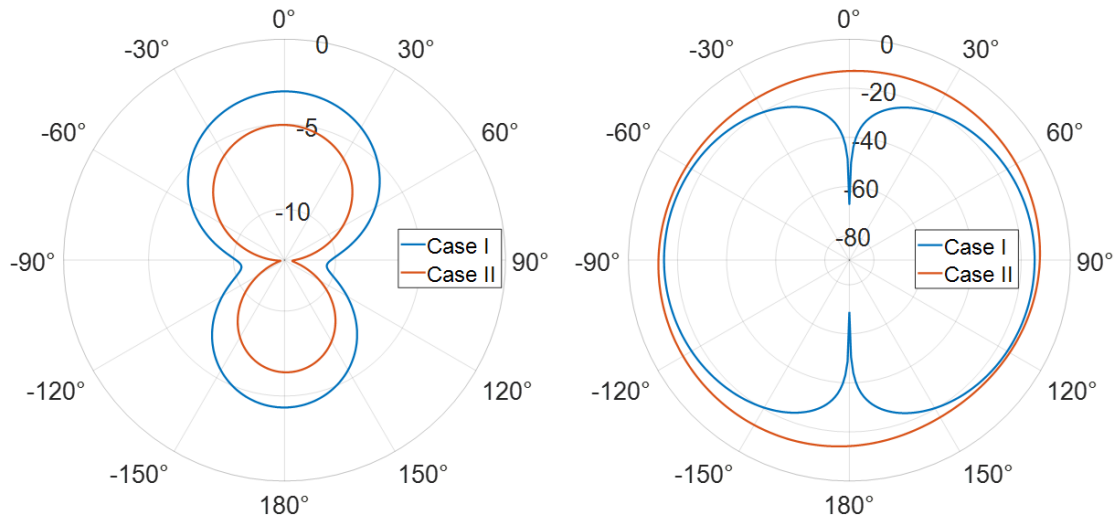


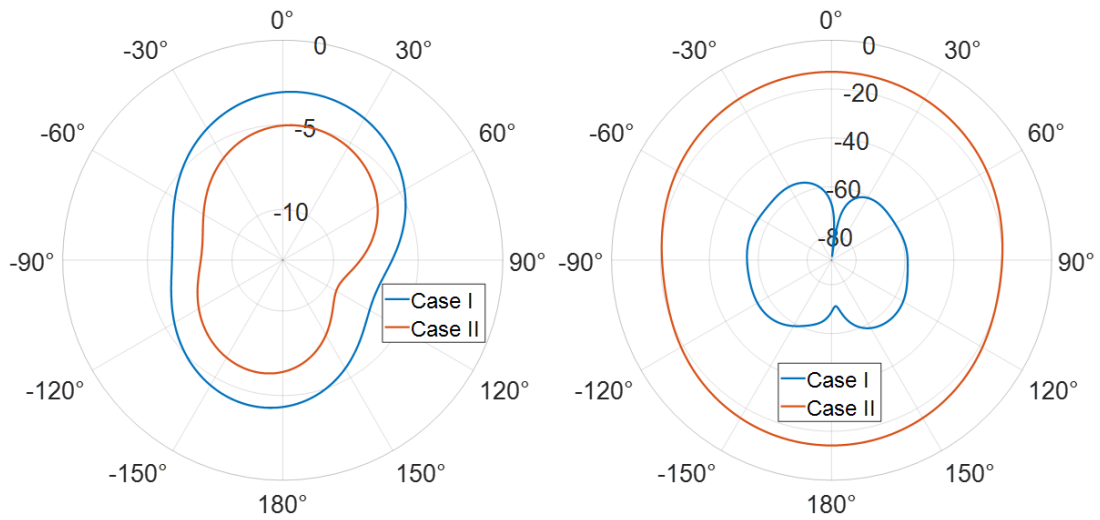
Figure 4.15: $|S_{11}|$ in dB plotted as a function of frequency (MHz) for antennas with a $6'' \times 6''$ ground plane

$\phi = 0^\circ$ is -14.5 dBi and that of co-polarization is -3.1 dBi. Moreover, the maximum elevation cross-polarization for the cut planes at $\phi = 90^\circ$ is -56.2 dBi, and that of co-polarization is -3.1 dBi. This means that the cross-polarization is 11.4 dB below the desired main beam for $\phi = 0^\circ$ and 53.1 dB below the desired main beam for $\phi = 90^\circ$. As for case II, it can be seen that the maximum elevation cross-polarization for the cut planes at $\phi = 0^\circ$ is -12.8 dBi and that of co-polarization is -6.3 dBi. Moreover, the maximum elevation cross-polarization for the cut planes at $\phi = 90^\circ$ is -14.3 dBi, and that of co-polarization is -6.3 dBi. This means that the cross-polarization is 6.5 dB below the desired main beam for $\phi = 0^\circ$ and 8 dB below the desired main beam for $\phi = 90^\circ$. Hence, the antennas maintain the desired polarization while minimizing cross-polarization but the isolation is minimal compared to the MAGTREX material and doesn't meet NASA's EcoSAR requirement.

To summarize what has been explored so far, table 4.1 can be referred to for antenna



(a) $\phi = 0$ degrees (H-plane)



(b) $\phi = 90$ degrees (E-plane)

Figure 4.16: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a 120-mil-thick single PIFA

performance comparison between the miniaturized antenna with magneto-dielectric material and PIFA based on the reference design. These comparisons explored the limitations of these antennas on a $6'' \times 6''$ ground plane; however, to get to the ideal performance that minimizes back lobe radiation, a ground plane of $11'' \times 11''$ is used because

the optimal distance is obtained by having the distance from the edge of the patch to the edge of the ground plane be between $\lambda/4$ and $\lambda/2$. The results of these findings are shown in table 4.2 where the main difference is in the realized gain values.

Table 4.1: Antenna performance comparison at 435 MHz with a $6'' \times 6''$ ground plane

Antenna	Return Loss (dB)	Realized Gain (dBi)	10dB BW (MHz)	Patch Size (cm ²)	X-pol Isolation (dB)
RO3006	12	-0.7	0.97	188.376	>30
PIFA	15	-3.1	1.22	48.301	<30
MAGTREX	19	-6.8	18.8	39.043	>30

Table 4.2: Antenna performance comparison at 435 MHz with a $11'' \times 11''$ ground plane

Antenna	Return Loss (dB)	Realized Gain (dBi)	10dB BW (MHz)	Patch Size (cm ²)	X-pol Isolation (dB)
RO3006	14	4.4	2.41	190.44	>30
PIFA	13.7	1.25	1.34	48.303	<30
MAGTREX	22	-1.6	21.29	39.043	>30

The data shows the trade-offs that come with using a lossy magneto-dielectric material with high magnetic and dielectric loss tangents. It can be seen that only the first case of PIFA shows acceptable results in terms of bandwidth. Compared to the reference design, PIFA achieved 74.63% miniaturization whereas MAGTREX achieved 79.3% miniaturization. Similarly, PIFA achieved a 25.77% increase in bandwidth whereas MAGTREX achieved an 1838.14% increase. This shows that a magneto-dielectric material offers better miniaturization while maintaining and improving antenna performance at the expense of a worst realized gain, -6.8 dBi compared to -3.1 dBi. However, although the gain is better in the PIFA, its cross-polarization isolation is less than that of

MAGTRES. Although the bigger ground plane helped fix the deficiency in the radiation performance, it was not enough to get the MAGTRES to a positive gain. A way to fix this is by using a phased array configuration, which is shown in section 4.3. Figures 4.17 and 4.18 can be referred to for the plots of the parameters summarized in table 4.2. It can still be seen that the broadest obtained bandwidth in this analysis is the MAGTRES's at 21.29 MHz. This is still far from achieving the bandwidth requirement of 120 MHz of NASA's EcoSAR antenna. Thus, further analysis is done in section 4.2 where a stacked configuration is designed to further explore the limitations of MAGTRES in terms of bandwidth.

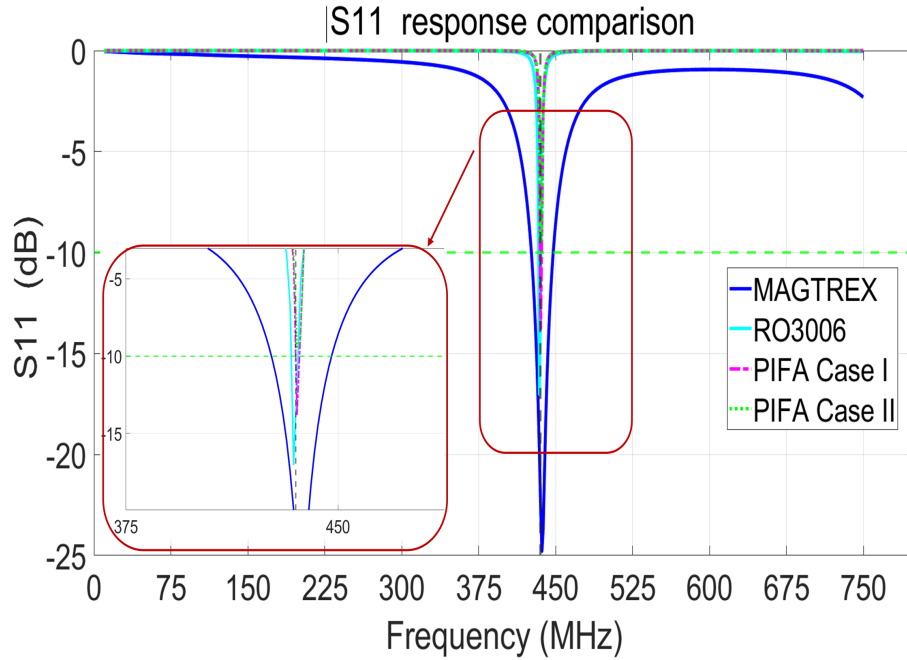
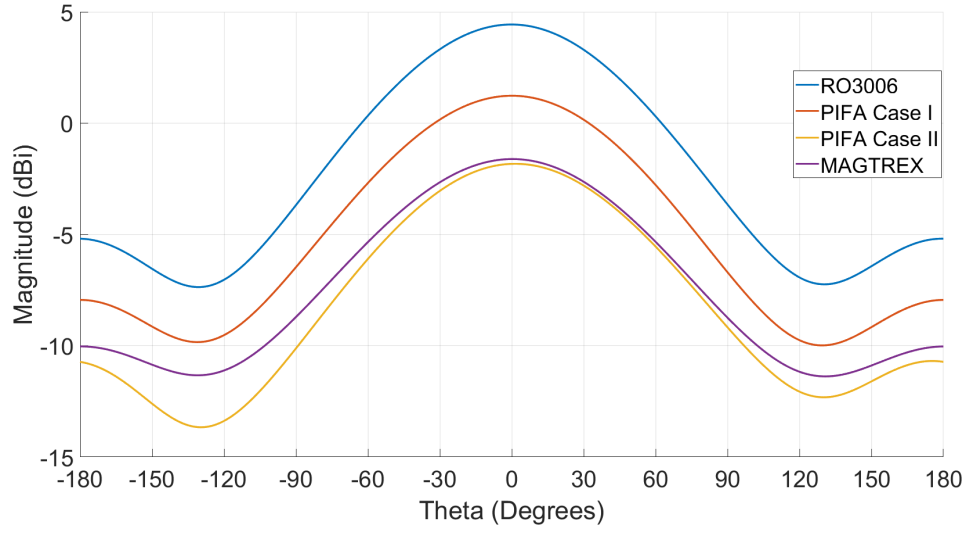
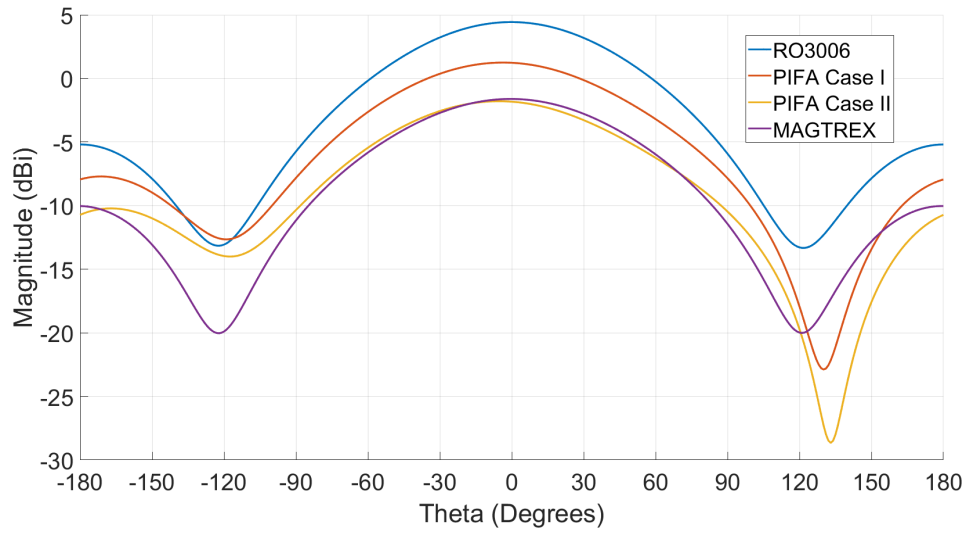


Figure 4.17: $|S_{11}|$ in dB plotted as a function of frequency (MHz) for antennas with an $11'' \times 11''$ ground plane



(a) $\phi = 0$ degrees (H-plane)



(b) $\phi = 90$ degrees (E-plane)

Figure 4.18: Elevation realized gain cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for the antennas designed with an $11'' \times 11''$ ground plane

4.2 Stacked Microstrip Patch Antenna

A stacked patch antenna is designed to broaden the bandwidth. The steps to design it are no different than that of a single patch antenna, where the top patch is designed by determining the effective dielectric constant for the top patch and then the resonant

length, using the same equations used for the bottom patch [8]. The size of the top patch will either be larger or smaller than the bottom patch, depending on the resonance frequency at which it is designed to resonate to achieve dual-band or broadband operation. Due to the unavailability of equations that include the effective magnetic constant, a slightly larger top patch will be started with. Different board thicknesses are also investigated to achieve the broadest bandwidth.

Regardless of thicknesses, patch sizes, and feed location, figure 4.19 shows the stack-up layers of the antenna.

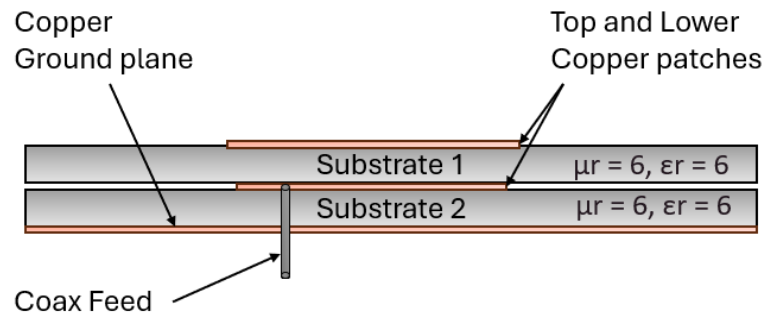


Figure 4.19: Microstrip patch antenna stack-up side view layout

Three main variables affect the antenna's behavior: substrate thickness, patch size, and feed location. As for the substrate thickness, the thicker the substrate, the broader the bandwidth and the less matched the return loss response gets at the resonant frequency. The bigger the patch size, the lower the resonant frequency. Changing the feed location along the y-axis will disturb the fields and produce more harmonic ripples in the response plot. On the other hand, changing the feed location along the x-axis helps control the return loss response due to the change in impedance. With this in mind, an exploration of limitations is shown.

Four main thicknesses have been investigated to see how the substrate thickness affects the behavior of the antenna: 20 mil, 40 mil, 60 mil, and 120 mil. Eigenmode

simulations have been conducted in HFSS to see how these thicknesses affect the resonance frequency of both patches. The lower patch is fed directly through a probe and controls the resonance frequency. The top patch is coupled indirectly through proximity coupling from the lower patch. Changing its size to slightly larger or smaller affects the antenna's bandwidth. Eigenmode simulations show that only 40-mil and 60-mil-thick lower substrates provide promising investigation results because the patches resonate at frequencies that can potentially span a 120 MHz bandwidth. Thus, 20-mil and 120-mil thick lower substrates are not considered in the investigation.

4.2.1 Substrate Thickness Selection

After deciding that only 40-mil and 60-mil thick lower substrates will be investigated, the HFSS modal network analysis mode is used to obtain simulation responses. Starting with a 40-mil thick lower substrate, figure 4.20 shows the $|S_{11}|$ response of a stacked microstrip patch antenna with different upper substrate thicknesses.

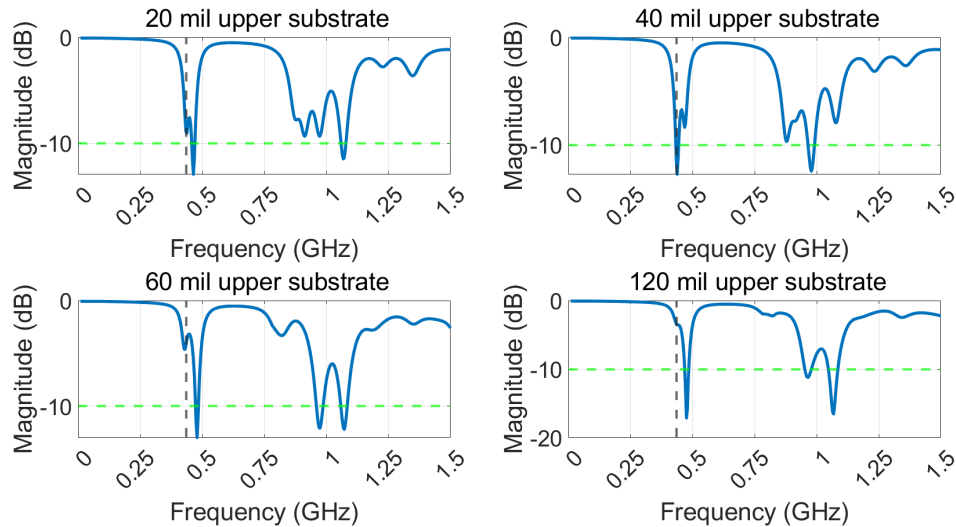


Figure 4.20: Microstrip patch antenna stack-up layers with 40-mil thick lower substrate and varying upper substrate thicknesses

Figure 4.20 shows that the antenna behaves differently with different upper substrate thicknesses. However, the bandwidth stays the same as the single patch antenna (10-20 MHz). Given that the desired bandwidth is 120 MHz, the 40-mil thick lower substrate will not be suitable for this thesis as it does not provide improvement compared to the single patch design. This confirms the results of the eigenmode simulation that excludes the 20-mil thick lower substrate from consideration. However, this process was not as simple as just changing the thickness of the upper substrate. Tuning has to be made for the patch sizes to get the best response of having both patches resonate at close enough frequencies, around 435 MHz. Once the best response is found, the feed location can be changed for better optimization. Before tuning the feed location, the initial feed position was kept at (16.51 mm, (lower patch length)/2). After tuning the results to get figure 4.20 that shows the best response (broadest bandwidth and closest resonant frequency of 435 MHz), the patch sizes were the following:

- *20 mil upper substrate:* 56.896 mm lower and upper patch.
- *40 mil upper substrate:* 56.413 mm lower and upper patch.
- *60 mil upper substrate:* 50.8 mm lower patch and 62.23 mm upper patch.
- *120 mil upper substrate:* 50.8 mm lower patch and 60.96 mm upper patch.

