X BAND PHASED ARRAY DESIGN FOR RADAR APLICATION

Master Thesis

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2015
X BAND PHASED ARRAY DESIGN FOR RADAR APPLICATION

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March 31, 2015
Dedication

First, I want to thank my mother, who despite not having scientific education has always encouraged me to continue in this line of study, taught me to trust my abilities, his love and dedication have been the best gift I received it would be much less than what I am today without her. To my family and friends who have seen in me, abilities and qualities of which I am not yet aware, my gratitude to them is great, even though I don’t say it to them frequently.

Iván Darío Duarte Brito.

I dedicate this work and give special thanks to my mother, father and brother, pillars of my life, for their unconditional support and words of encouragement to improve myself, the source of love and strength all these years. To my girlfriend, who helped me keep perspective on what is important in life, for her patience and love.

Jhon Cárcenas Triana.
Acknowledgments

The authors wish to express gratitude to CODALTEC (Corporación de Alta Tecnología para la Defensa) for the assignment of the financial resources in order to complete this Master degree. Likewise, to UPM (Universidad Politécnica de Madrid), specially to Radar and Microwave Group (GMR) and Radiation Group (GR) for the installations access, resources and professors who always shared their knowledge with us.

To Manuel Sierra Castañer, who from the beginning accepted the challenge of leading this work, the development process has not been easy, however with his advice and knowledge has been possible to reach a satisfactory conclusion on this project adventure.

To José Manuel Fernandez, who joined this adventure from the beginning, because without his support, knowledge, dedication and unbreakable constancy would not have been possible to complete it.

Our deepest appreciation to both, thank you for showing us the way, for the trust placed in us, for being part of the whole process, for that enthusiasm and optimism to overcome the issues, thanks for your geniality who did the work flows and accelerated the development of this master thesis.
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Abstract

This book compiles the complete design process of a Phased Array Antenna for a coastal surveillance radar application in X band (8.8 – 9.6 GHz). The mentioned project is presented in order to accomplish all the requirements for the graduation program “Master en Radar, Tecnologías, Equipos y Diseño de Sistemas” at Universidad Politécnica de Madrid. In the first chapter, an introduction of phased array technology and a comparison of diverse coastal surveillance radar products available in the market, are presented. In the second chapter, general and specific objectives are listed. In the third chapter, the concept of antenna arrays and the principle of phased array operation, is introduced. In the fourth chapter, the complete phased array design is described, detailing the main parameters of azimuth and elevation array design, the influence of quantification error, a summary of phase sub array techniques and limitations, the radiating array element design, the distribution network design and finally its integration. In the fifth chapter, four elements sub array prototype is designed, implemented and measured. Finally, the conclusions and future works are presented.
Key Words

Antennas Design, Phased Array, Radar Antennas, Electronic Scanning
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Chapter 1

Introduction

In Radar systems, the antenna has the function to transmit the electromagnetic energy through the medium and collect the reflected energy of a distant target. In order to do the described process in an efficient way, the device must have an appropriate impedance matching, high gain, low side-lobes level and guarantee a precise angular resolution enough for the application. For that reason, it is almost mandatory to use large aperture antennas, commonly reflectors and large arrays to achieve the mentioned specifications. Antenna arrays have the capability to steer their beam electronically, avoiding the mechanical wear and making possible a dynamic and efficient beam control, one of the fundamental characteristics of modern radar systems. These kind of antennas are the Phased Arrays, its method of operation is primarily based on introducing an appropriate phase difference between the adjacent array elements, so the desired beam tilt is achieved.

Historically, the phased array