To make a better call about which thickness to pick, another investigation with a 60-mil thick lower substrate is implemented, which gives figure 4.21.

Now, with a 60 mil lower substrate, following a similar process, figure 4.21 displays the return loss response of a stacked microstrip patch antenna with different upper substrate thicknesses. The responses in each subfigure indicate that a 60-mil-thick lower substrate offers potential for optimization and tuning than a 40-mil-thick lower substrate. Up to this point, the feed location was maintained at (16.51 mm, (lower patch

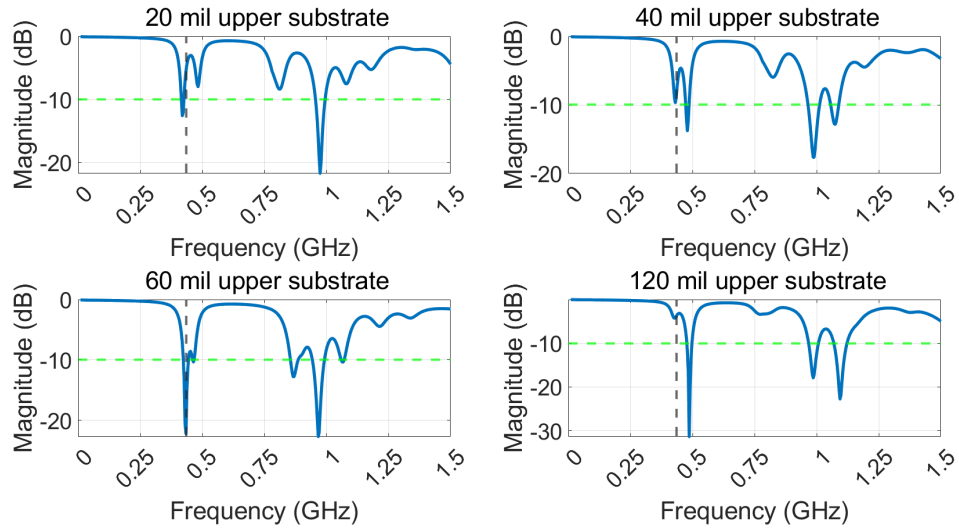


Figure 4.21: Microstrip patch antenna stack-up with 60 mil Thick lower substrate and varying upper substrate thicknesses

length)/2). After tuning, the results in figure 4.21 showcased the best response (broadest bandwidth and closest resonant frequency to 435 MHz). The patch sizes were as follows:

- *20 mil upper substrate:* 50.8 mm lower patch and 63.5 mm upper patch.
- *40 mil upper substrate:* 50.8 mm lower patch and 62.23 mm upper patch.
- *60 mil upper substrate:* 56.896 mm lower and upper patch.
- *120 mil upper substrate:* 49.53 mm lower patch and 62.23 mm upper patch.

From these two lists, it becomes apparent that the patch sizes do not conform to a specific trend, primarily due to the magnetic and lossy nature of the MAGTREX substrate. The feed location is yet to be tuned; this is a crucial part of the final design tuning, in subsection 4.2.2.

4.2.2 Final Design

After seeing the results of subsection 4.2.1, a 60 mil lower and upper substrate was chosen to explore the limitations of the MAGTREX material in a stacked configuration. The realized gain was obtained in figure 4.22 to be -13.4 dBi. Although stacked patches should show better gain, it is much worse than the single patch due to the stack's increased dielectric and magnetic losses. The electromagnetic waves pass through more lossy material with cumulative losses dissipating more energy as heat reducing the overall efficiency. Besides, the coupling efficiency between the patches is reduced as the intended energy for the upper patch gets lost in the substrate material.

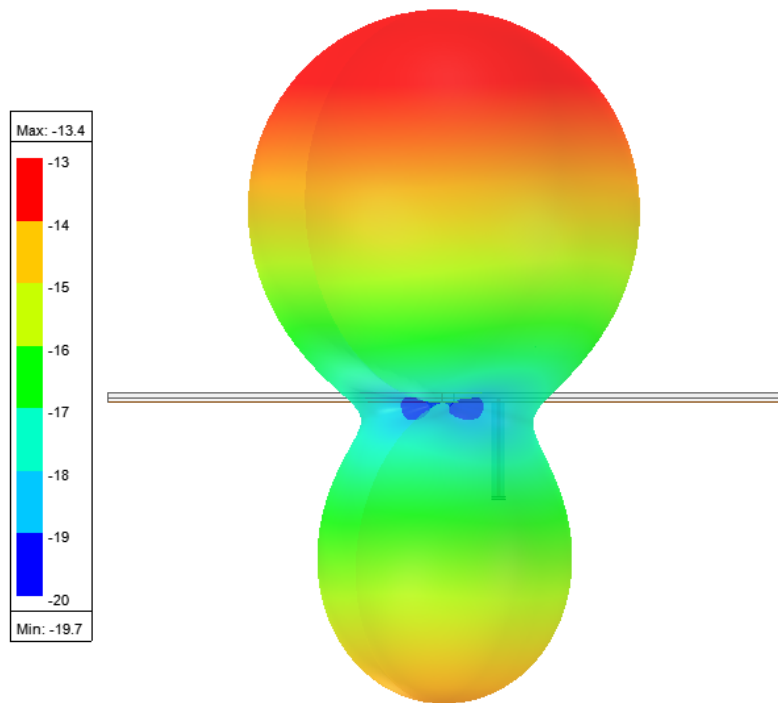


Figure 4.22: Microstrip stacked patch antenna side view and gain scale with middle feed configuration

To tune what was found in subsection 4.2.1 for better performance, the feed location was changed from (16.51 mm, (lower patch length)/2) to (6.096 mm, (lower patch

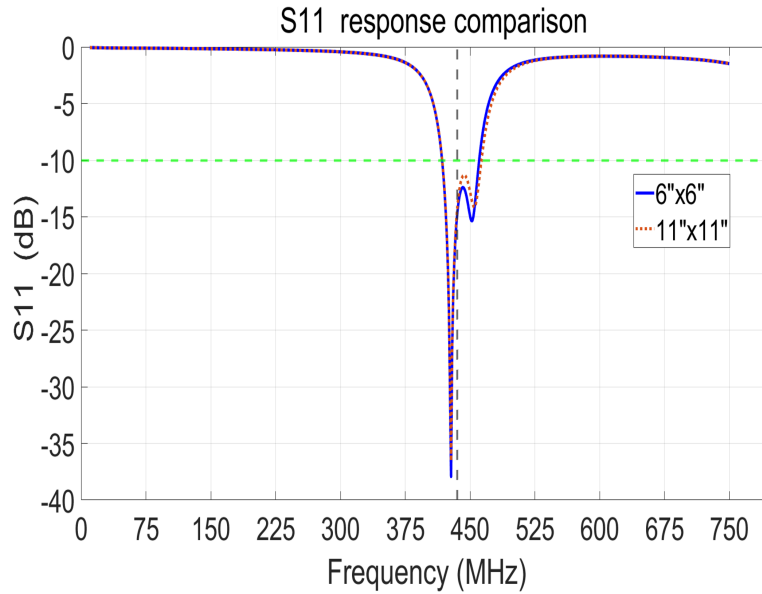
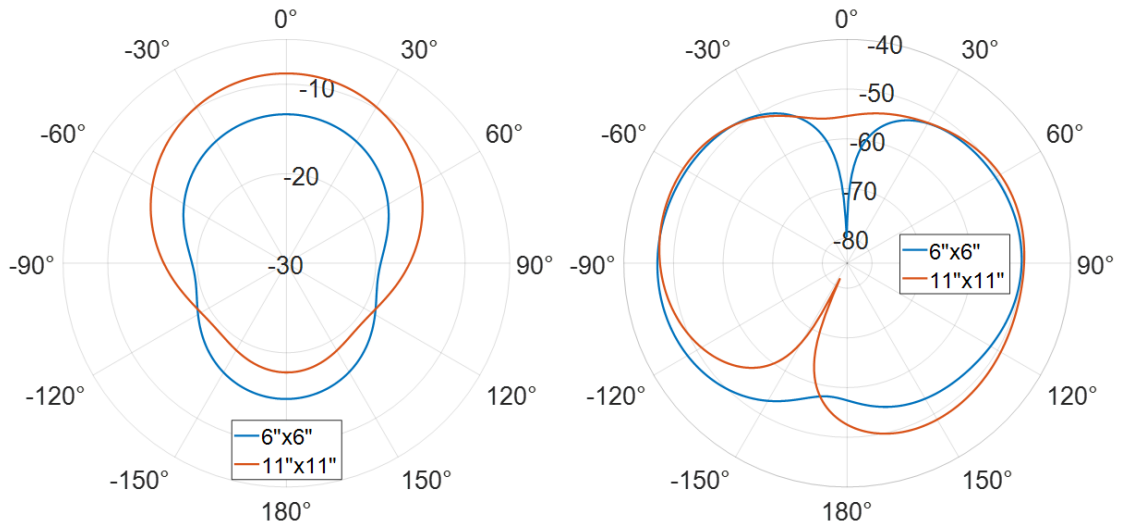


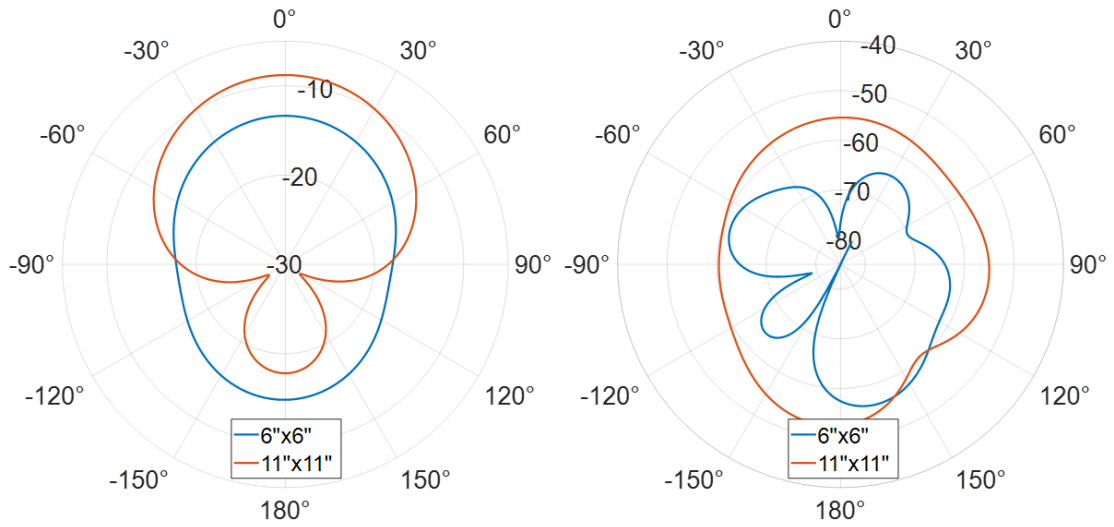
Figure 4.23: Return loss response of microstrip stacked patch antenna with middle feed configuration

length)/2) to match the network at the resonance frequency. Besides, the lower and upper patches were changed from 56.896 mm for both patches to 57.785 mm for the lower patch and 58.42 mm for the upper patch to center the bandwidth around the resonant frequency of 435 MHz. From figure 4.23, a maximum bandwidth of 42.97 MHz was obtained, a fractional bandwidth of 9.8%, with a resonant frequency of 435 MHz at -14.96 dB. The figure also compares the effect of a bigger ground plane, 11" $\times$ 11" on bandwidth.

Focusing on the 6" $\times$ 6" design from figure 4.24, the maximum elevation cross-polarization for the cut planes at $\phi = 0^\circ$ is -46.8 dBi and that of co-polarization is -13.4 dBi. Moreover, the maximum elevation cross-polarization for the cut planes at $\phi = 90^\circ$ is -56.0 dBi, and that of co-polarization is -13.4 dBi. This means that the cross-polarization is 33.4 dB below the desired main beam for $\phi = 0^\circ$ and 42.6 dB below the desired main beam for $\phi = 90^\circ$. This means that the antenna effectively maintains the desired polarization while minimizing cross-polarization.



(a) $\phi = 0$ degrees (H-plane)



(b) $\phi = 90$ degrees (E-plane)

Figure 4.24: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a stacked MPA with different ground plane sizes

Figure 4.25 shows a comparison between the stacked and single designs which shows that the bandwidth was successfully broadened from 18.8 MHz to 42.97 MHz. Thus, compared to the reference design, the bandwidth increased by 4320.62%. Although this is not close to NASA's EcoSAR antenna operating bandwidth of 120 MHz,

this is the broadest that can be achieved with the stacked magneto-dielectric test substrate provided by ROGERS due to its lossy nature using this thickness. As for the gain, section 4.3 explores ways to fix the negative gain whereas chapter 7 suggests future research that can be done. Table 4.3 can be referred to for antenna performance comparison.

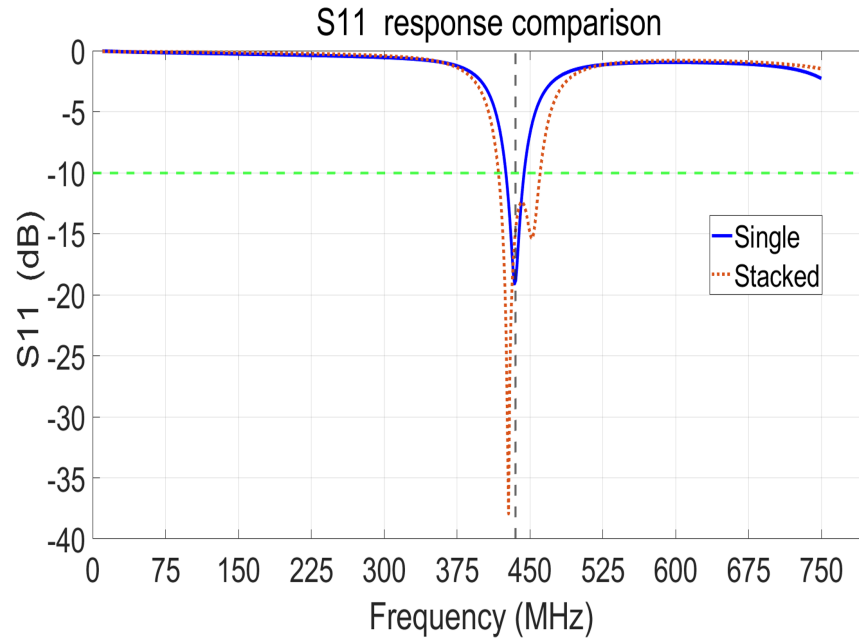


Figure 4.25: Return loss response comparison of single vs. stacked MPA

Table 4.3: Updated antenna performance comparison at 435 MHz with a $6'' \times 6''$ ground plane

Antenna	Return Loss (dB)	Realized Gain (dBi)	10dB BW (MHz)	Patch Size (cm ²)	X-pol Isolation (dB)
RO3006	12	-0.7	0.97	188.376	>30
PIFA	15	-3.1	1.22	48.301	<30
MAGTREX	19	-6.8	18.8	39.043	>30
Stacked MAGTREX	15	-13.4	43.0	33.391L 34.129U	>30

4.2.3 Coupling Coefficients

Coupling coefficients are a measure of the interaction between two resonant structures or elements, quantifying the degree of energy transfer between them. In the context of stacked MPAs, coupling coefficients are used to understand and optimize the resonance modes of the antenna system. They play a critical role in determining the bandwidth, efficiency, and overall performance of the antenna. Thus, obtaining the coupling coefficients helps better understand stacked MPAs. This parameter is investigated using the eigenmode simulation mode in HFSS where a thorough eigenmode investigation with four modes was followed using HFSS to obtain resonant frequencies and the quality factor. The objective is to have modes 1 and 2 resonate at 395 MHz, whereas modes 3 and 4 resonate at 475 MHz. This will span a 100 MHz bandwidth centered at the desired 435 MHz resonant frequency for the stacked antenna due to the roll-off from the resonant frequencies.

$$k_{i,j} = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2} \quad (4.3)$$

Similar to the previous investigation, the thickness of the lower substrate is fixed while the upper substrate is changed. Lower substrates with thicknesses 40 and 60 mil are used, and $k_{1,3}$ and $k_{2,4}$ is plotted as a function of varying thicknesses of the upper substrate. Upon investigation, the quality factor is obtained to be fixed at $Q = 16.808$ across all variations. This means that the only variable in this analysis is the thickness of the upper substrate because that of the lower substrate is fixed in each analysis, with feed location at (6.096 mm, (lower patch length)/2) from the final design obtained in 4.2.2.

For figure 4.26, the patch sizes (L and W) were the following:

- *20 mil upper substrate:* 56.896 mm lower and upper patch.

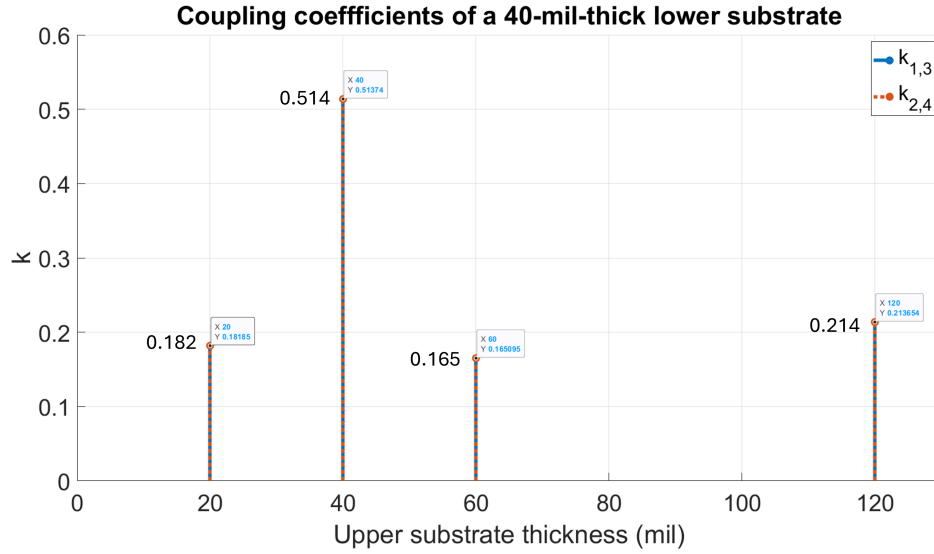


Figure 4.26: Coupling coefficients for a stacked MPA with a fixed 40-mil-thick lower substrate as a function of varying upper substrate thickness

- *40 mil upper substrate*: 53.34 mm lower patch and 62.23 mm upper patch.
- *60 mil upper substrate*: 53.34 mm lower patch and 62.23 mm upper patch.
- *120 mil upper substrate*: 50.8 mm lower patch and 60.96 mm upper patch.

For figure 4.27, the patch sizes (L and W) were the following:

- *20 mil upper substrate*: 50.8 mm lower patch and 63.5 mm upper patch.
- *40 mil upper substrate*: 50.8 mm lower patch and 62.23 mm upper patch.
- *60 mil upper substrate*: 49.53 mm lower patch and 62.23 mm upper patch.
- *120 mil upper substrate*: 49.53 mm lower patch and 60.96 mm upper patch.

To reiterate, Coupling coefficients represent the efficiency of energy transfer between different patches by quantifying how effectively the electromagnetic energy is coupling from one patch to another. When patches are well coupled, multiple resonant modes can be excited, which can help enhance the bandwidth and the radiation pattern.

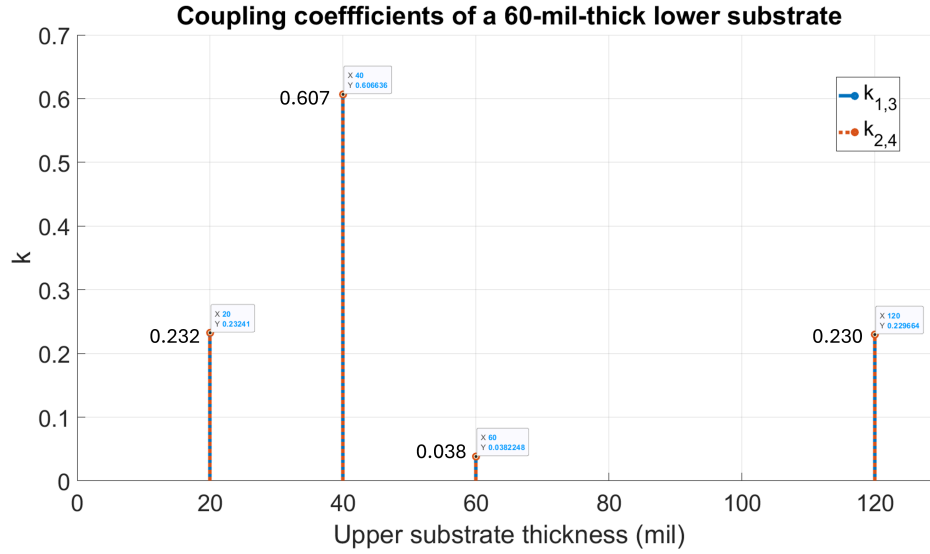


Figure 4.27: Coupling coefficients for a stacked MPA with a fixed 60-mil-thick lower substrate as a function of varying upper substrate thickness

The thicker the substrate, the weaker the coupling will be due to the more extensive distribution of the electromagnetic fields, the changing effective dielectric constant, and the increased loss tangent, which will influence how much energy is dissipated as heat rather than couple into the other patch [34].

As for the case with figures 4.26 and 4.27, the initial increase in substrate thickness seems to improve coupling by enhancing the field overlap. However, the fields become too spread out, with increasing thicknesses reducing coupling efficiency. Besides, after the limit for improving coupling is reached with increasing thickness, it becomes a question of mode matching as well as standing wave patterns, as increasing thicknesses cause divergence in the resonance frequency, which will not always lead to mode coupling alignment [35]. With a high coupling coefficient, an antenna will exhibit higher energy transfer efficiency, broader bandwidth, better impedance matching, and higher radiation efficiency. The typical coupling coefficient values are < 0.1 for low coupling, > 0.1 and < 0.3 for moderate coupling, and > 0.3 for high coupling. The analysis sug-

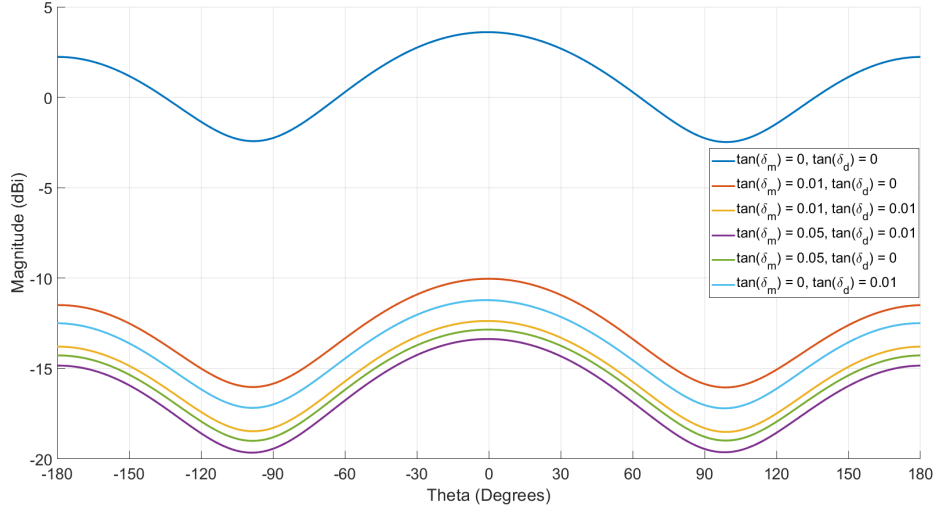
gests that the 60-mil lower substrate with a 40-mil upper substrate (60-40) might be ideal based on coupling coefficients. However, practical challenges arise when attempting to bring the resonances closer in this configuration. In the 60-40 setup, achieving proper resonance alignment without disintegration of the resonant modes was not possible due to the constraints imposed by the substrate thicknesses and the physical properties of the patch antenna. On the other hand, the 60-60 configuration, while exhibiting weaker coupling coefficients, provided a more stable and predictable behavior for aligning the resonances at the target frequencies. This stability is critical in ensuring the antenna's performance, even if the coupling coefficients are not optimized for maximum energy transfer. Therefore, the 60-60 configuration was chosen to balance resonance alignment and practical design feasibility.

In addition, the next section investigates the gain to see why the obtained gain has been negative for all the antennas and how it can be improved.

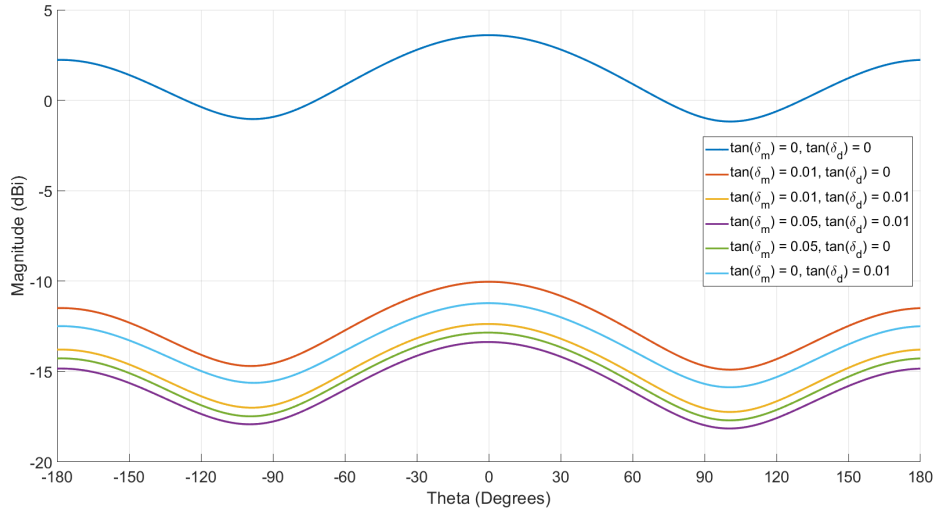
4.3 Gain Investigation

The three antenna designs with MAGTREX have shown a negative gain. The primary factor contributing to this result is the lossy nature of the MAGTREX material, which exhibits a dielectric loss tangent of 0.01 and a magnetic loss tangent of 0.05 under 500 MHz, according to the datasheet. While these material losses are significant, the patch antenna's ohmic losses from the conductor are not the main reason for the negative gain, as the antenna could radiate effectively if only ohmic losses were present. Instead, the combination of dielectric, magnetic, and conductor losses collectively reduces the efficiency and, consequently, the gain. Despite this, the directive nature of the antenna is evident, as shown by the simulation results: a directivity of 3.4 dB for the 60-mil-thick and 120-mil-thick single MPAs and 3.5 dB for the stacked MPA.

While it is directive, it is still below what is expected of microstrip patch antennas, as discussed in section 2.3.4, which is supposed to range from 5 to 9 dB. In this case, two ways exist to fix directivity and improve gain. The first approach is to consider the effect of a larger ground plane, representing an equivalent infinite ground plane compared to the patch size. The second approach is to rely on phased array functionality to compensate for the low gain. However, before investigating these two approaches, an analysis of the MAGTREX losses needs to be made to see what's contributing the most between both loss tangents. The results of this analysis can be summarized in table 4.4 and figure 4.28 where the contributions from the loss tangents are shown. It can be seen that they are both contributing equally to the lossy nature of the MAGTREX material. The plots show that a magnetic loss tangent of 0.05 has a greater role in making a substrate lossy than a magnetic and dielectric loss tangent of 0.01 each.



(a) $\phi = 0$ degrees (H-plane)



(b) $\phi = 90$ degrees (E-plane)

Figure 4.28: Elevation realized gain cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for the stacked MPA with different tangent loss values

4.3.1 HFSS Fields Calculations

Before exploring methods to compensate for the low gain, an analysis using the HFSS field calculator is conducted to gain a deeper understanding of the power distribution in the stacked MPA.

Table 4.4: Effects of the loss tangents on the realized gain of the stacked MPA with MAGTRES

Magnetic loss tangent	Dielectric loss tangent	Realized Gain (dBi)
0.05	0.01	-13.4
0.01	0.01	-12.4
0	0.01	-11.2
0.05	0	-12.8
0.01	0	-10.04
0	0	3.6

Equation (4.4) illustrates the calculation of total power in an antenna, where P_T represents the total input power (set to 1), P_d denotes the power stored in the dielectric substrates, P_c represents the power stored in the conductors, P_r is the radiated power, and P_{sw} is the power dissipated as surface waves. To explore this equation further, (4.5), (4.6), and (4.7) can be referenced [36].

$$P_T = P_d + P_c + P_r + P_{sw} = 1 \text{ W} \quad (4.4)$$

$$P_d = \frac{\omega\epsilon\delta_d}{2} \iiint |\vec{E}|^2 dv + \frac{\omega\mu\delta_m}{2} \iiint |\vec{H}|^2 dv = 0.61953 \text{ W} \quad (4.5)$$

$$P_c = 2 \frac{R_s}{2} \iint |\vec{H}|^2 = 0.005223 \text{ W} \quad (4.6)$$

$$P_r = \frac{1}{2} \int_0^{2\pi} \int_0^\pi \frac{|\vec{E}|^2}{2\eta_0} r^2 \sin(\theta) d\theta d\phi = 0.0204 \text{ W} \quad (4.7)$$

Where R_s is the surface resistance of copper which is denoted by the equation $(\omega\mu/\sigma)$ that equals $5.92 \times 10^{-5} \Omega$, σ is the electrical conductivity of copper, and $P_{sw} = P_T - (P_d + P_c + P_r) = 0.3548 \text{ W}$. This shows that most of the input power is lost in the dielectric and as surface waves instead of being radiated into space. To visualize that in a pie chart, figure 4.29 can be referenced.

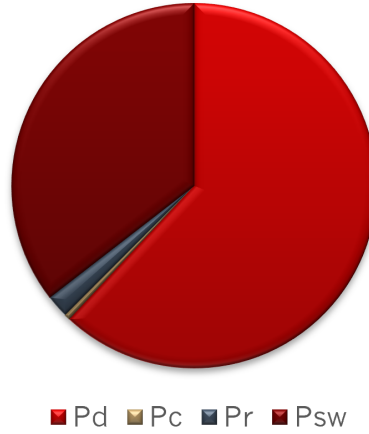


Figure 4.29: Pie chart visualizing different power allocations for the designed stacked MPA in subsection 4.2.2

To further support figure 4.29, a mathematical calculation can be done as shown in (4.8). This shows that the antenna is inefficient with an efficiency of 2.08%.

$$e\% = \frac{P_r}{P_T - P_r} \times 100\% = 2.08\% \quad (4.8)$$

4.3.2 Infinite Ground Plane Approach

To understand the effect of ground plane size on the behaviors of antennas, five different ground plane areas were considered: 36 square inches, 81 square inches, 121 square inches, 256 square inches, and 400 square inches. Patch dimensions and substrate height were not modified. Upon investigating the three boards, figures 4.30, 4.31, and 4.32 were obtained which can all be summarized in table 4.5.

Directivity, gain, and realized gain were significantly improved by increasing the ground plane size; however, gain and realized gain do not exhibit a linear response, unlike directivity. Directivity improved by falling in the expected range of 6 to 7 dB. Although gain and realized gain improved drastically for the stacked microstrip patch antennas (-13.4 to -8.35 dB), it is still negative, meaning that the antenna's conductors

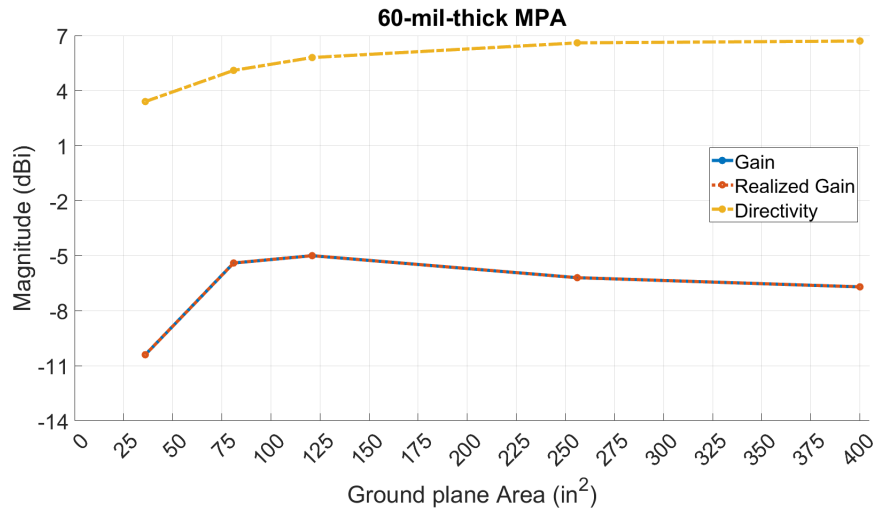


Figure 4.30: Radiation parameters pattern of a 60-mil-thick MPA with different ground plane areas

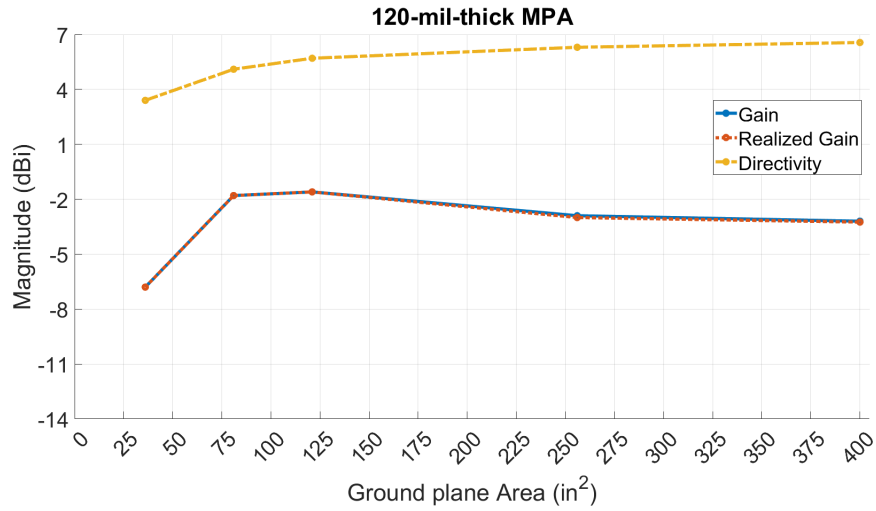


Figure 4.31: Radiation parameters pattern of a 120-mil-thick MPA with different ground plane areas

still exhibit significant loss.

The reason behind this improvement is the effect of the finite ground plane in the original $6'' \times 6''$ configuration. It does not account for radiation diffraction in the edge as the distance between the ground plane edge and the edge of the radiating patch is less than a quarter a wavelength [37]. The small distance causes surface wave generation,

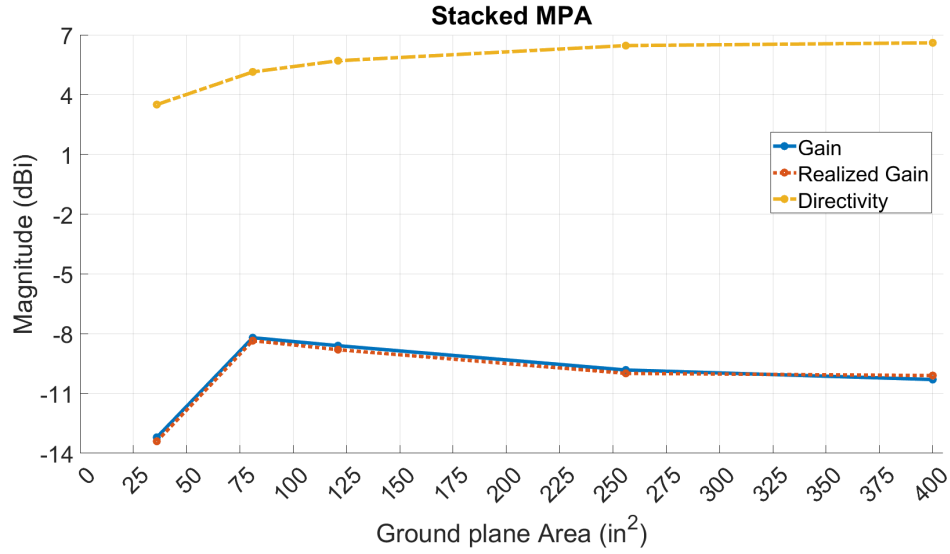


Figure 4.32: Radiation parameters pattern of a stacked MPA with different ground plane areas

Table 4.5: Effects of different ground plane sizes on the radiation patterns (dBi) of the three designed MPAs

Antenna	Parameters	6" × 6"	9" × 9"	11" × 11"	16" × 16"	20" × 20"
60-mil-thick MPA	Gain	-10.4	-5.4	-5	-6.2	-6.7
	Realized Gain	-10.4	-5.4	-5	-6.2	-6.7
	Directivity	3.4	5.1	5.8	6.6	6.7
120-mil-thick MPA	Gain	-6.8	-1.8	-1.6	-2.9	-3.2
	Realized Gain	-6.8	-1.8	-1.6	-3	-3.25
	Directivity	3.4	5.1	5.7	6.3	6.56
Stacked MPA	Gain	-13.2	-8.2	-8.6	-9.82	-10.3
	Realized Gain	-13.4	-8.35	-8.8	-9.99	-10.1
	Directivity	3.5	5.14	5.7	6.46	6.6

which leads to distortions in the radiation pattern, resulting in both side and back lobes.

A finite ground plane is often associated with additional losses due to edge diffractions and high reflection coefficients affecting the antenna gain, which affects the radiation pattern that gets directed in unwanted directions. Thus, by increasing the size of the ground plane and making it act as an infinite ground plane, the radiation pattern will improve, which fixes gain and directivity due to the eliminated diffractions at the

edge. The pattern is also improved due to stable impedance as the ground plane acts as an ideal reflector satisfying ideal boundary conditions where the tangential electric field on the ground plane is zero, minimizing impedance variations. Reducing surface waves means more power is radiated in the main lobe[35].

Besides, the variation in gain and realized gain weren't linear as expected as different edge diffractions occur once the ground plane's size becomes larger than what's required. The sweet point is when the distance between the edge of the patch and that of the ground plane is between $\lambda/4$ and $\lambda/2$ which also gets affected by the added losses of the substrate. An example of this is that both of the single MPAs showed the best gain result with a $9'' \times 9''$ ground plane whereas the stacked MPA showed the best gain result with an $11'' \times 11''$ ground plane. Once the sweet point is passed, the reflected waves in the substrate will add constructively or destructively affecting the overall radiated energy.

4.3.3 Phased Array Approach

The other approach to compensate for the negative values in the radiation patterns is the implementation of a phased array structure due to their various advantages, such as increasing gain, which is needed here. The reason behind that is the ability to direct most of the energy in a specific wanted direction as stated in subsection 3.5.1 [20].

Before designing the phased array structure, it has to be mentioned that a stacked configuration behaves less efficiently than a single configuration due to the presence of two patches that are coupled in a way that reduces efficiency and degrades performance, lowering the gain. The upper substrate is causing more losses to be added, causing energy dissipation as heat rather than being transmitted into space. Thus, for MAGTREX, having a stacked microstrip patch antenna increases the losses in the structure, which

affects gain. One way to compensate for that is to implement a phased array system that would be able to replace NASA's EcoSAR antenna. The phased array configuration of NASA's antenna can be shown in fig 4.33. Seeing the 1×3 square meter space where the phased array can be installed, five different configurations will be implemented with the results shown in table 4.6.

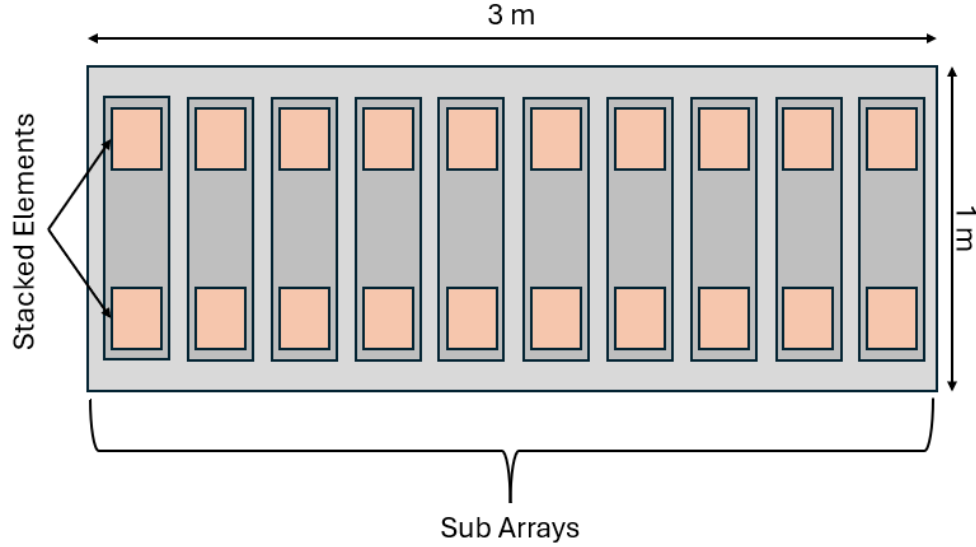


Figure 4.33: Layout of NASA's EcoSAR antenna showing 10 sub-arrays and 20 total elements

Table 4.6: Effects of element spacing on realized gain and degree of steering for a stacked MPA with MAGTREX

Array Size	Spacing	Realized Gain (dBi)	Degree of Steering
2×8	0.5λ	0.4	$\pm 45^\circ$
5×17	0.25λ	6.8	$\pm 45^\circ$
11×34	0.125λ	11.5	$\pm 45^\circ$
15×45	0.095λ	10.0	$\pm 45^\circ$

For table 4.6, the realized gain is taken from the peak of the main lobe whereas the degree of steering is calculated from the equation of maximum element spacing (4.9). This equation shows that a θ_{max} of 90° ($\pm 45^\circ$) is achievable with all the chosen

element spacing in table 4.6. This is mainly because grating lobes that would limit this steering angle appear at $d > 0.5\lambda$. Then, to visualize the radiation patterns of table 4.6, figure 4.35 can be referred to. The figures show that the side-lobe level is at 13.5 dB for all the arrays of table 4.6 providing flexibility in choosing the desired design without compromising side-lobe performance. Besides, the drop in gain beyond 11×34 elements is due to mutual coupling (MC) between the array elements. MC can lead to increased losses when elements are very close, affecting the radiated power as it is absorbed or scattered by neighboring elements, reducing the total radiated power [8].

Figure 4.34 illustrates the density of the MAGTREX array, which can be compared to the EcoSAR system shown in Figure 4.33. With the denser array presented as a potential replacement, the weight of the array is successfully reduced by 66.82%, from 200 lbs to 66.3 lbs. Additionally, it is important to note that the small element spacing, $d \ll 0.5\lambda_0$, in such a dense array results in a narrower beamwidth. To compensate for this, an amplitude synthesizer can be employed.

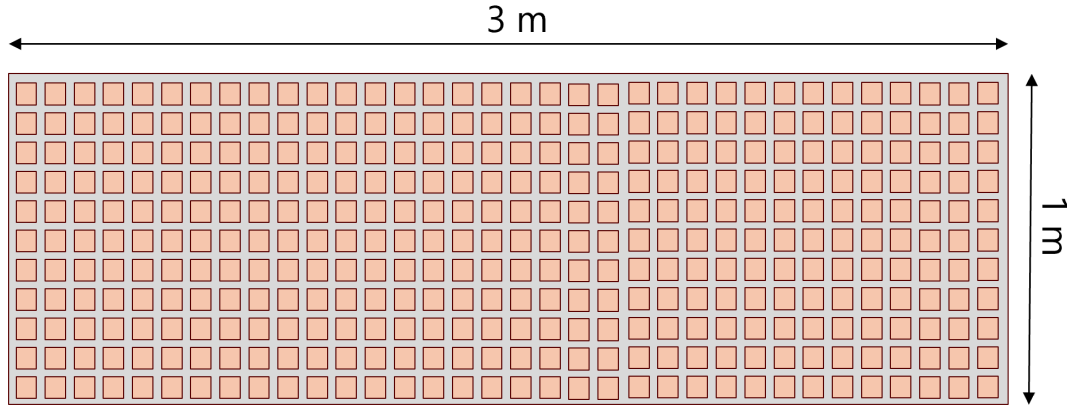
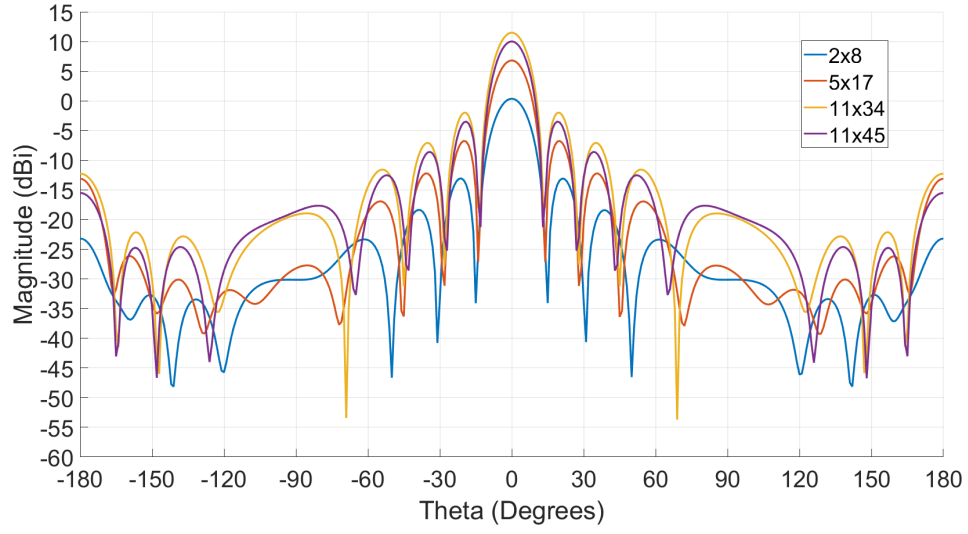
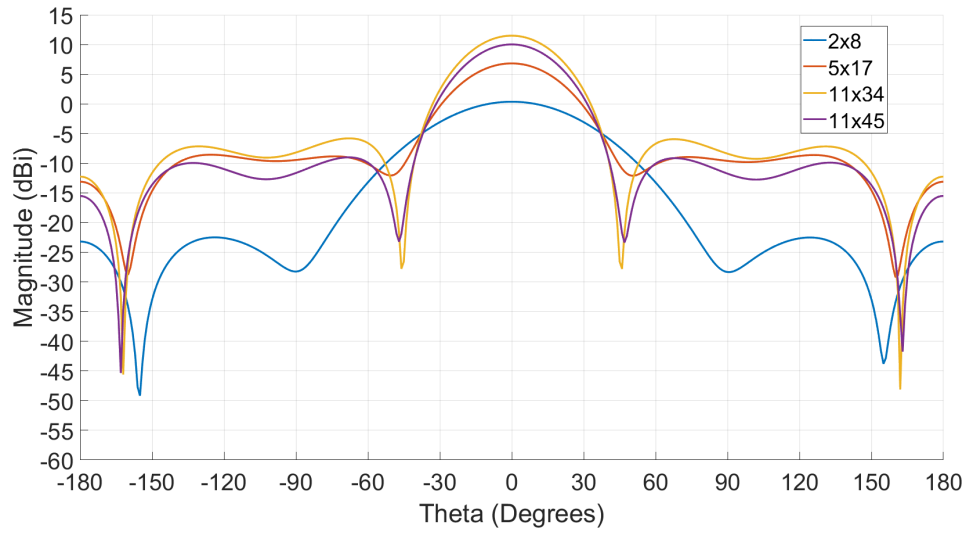


Figure 4.34: Layout of the MAGTREX array, a potential replacement to NASA's EcoSAR antenna, showing a total of 11×34 elements

$$d_{max} \leq \frac{\lambda_0}{1 + \sin(\theta_{max})} \quad (4.9)$$



(a) $\phi = 0$ degrees (H-plane)



(b) $\phi = 90$ degrees (E-plane)

Figure 4.35: Elevation realized gain cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for the stacked MAGTRES MPA array of different number of elements

Refer to Appendix A for further mathematical verification of phased array antennas using an ideal case analysis that utilizes MATLAB. After obtaining the required designs theoretically, the next step is to test them and see their behavior. To do that, an in-house fabrication process must occur first, which is detailed in the next chapter.

Chapter 5

Fabrication Process

The successful fabrication of MPAs requires careful consideration of several key steps to ensure the desired performance and precision. In this section, the fabrication process is shown in detail for the three main antenna designs developed in this thesis: the 60-mil-thick single MPA, the 120-mil-thick single MPA, and the stacked MPA. Each of these antenna designs involves distinct challenges and requires tailored approaches to meet specific design objectives.

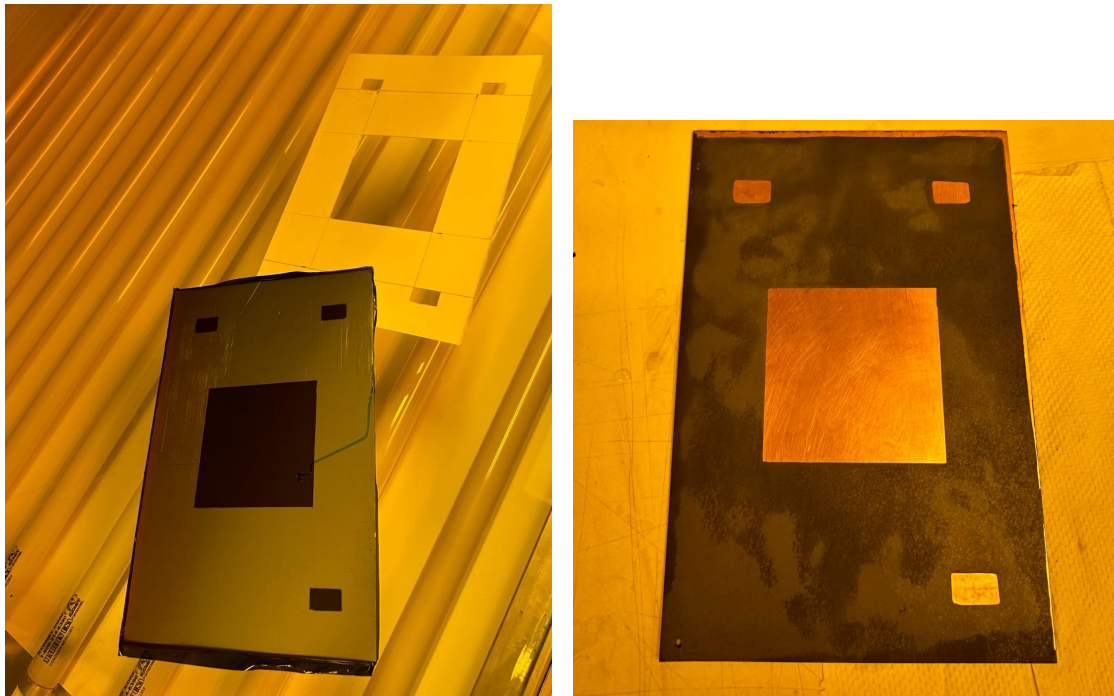
To guide the fabrication process, a series of steps are followed, as outlined in the flowchart shown in figure 5.1. This flowchart summarizes the various stages in the fabrication of the MPAs, with each design undergoing a combination of these stages based on its complexity. While not every board went through the same set of steps, the etching process remains a crucial starting point for all three designs. This fabrication process is conducted at the Advanced Radar Research Center (ARRC) at the University of Oklahoma (OU).



Figure 5.1: The flowchart of the fabrication process

5.1 Etching

The fabrication of all three antennas involves an active process of etching. For the single microstrip patch antennas, a mask was created to etch a desired area of copper off one of the sides. This was to make it easier to pattern it later using the milling machine. Once the mask was ready, a photoresist layer was added and then exposed to UV light. Afterward, the antenna was treated with sodium carbonate to remove the photoresist layer around the mask that needed to be etched. Then, the board was soaked in an etching tank to etch unwanted copper. Finally, the board was washed with the following solutions: distilled water, acetone, and isopropyl alcohol. The distilled water was used to clean the etching solution, the acetone was used to remove the photoresist, and the isopropyl alcohol is used to clean the board at the end.



(a) UV exposed lithography mask

(b) Post etching and treatment

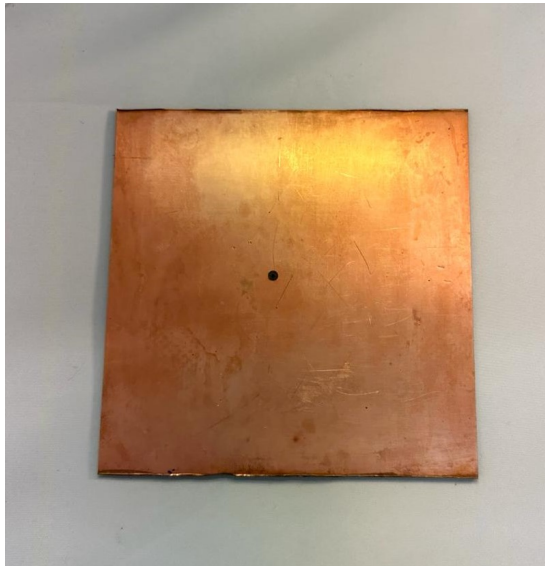
Figure 5.2: Board with patch and fiducials mask before and after etching

The same procedure was followed for the stacked antennas. The only difference is

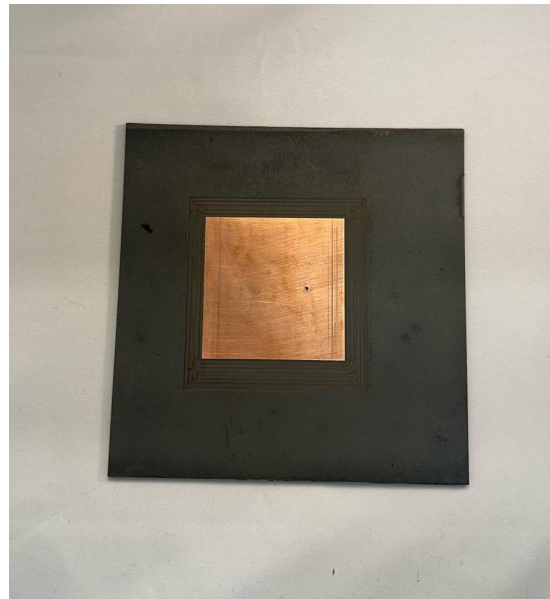
that one of the layers did not require a mask, as the entire copper area needed to be etched. The next step is to pattern the layers for the coax and the patches.

5.2 Milling

Once the etching process is complete, the milling procedure starts for all three boards. At this point, the 60-mil-thick microstrip patch antenna can be fully fabricated. Fiducials and coax via feed were drilled using the LPKF ProtoMat S104 machine, a crucial tool in this fabrication process due to its precision and efficiency. Then, the bottom and top layers were patterned using the same machine.



(a) Bottom layer



(b) Top layer

Figure 5.3: 60-mil-thick single microstrip patch antenna

The next antenna with the fewest fabrication steps is the 120-mil-thick microstrip patch antenna. Only 60-mil-thick substrates were available; thus, a 120-mil board was fabricated by laminating two boards together, which is described in section 5.3. After the lamination process, the steps to finish fabricating the design on the lower substrate

were similar to those with the 60-mil-thick antenna. First, the via was drilled, and then both the bottom and top layers were patterned using the ProtoMat S104 machine.

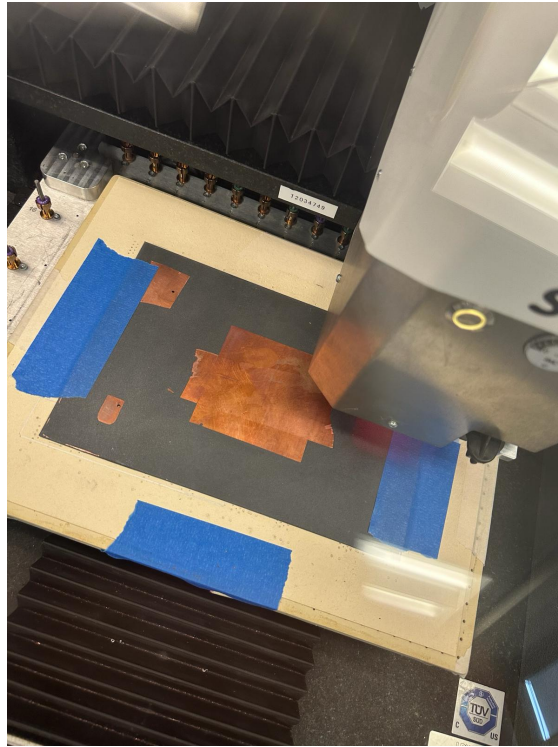


Figure 5.4: LPKF ProtoMat S104 and board setup

The fabrication of the stacked microstrip patch antennas involved a complex two-step milling process. In the first step, the lower substrates were carefully milled to drill the via, while the upper substrate was patterned to create the patch. Following this, the substrates were plated, as explained in Section 5.4. The second step involved patterning both sides of the plated lower substrate, which required meticulous attention to detail. Once this was completed, the final step was laminating the substrates together to achieve the desired design thickness.

5.3 Laminating

As stated, the 120-mil-thick microstrip patch antenna and the stacked antennas require lamination. A lamination machine was used, and a 0.5-mil-thick LF1500 prepreg from DuPont was used between the substrates that are to be laminated, as shown in figure 5.5. The first step in using the machine was to isolate the prepreg film from its protective sheet, whereas the second step was to laminate the lower and the upper substrates of each board together.



Figure 5.5: Lamination machine with a sandwiched stacked board.

To carry out the second step, alignment pins were used between the two boards to ensure that the position of the upper patch remained accurately aligned with its designed location, minimizing any shift, as shown in figure 5.6.

5.4 Plating

For the stacked MPA, the via of the lower substrate was plated to ensure the best connection when adding the SMA connector. The plating procedure also consisted of

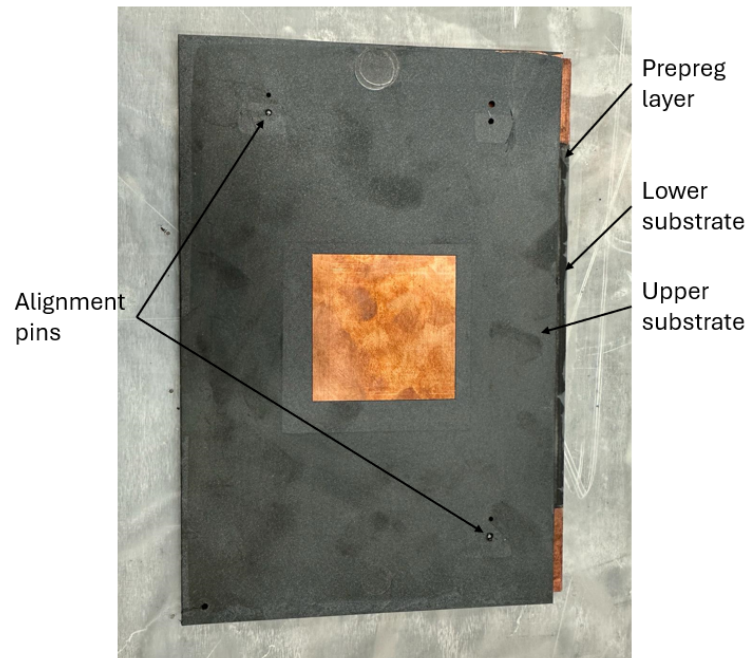


Figure 5.6: Antenna with alignment pins before lamination

two steps. The first step was electroless plating, which included performing a titration procedure followed by dipping the boards into four different tanks in a specific order for an appropriate time. The titration was performed to ensure that the plating tank reached the correct operational level.

After completing the electroless plating step, electroplating was performed to further enhance the conductivity and thickness of the metal layers. Once all plating processes were finalized, the next step involved attaching an SMA connector to the antenna. The soldering process for the SMA connector requires precision to ensure a reliable electrical connection as shown in figure 5.8. For the single MPAs, the SMA connector was inserted through the designated holes in the substrate. Solder paste was then applied, and a heat gun was used to melt the paste, ensuring a solid electrical connection between the connector and the antenna. In the case of the stacked MPA, the SMA connector was slightly trimmed to prevent it from penetrating into the second layer of the antenna. The

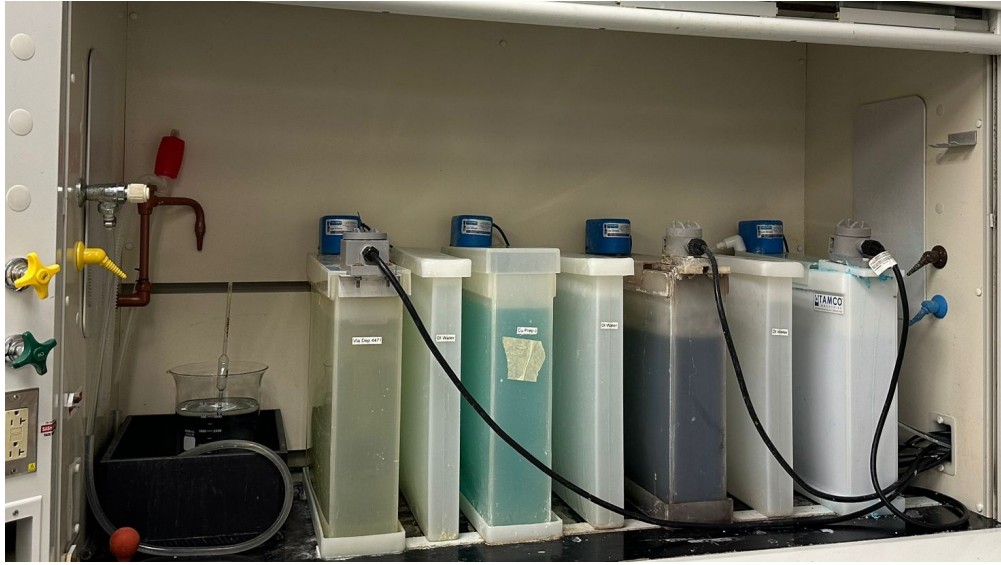
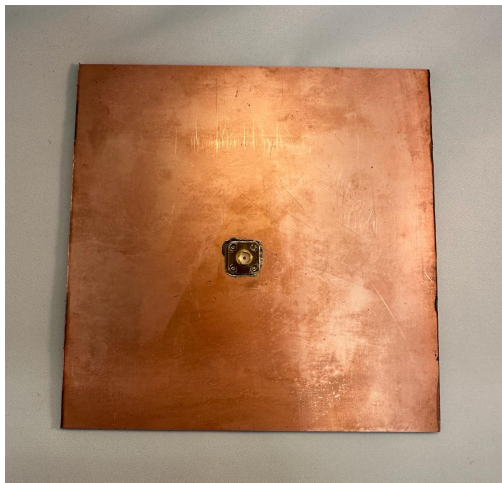
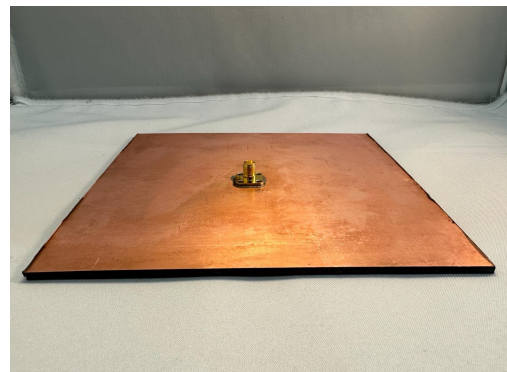


Figure 5.7: Electroless plating tanks

same solder paste procedure was applied to ensure proper connectivity, with the heat gun carefully heating the paste to form a reliable bond without damaging the layers of the antenna.



(a) Top view



(b) Isometric view

Figure 5.8: Soldered SMA connector to the antenna surface

Now that the fabrication is done, the next step is to take measurements to compare the performance of the fabricated antennas with the simulated designs.

Chapter 6

Experimental Validation

This chapter focuses on validating the antenna designs presented in this thesis by testing the fabricated prototypes and obtaining the necessary measurements for comparison. These measurements provide insights into the performance of the microstrip patch antennas and their alignment with the design objectives. Additionally, the results will help evaluate the implementation of the antennas using MAGTREX material, illustrating the extent to which the challenges outlined in the problem statement in section 1.3 have been addressed.

6.1 Fabrication Measurements

The experimental setup involves using a vector network analyzer (VNA) and reference antennas to perform different types of measurements. The VNA is utilized to obtain the S-parameter plots, while the reference antennas will be employed to measure the realized gain of the fabricated antennas inside an anechoic chamber.

6.1.1 S-parameter Measurements

A VNA provides full S-parameter response for an antenna. Measurements are conducted for the three MPAs, with the measurement setup illustrated in figure 6.1.

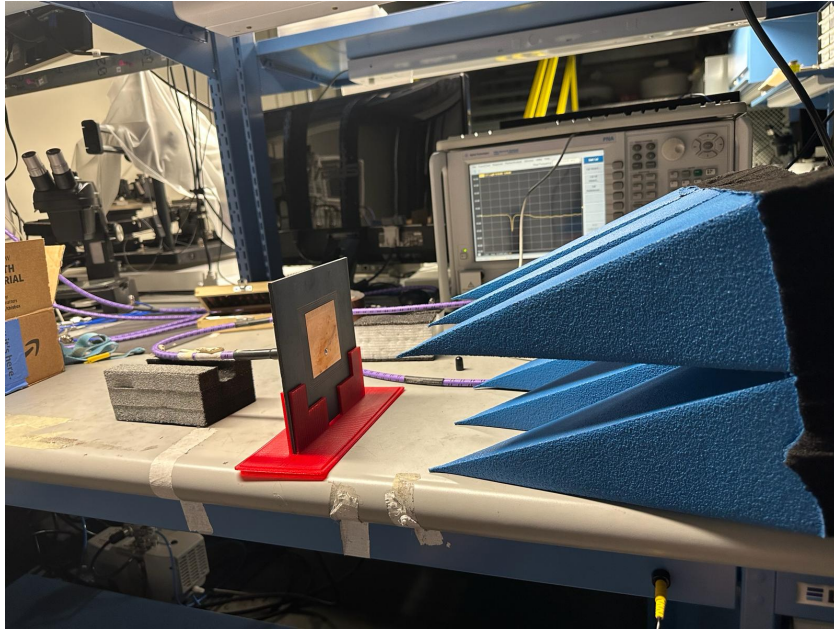


Figure 6.1: VNA network analyzer and antenna setup

After setting up and obtaining the measurements, the resonant frequency for all antennas shifted by approximately 40 MHz toward a higher frequency as can be seen in figure 6.2. For the single MPAs, this issue was resolved by increasing the patch size using copper tape, as illustrated in figure 6.3.

The patch size of the 60-mil-thick microstrip patch antenna was measured to be $61.65\text{mm} \times 61.65\text{mm}$ after fabrication. Copper tape was added to increase the patch size to $68.06\text{mm} \times 68.06\text{mm}$. Figure 6.4 shows a resonance frequency response of 435 MHz at -17.88 dB with an operating bandwidth of 14.63 MHz from 429.57 MHz to 444.42 MHz, a fractional bandwidth of 3.39%.

As for the patch size of the 120-mil-thick microstrip patch antenna, it was measured to be $61.98\text{mm} \times 61.98\text{mm}$ after fabrication. Copper tape was added to increase the patch size to $68.87\text{mm} \times 68.87\text{mm}$. Figure 6.5 shows a frequency response resonant at 435 MHz at -21.35 dB with an operating bandwidth of 22.92 MHz from 421.66 MHz to 444.58 MHz, a fractional bandwidth of 5.29%.

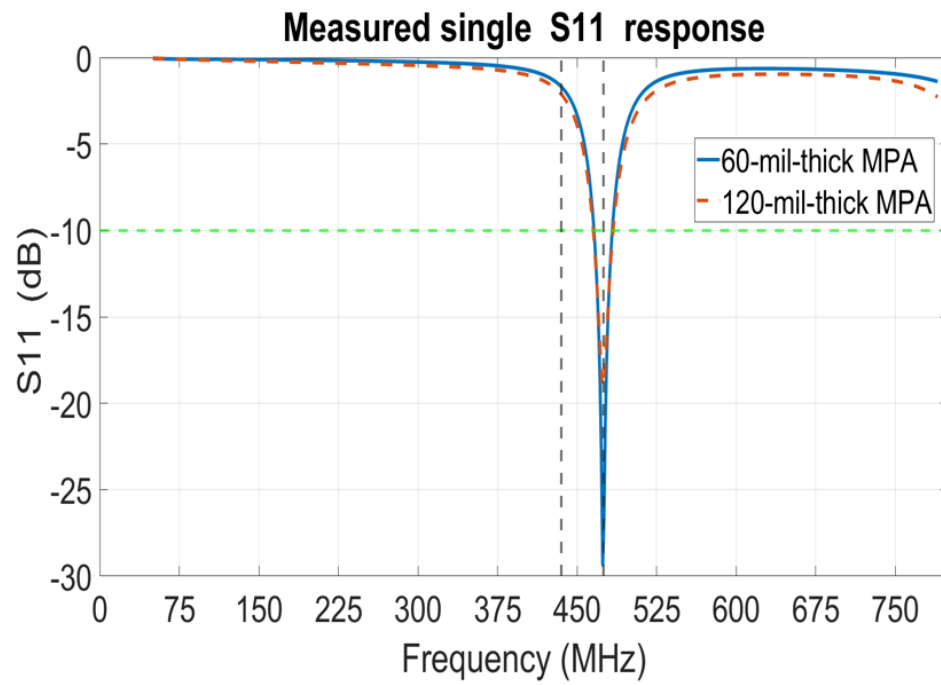


Figure 6.2: Measured return loss response comparison for the single MPAs

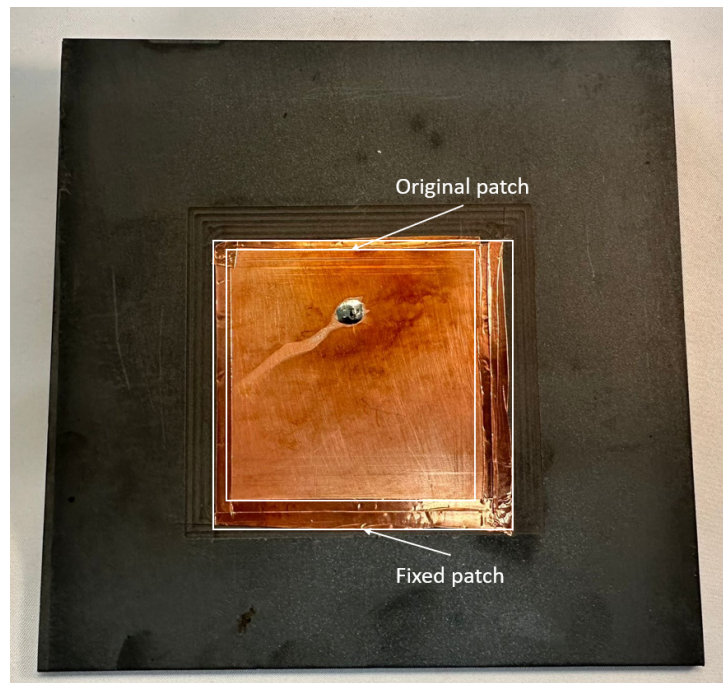


Figure 6.3: Microstrip patch antenna response tuned by making the patch bigger

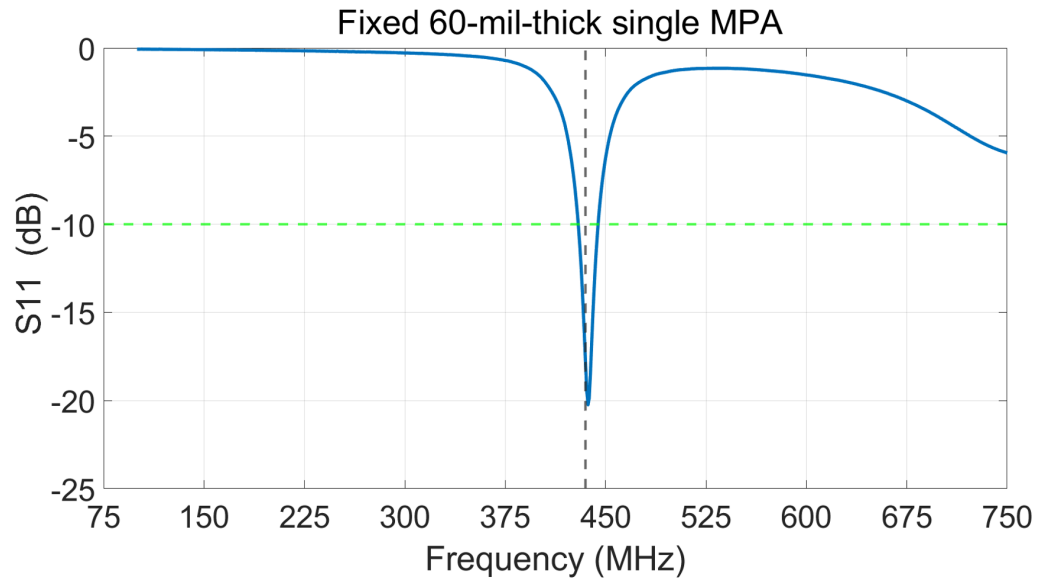


Figure 6.4: Measured return loss response of the tuned 60-mil-thick single MPA

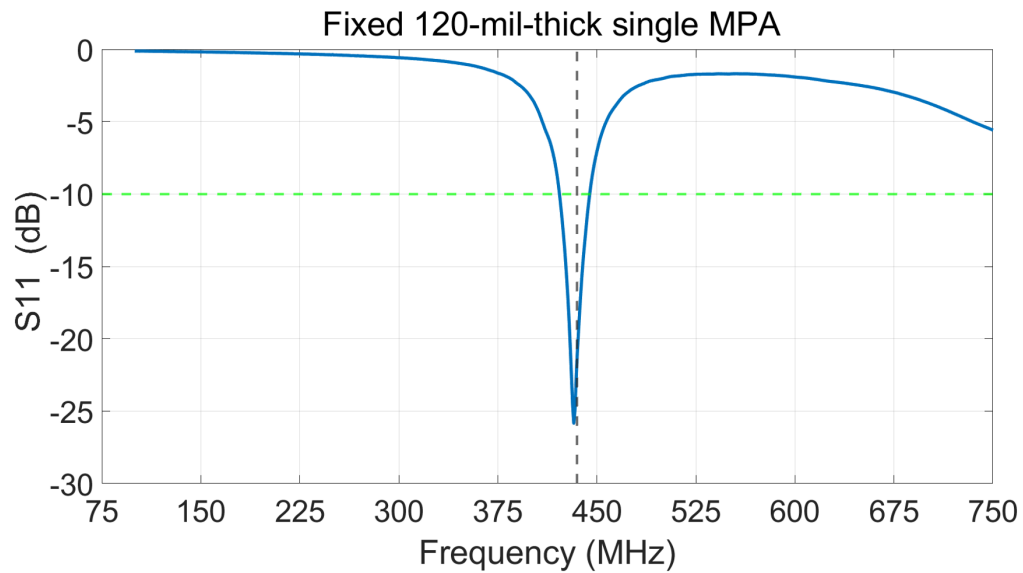


Figure 6.5: Measured return loss response of the tuned 120-mil-thick single MPA

The stacked MPA was also measured and exhibited a 40 MHz shift in its resonant frequency toward a higher frequency. However, tuning it with copper tape was not feasible due to the inaccessible buried patch located between the substrates.

Due to this shift, the stacked patch antennas cannot be tuned and should be kept

as measured. The obtained responses are shown in figure 6.6. The figure shows the resonance frequency at 473.625 MHz at -20.45 dB with an operating bandwidth of 40.14 MHz from 462.72 to 502.860 MHz, a fractional bandwidth of 8.3%.

For the stacked MPA, any shift in alignment can affect the coupling between the patches. For a better comparison as stated in section 4.1, the 120-mil-thick single MPA is the one that is compared with the stacked MPA for bandwidth comparison as shown in figure 6.7.

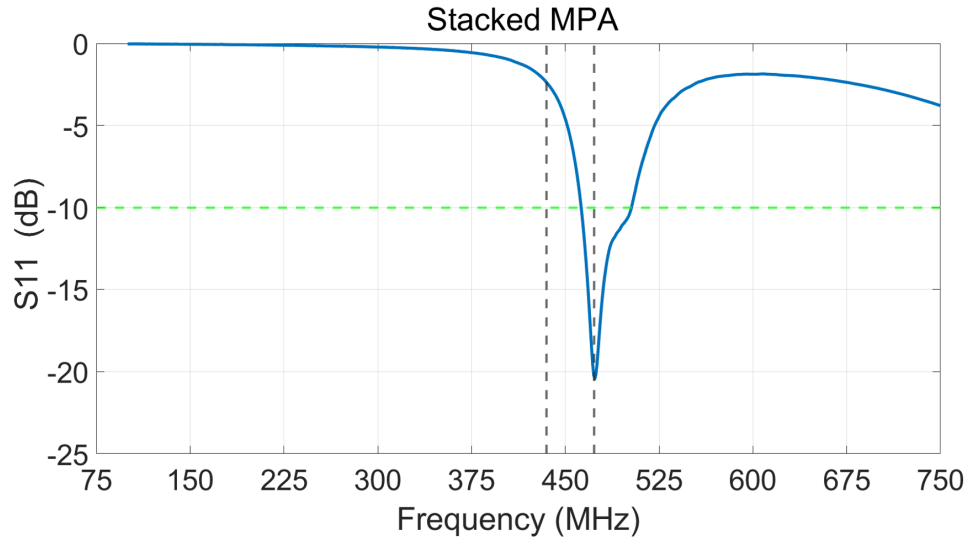


Figure 6.6: Measured return loss response of the stacked MPA

6.1.2 Realized Gain Measurements

Two TBMA1 30MHz - 1000 MHz Biconical antennas were used to obtain the realized gain measurements of the fabricated antennas. The Biconical antennas were set up as shown in figure 6.8 with reflectors in the middle to minimize ground reflections. S_{21} plot was obtained from the two reference Biconical antennas using a FieldFox. Following that, one of the Biconical antennas was removed and replaced by the designed antennas, AUTs, one after the other to obtain the S_{21} values.

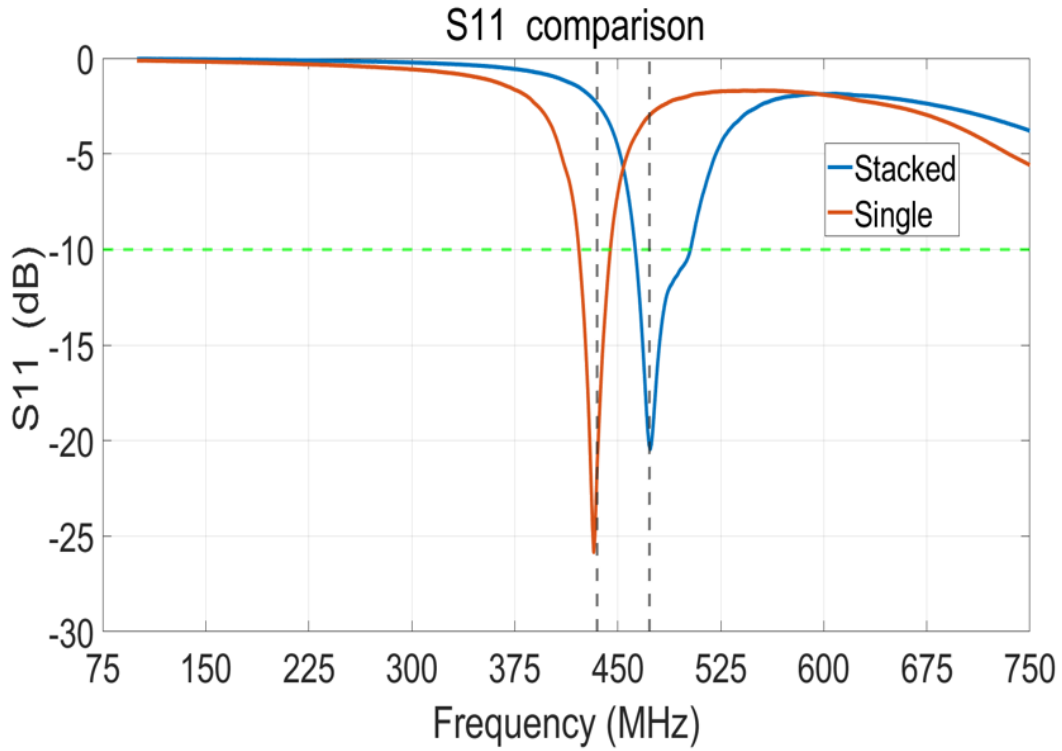


Figure 6.7: Measured return loss responses comparison of the stacked MPA vs the 120-mil-thick single MPA

Following that, the Biconical datasheet provides AF values for certain frequencies. A Matlab script was used to interpolate the AF values for the non-mentioned frequencies and then the reference isotropic gain was obtained. Then, an offset S_{21} value for each frequency was obtained by subtracting the obtained S_{21} values from the calculated reference isotropic gain. Following that, the obtained S_{21} offset was added to the measured S_{21} of the antennas, AUTs. All the needed data was obtained at this point before it was plotted and compared as shown in figures 6.9, 6.10, and 6.11.

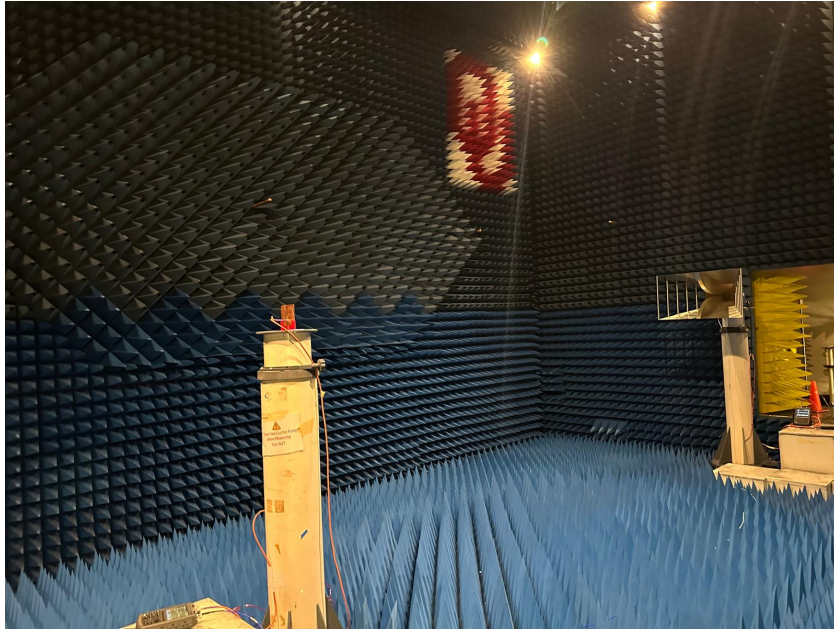


Figure 6.8: Realized gain measurement setup showing a horn and an MPA inside a far-field anechoic chamber

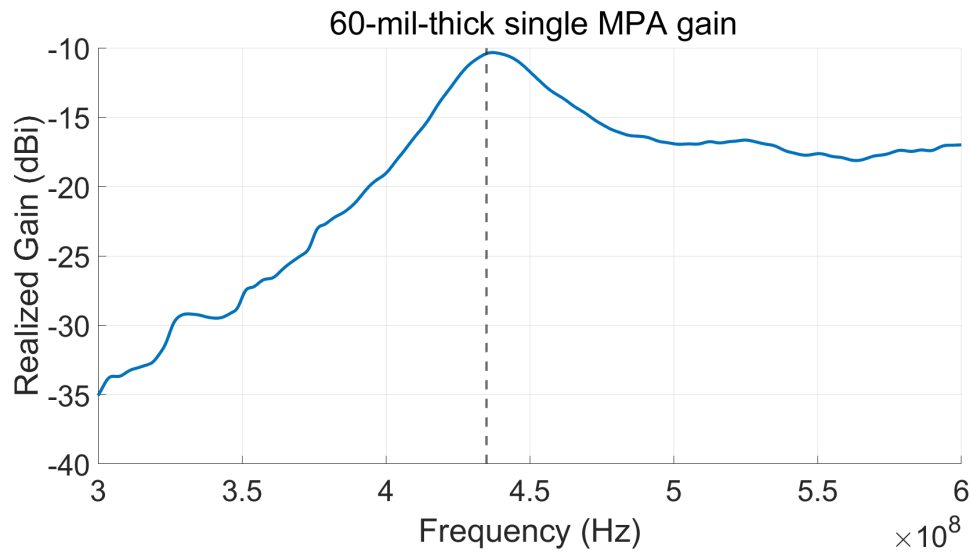


Figure 6.9: Measured realized gain of the 60-mil-thick single MPA

From the figures, it can be seen that the antennas are behaving as expected. To show how well measurement matches simulations, a comparison is thoroughly performed in the next section, 6.2.

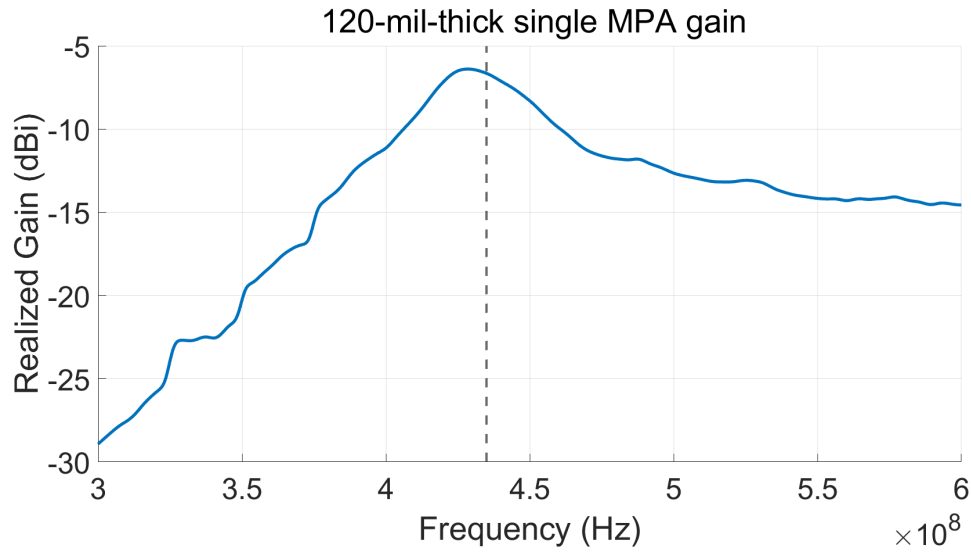


Figure 6.10: Measured realized gain of the 120-mil-thick single MPA

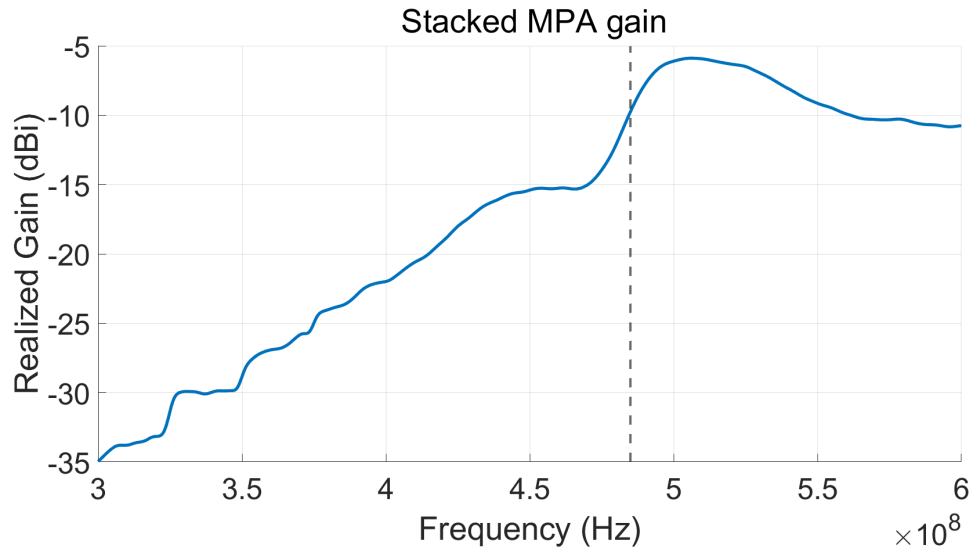


Figure 6.11: Measured realized gain of the stacked MPA

6.2 Comparison of Design Simulations and Fabrication Results

This section shows plots of measured vs. simulated return loss and gain for all three antennas to verify the simulations of the fabricated antennas with the obtained measurements. One reason behind the shift of the resonance shown in subsection 6.1.1

is the discrepancy in substrate properties between what is sent by ROGERS and what is reported on the datasheet. A 40 MHz shift represents a $\sim 10\%$ change which can be obtained through simulation by lowering the multiplication of $\sqrt{\epsilon_r \mu_r}$ by 10%. This can be done by using a $\mu_r = \epsilon_r = 5.4$. Also, the substrate is designed for optimal performance up to 500 MHz, while some of these designed antennas have performance that extends beyond that as per the bandwidth requirement.

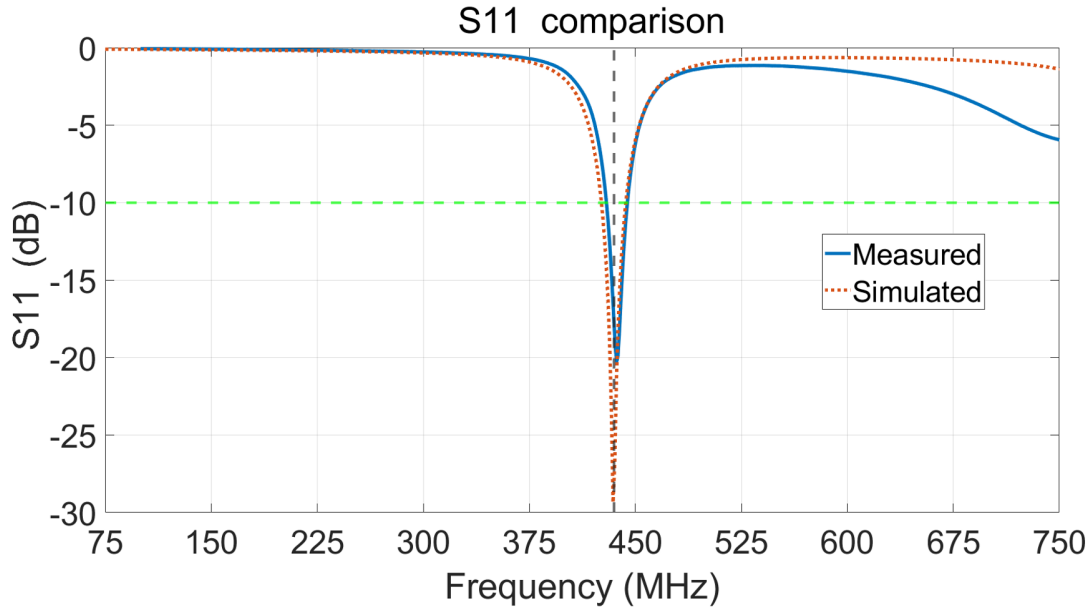


Figure 6.12: Measured vs simulated return loss responses of the 60-mil-thick single MPA

Figures 6.12 and 6.13 show that the behavior of the MAGTRES is similar between measurement and simulation up to 500 MHz for the single patches, which is expected from the material as it is designed with an operating frequency up to 500 MHz. Once the frequency exceeds 500 MHz, the simulated responses do not show any realizable change whereas the measured ones start showing a return loss difference greater than 3 dB.

However, figure 6.14 shows that the resonance frequency of the fabricated antenna shifted to the right, whereas the expectations of the bandwidth remained approximately

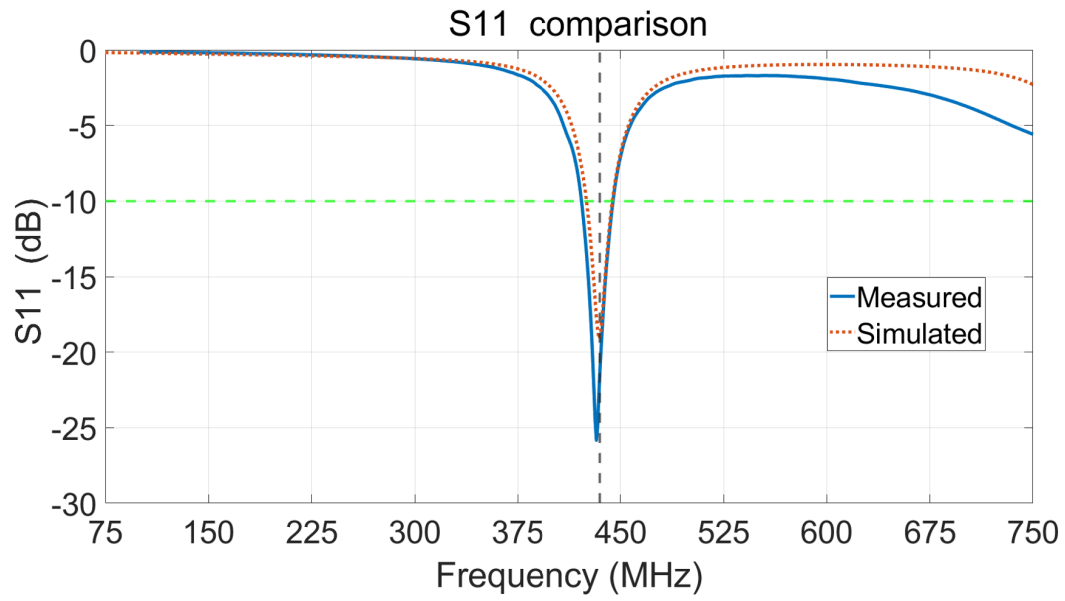


Figure 6.13: Measured vs simulated return loss responses of the 120-mil-thick single MPA

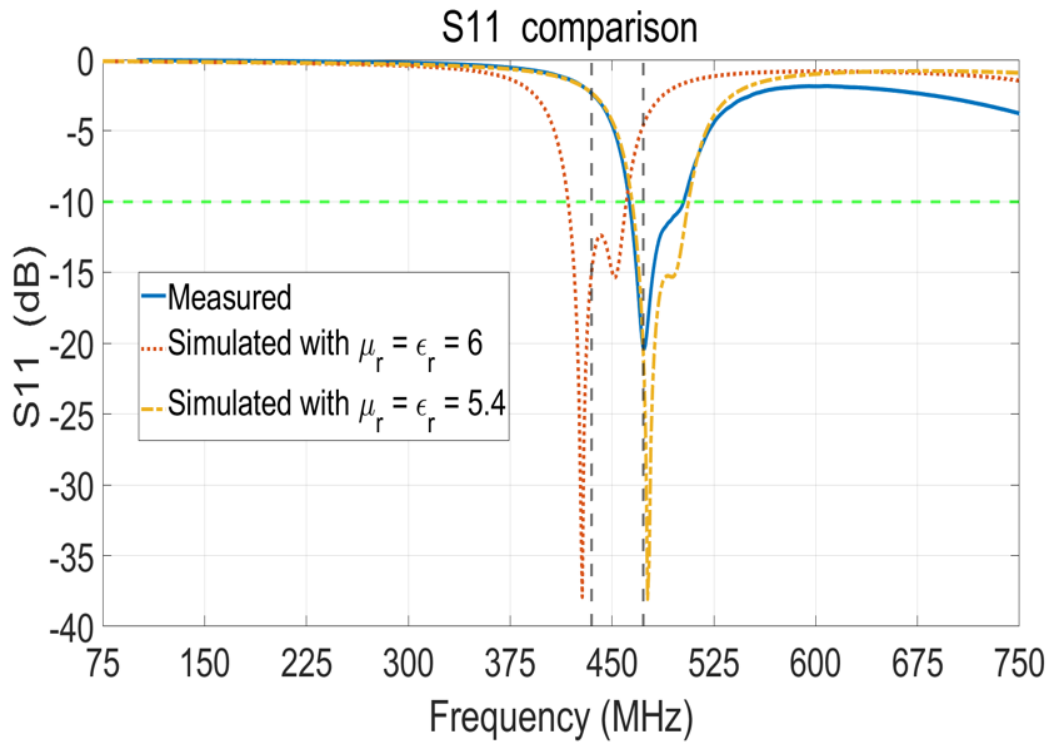


Figure 6.14: Measured vs simulated return loss responses of the stacked MPA

the same. It was found to be 41.6 MHz for the simulated design and 40.1 MHz for the fabricated design. Two simulations were shown for a better comparison because the shift in resonance in the stacked comparison was not possible to tune with copper tape. In conclusion, having a stacked MPA instead of a single MPA increased the bandwidth for a magneto-dielectric material by 75.13%, from 22.92 MHz to 40.14 MHz.

Now that the $|S_{11}|$ comparison is done, the gain of the antennas needs to be compared as well which is shown in figures 6.15, 6.16, and 6.17.

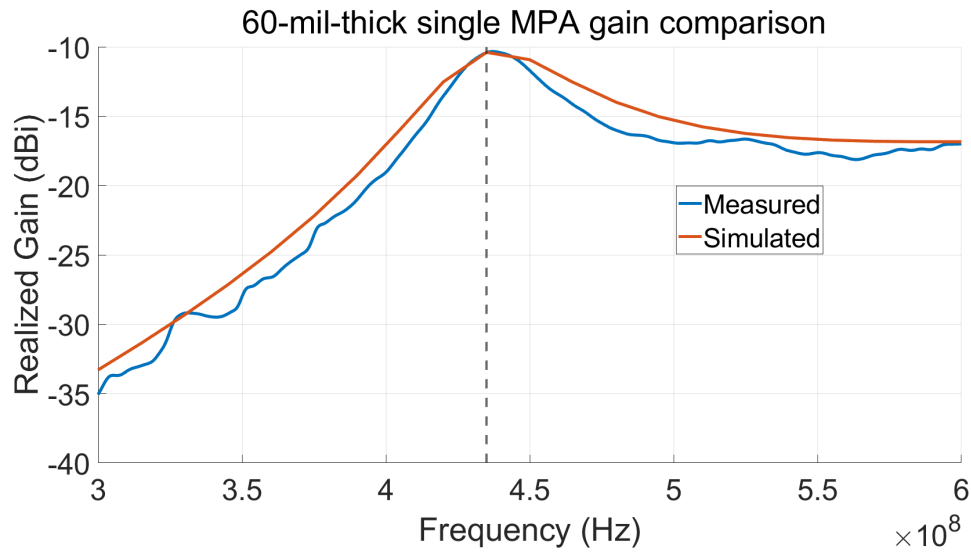


Figure 6.15: Measured vs simulated realized gain responses of the 60-mil-thick single MPA

Similar to the case with the return loss response comparisons, figures 6.15 and 6.16 show that the behavior of the MAGTREX is similar between measurement and simulation up to 500 MHz for the single patches. Besides, the realized gain shows an exact match at the resonant frequency of 435 MHz. The simulated plots are done at certain points in the range from 300 MHz to 600 MHz which explains some of the discrepancies around 435 MHz.

On the other hand, figure 6.17 shows that the frequency shift matches in radiation

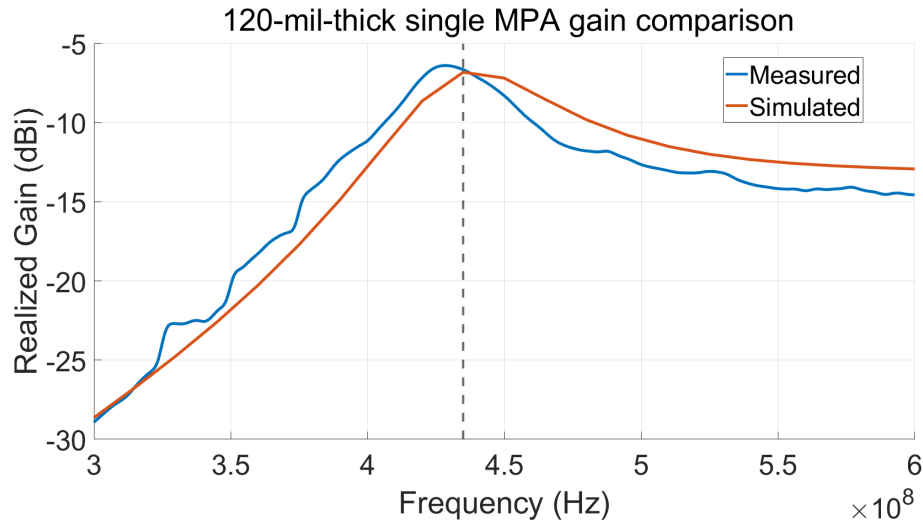


Figure 6.16: Measured vs simulated realized gain responses of the 120-mil-thick single MPA

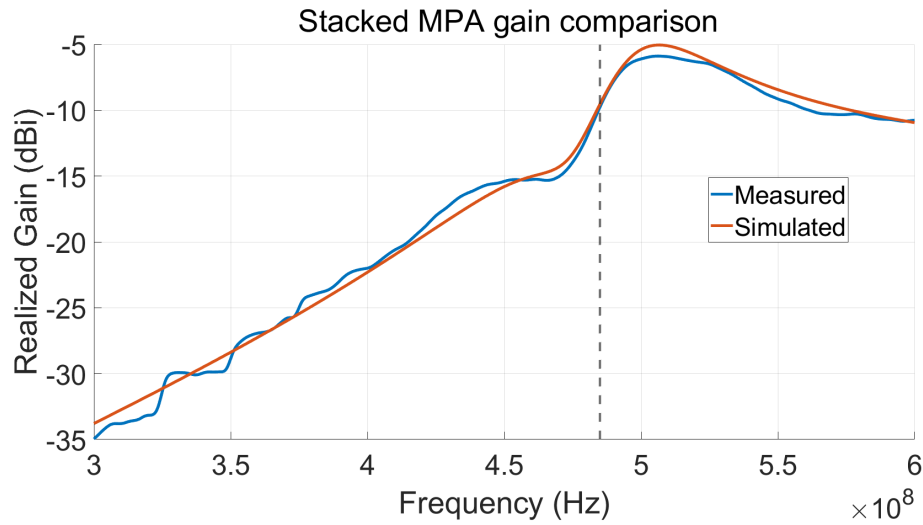


Figure 6.17: Measured vs simulated return realized gain responses of the stacked MPA

between simulation vs measurement where the overall performance shows peak radiation at 500 MHz instead of 485 MHz. Given that the stacked MPA spans a 41.4 MHz bandwidth from 465.10 to 506.5 MHz, the radiation is happening towards the higher bandwidth end with a peak realized gain of -5.89 dBi at 500 MHz. To see how that compares to the return loss response directly, figure 6.18 can be referred to.

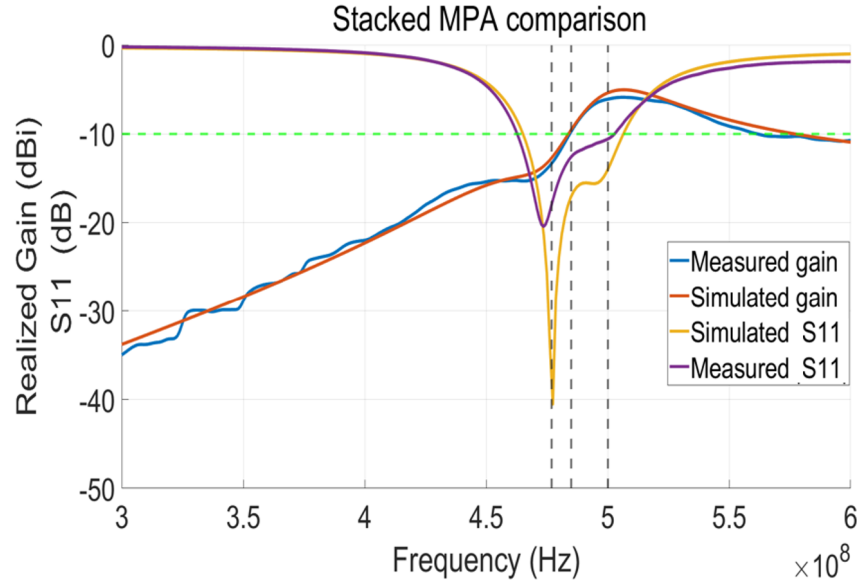


Figure 6.18: Measured vs simulated realized gain responses of the stacked MPA alongside their return loss responses

This shift in peak realized gain in a stacked MPA configuration is expected as seen in previous work such as [38] and [39].

In addition to verifying the realized gain, the radiation patterns can be compared between measurement and simulation for all three antennas. Starting with the 60-mil-thick MPA, figure 6.19 can be referred to. From the figure, it is evident that the co-polarization values closely match between the simulation and the measured results. However, the cross-polarization is significantly higher in the measured data compared to the simulation. The cross-polarization isolation is approximately 30 dB in the measured results, whereas in the simulation, it was expected to be much higher than 30 dB. This discrepancy is consistent across all three antenna designs. The difference arises because cross-polarization isolation is calculated under ideal conditions in HFSS, while in the anechoic chamber, external factors, such as interference from surrounding wires and reflections, lead to lower isolation in the actual measurements than predicted by the simulation.

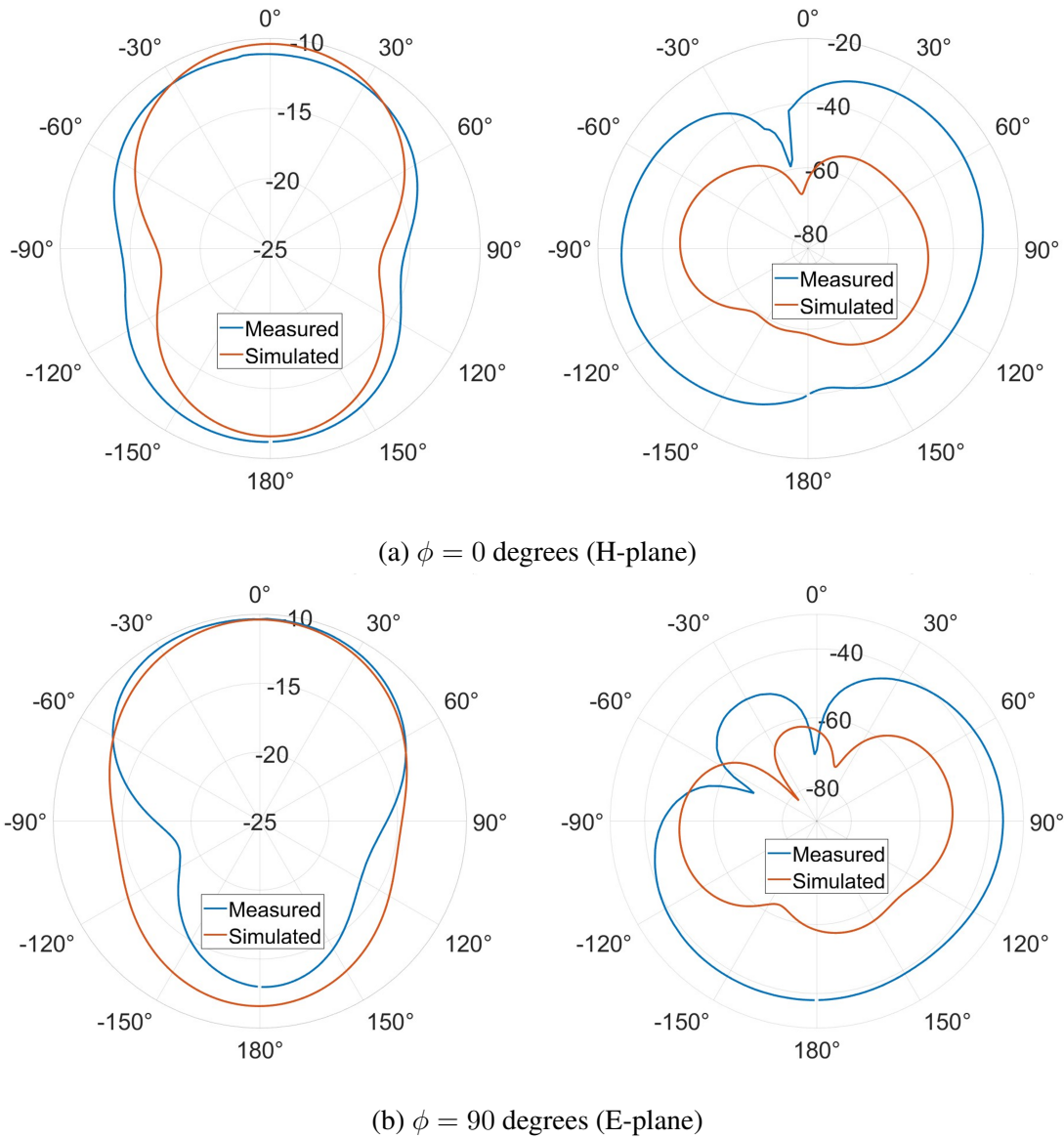


Figure 6.19: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a measured vs simulated 60-mil-thick MPA

A similar comparison can be made to the 120-mil-thick MPA with a similar trend in the result. Figure 6.20 shows cross-polarization isolation greater than 30 dB, which matches the EcoSAR requirement. However, this is not the stacked configuration yet; thus, a measurement conclusion about the cross-polarization isolation cannot be made yet. To make this conclusion, figure 6.21 can be referenced.

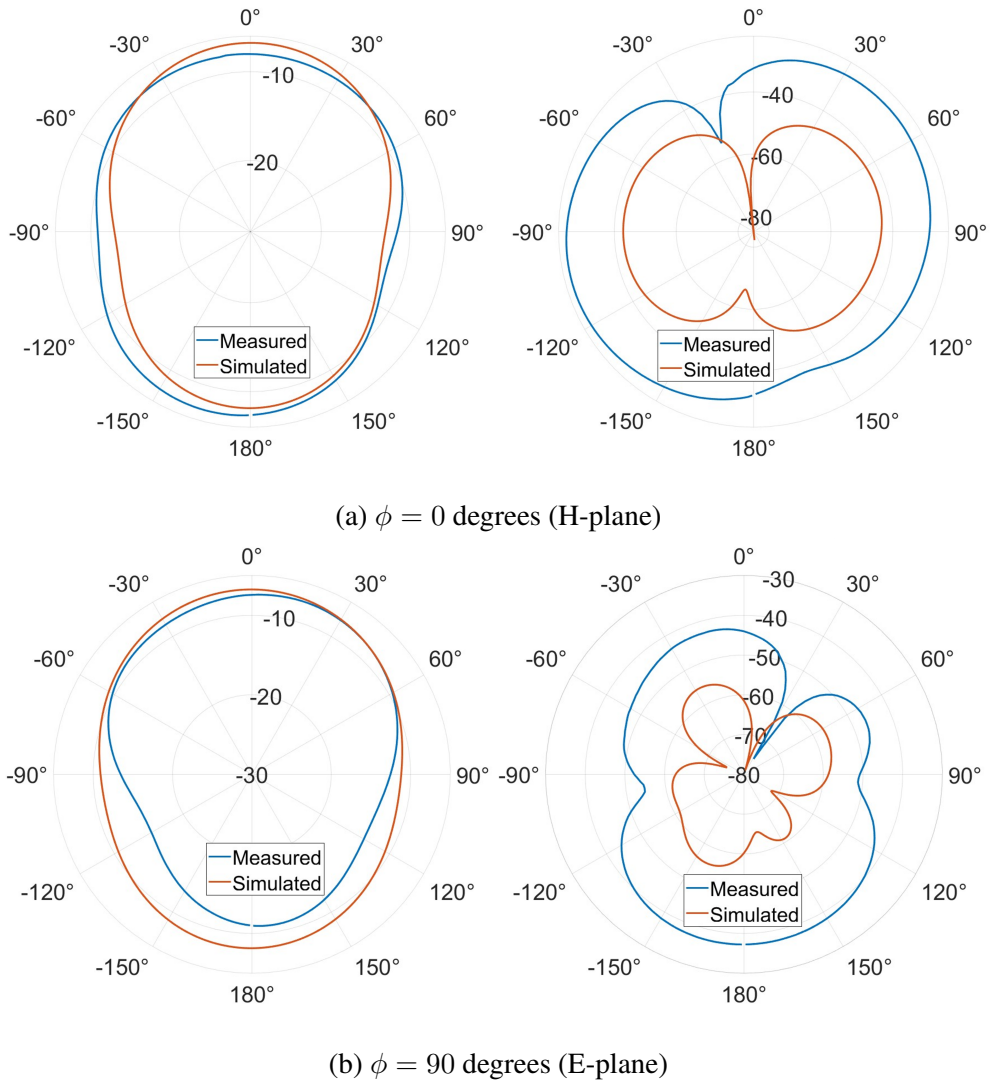


Figure 6.20: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a measured vs simulated 120-mil-thick MPA

Figure 6.21 shows that the stacked microstrip patch antenna achieved similar results in co-polarization while managing to meet the EcoSAR requirement by NASA of having a cross-polarization isolation greater than 30 dB.

After verifying the simulated designs with the fabricated designs, a comprehensive conclusion needs to be drawn, along with a discussion for future work in the next chapter.

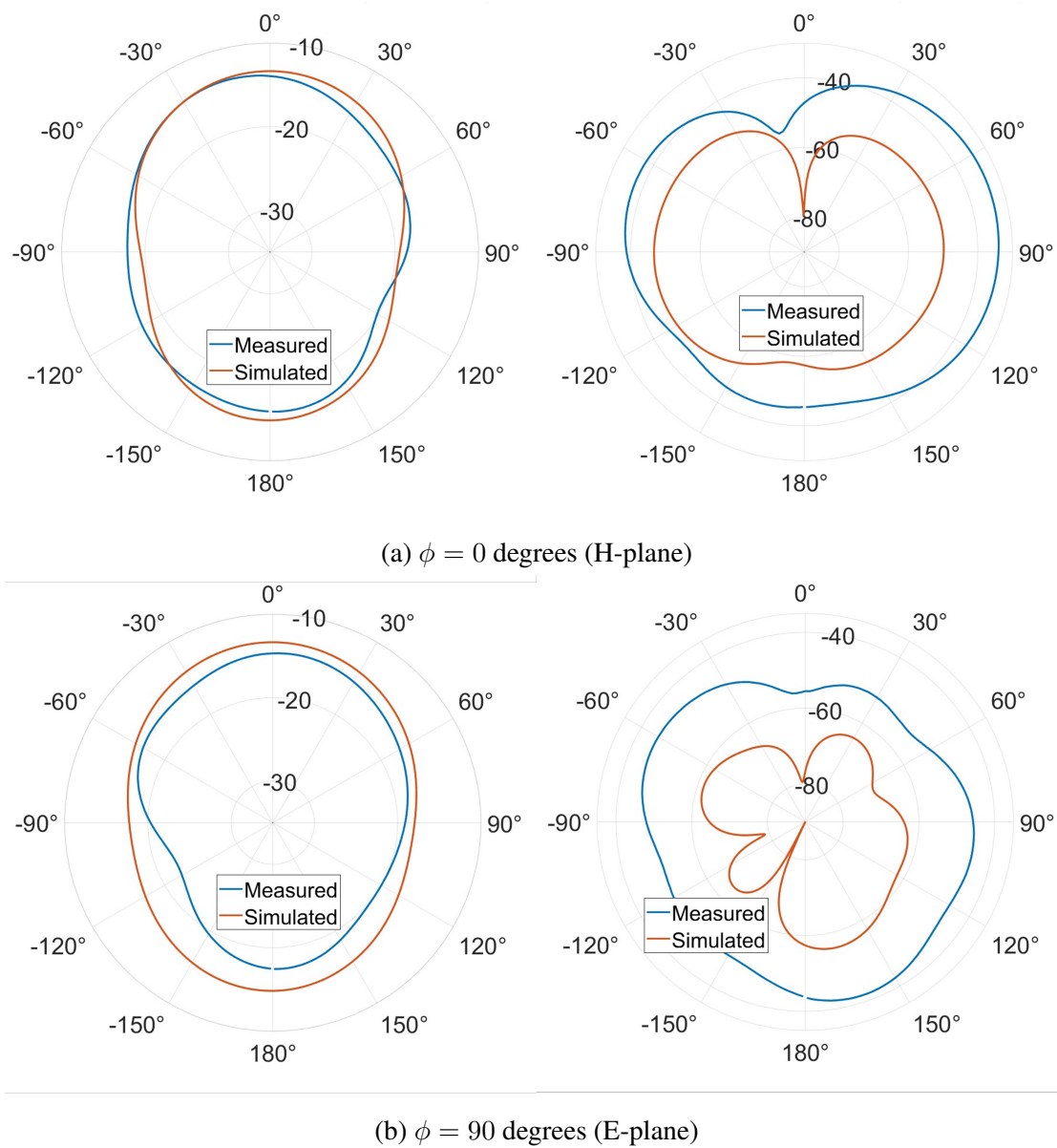


Figure 6.21: Elevation co-polarization (left) and cross-polarization (right) cut planes at $\phi = 0^\circ$ (top) and $\phi = 90^\circ$ (bottom) for a measured vs simulated stacked MPA

Chapter 7

Conclusion and Future Work

7.1 Conclusion

This thesis explored the design, implementation, and fabrication of microstrip patch antennas, with a focus on using magneto-dielectric materials for miniaturization in airborne radar systems. The research evaluated both single and stacked microstrip patch antennas, examining the design process using the cavity model and emphasizing the importance of precise length adjustments for accurate resonance frequency tuning. The study revealed that material losses, especially in single microstrip patch antennas with a lossy magnetic substrate, limit performance. In contrast, stacked configurations were found to improve bandwidth and gain, although the increased complexity and loss mechanisms generally resulted in a lower gain for stacked antennas using a lossy substrate such as MAGTREX.

The use of MAGTREX, was shown to offer significant advantages in miniaturization, due to its non-unitary permeability compared to conventional substrates. For MAGTREX, $\epsilon_r \simeq \mu_r$, a property that enables better impedance matching to free space, resulting in improved performance and broader bandwidth. Additionally, the research demonstrated that MAGTREX substrates could effectively reduce antenna size by 83.33% while maintaining functional properties.

In terms of antenna performance, MAGTREX allowed for the development of stacked microstrip patch antennas with a resonant frequency of 435 MHz and a maximum bandwidth of 40.14 MHz, a fractional bandwidth of 8.3%, with the chosen thickness. Regardless of the negative gain, further optimization of the antenna's ground plane dimensions enhanced the directivity and realized gain, indicating the potential for increased system performance. Besides, when an array of stacked antennas was designed, the radar system's efficiency improved, with a realized gain of 11.5 dBi.

The findings of this thesis also provide valuable insights into the limitations of using MAGTREX in antenna design, particularly regarding bandwidth and radiation efficiency. Despite the advantages in miniaturization and bandwidth, trade-offs in radiation efficiency were observed, especially when comparing MAGTREX to alternative designs meant for miniaturization like PIFA. Additionally, the effects of substrate thickness and losses on antenna performance were examined, highlighting the importance of careful design choices to optimize performance for specific applications.

Finally, this work demonstrates the potential for using MAGTREX to reduce the size and weight of radar systems, offering a 66.85% weight reduction compared to NASA's EcoSAr. However, this benefit comes with the trade-off of decreased radiation efficiency, 11.5 dBi compared to 19 dBi by EcoSAr. Also, the volume of NASA's EcoSAr antenna array is $1 \times 3 \times 0.19 \text{ m}^3$ whereas an array made of the stacked MAGTREX in the same volume leaves a volume of $1 \times 3 \times 0.178 \text{ m}^3$ that can be used to store electronics. Furthermore, while it was previously difficult to investigate dense arrays with element spacing d less than $0.5\lambda_0$, using a magneto-dielectric substrate like MAGTREX allows the patch size to become smaller than in free space, enabling more elements to fit within the same aperture and thereby increasing gain.

In conclusion, the results of this thesis underscore the potential of MAGTREX to replace existing antenna structures, contributing to cost reductions for future space explo-

rations and enabling current aircraft to become smaller and lighter. The study demonstrated how MAGTREX can effectively miniaturize antenna systems while providing valuable insights into the trade-offs between size, bandwidth, and radiation efficiency. These findings highlight MAGTREX as a promising material for advancing antenna technology, particularly for aerospace applications, where weight and size reductions are crucial.

7.2 Future Work

Regarding the work shown in this thesis, much more can be done to continue the work and improve the bandwidth. Firstly, investigations with cavity-loaded elements can be used to achieve Z-axis miniaturization which can also help examine how the bandwidth can be broadened. This cavity can be either left empty or filled with other substrates of different permittivity and permeability to examine different effects on the antenna. Besides, only one polarization was examined here; further research can include two different feeding positions that can be excited at different times to achieve different polarizations, vertical or horizontal. Moreover, the MAGTREX can be further researched by the manufacturer to see how dielectric and magnetic losses can be reduced to fix its efficiency with radiating energy. Also, redesigning a substrate to operate at higher frequencies than 500 MHz can help mitigate these problems. Thus, an investigation with the MAGTREX that will be released by Rogers later can be helpful as it promises operation up to 1.6 GHz which lowers the losses in the P-and-L-band regions.

Although there is still plenty to be done, the work presented by this thesis can give antenna engineers a better grasp on the effect of magneto-dielectric materials and how they can help miniaturize antennas. It also shows that with these results, the MAGTREX material should only be used in a phased array configuration to get a good gain as a

trade-off for its lossy behavior.

Finally, future research should explore alternative materials with lower loss tangents, advanced modeling to predict coupling coefficient behavior better, and refined fabrication methods to ensure performance consistency. This thesis can serve as a reference for future innovations in compact, high-performance antenna systems, bridging the gap between theoretical design and practical implementation.

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Appendix A

Phased Array Mathematical Verification

The results of section 4.3.3 can be verified mathematically by looking at the antenna pattern as well as the array factor to see how they behave together. To simplify this verification, a linear 1-D array is designed. For this array, the element patterns are assumed to be uniform and the same, with the pattern modeled as a cosine function. Given that this thesis works with MPAs, the general shape of an ideal MPA follows a cosine function.

As for the array factor, the following equation is used:

$$(AF)_n = \sum_{n=1}^N e^{j(n-1)\psi} = \frac{1}{N} \left[\frac{\sin(\frac{N}{2}\psi)}{\sin(\frac{1}{2}\psi)} \right] \quad (\text{A.1})$$

where $\psi = kdcos(\theta) + \beta$, d is the spacing between adjacent elements, $k = \frac{2\pi}{\lambda}$, and $\beta = kdcos(\frac{\pi}{2} - \theta_s)$.

Figure A.1 shows that decreasing the number of elements while keeping the same spacing between elements reduces the gain whereas increasing the number of elements while decreasing the spacing between elements increases the gain, which supports the results shown in section 4.3.3. This is also summarized in table A.1.

To visualize the data of table A.1 in a plot form, figures A.4 and A.3 can be referred to where figure A.4 shows the changing number of elements as a function of spacing

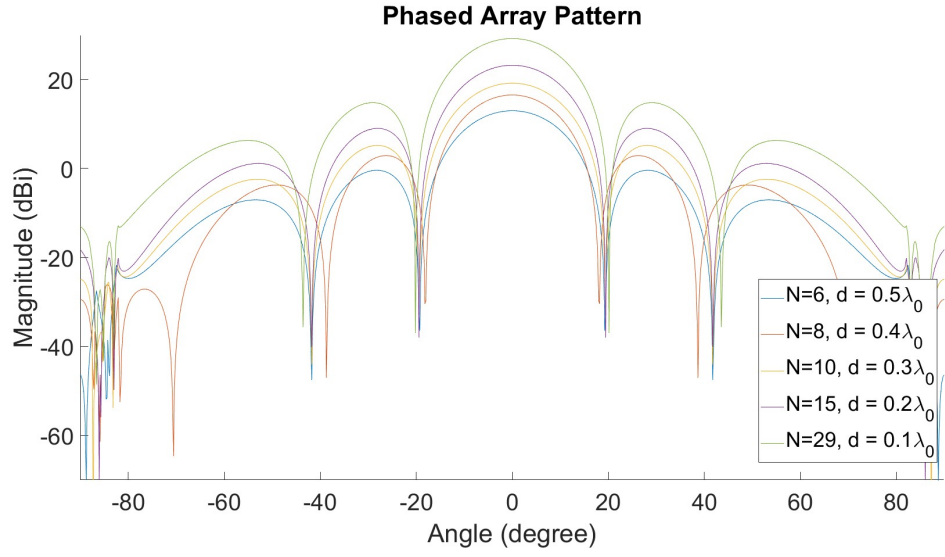


Figure A.1: Phased array patterns based on MATLAB simulations

Table A.1: Effects of element spacing on realized gain and side lobe level using ideal conditions

Array Size	Spacing	Realized Gain (dBi)	Sidelobe Level (dB)
1×6	0.5λ	12.9	13.3
1×8	0.4λ	16.5	13.6
1×10	0.3λ	19.1	13.9
1×15	0.2λ	23.2	14.2
1×29	0.1λ	29.2	14.4

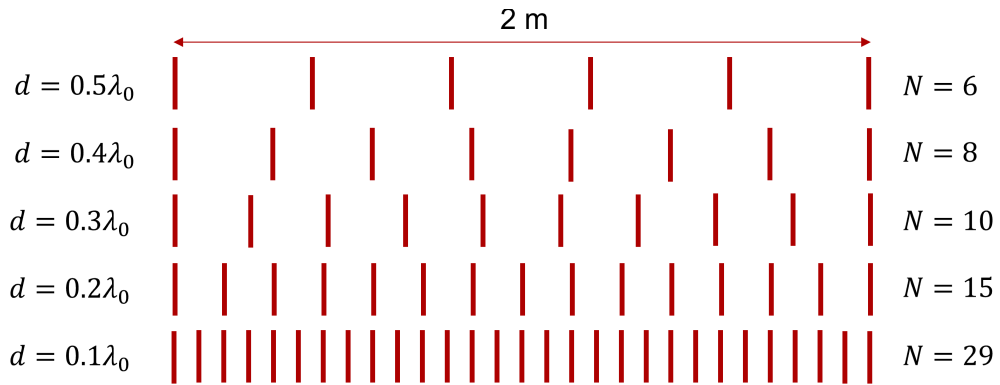


Figure A.2: Visualization of the effect of Element Spacing on Number of Elements in a Fixed-Length Phased Array Antenna

between the elements whereas figure A.3 shows the effect of number of elements on gain, all within the same length space as visualized in figure A.2. This demonstrates that, while it was previously difficult to investigate dense arrays with element spacing, d , less than $0.5\lambda_0$, using a magneto-dielectric substrate like MAGTRES can reduce the patch size in terms of wavelength, smaller than in free space. This enables fitting more elements into the same aperture, thereby increasing the gain.

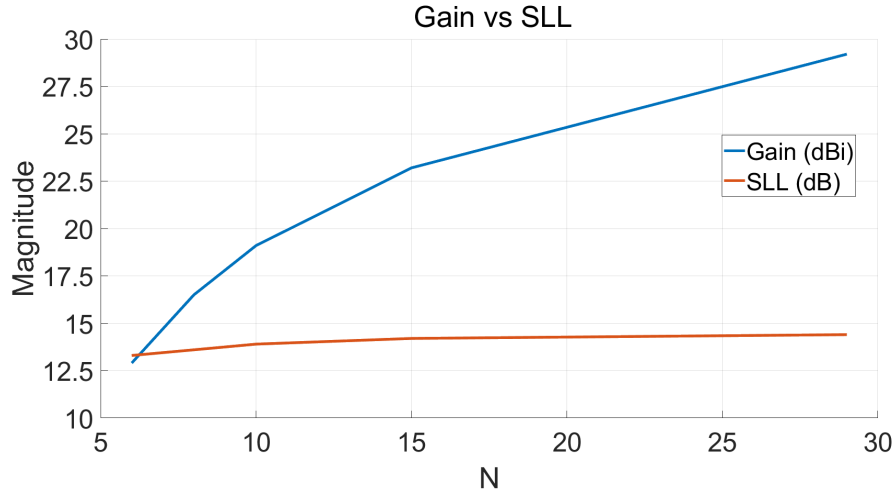


Figure A.3: Gain vs side lobe level as a function of number of elements

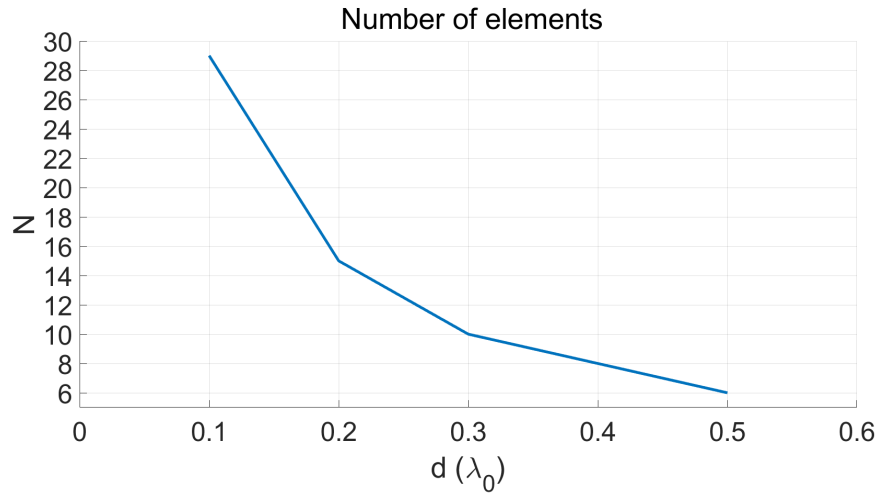


Figure A.4: Gain vs side lobe level as a function of number of elements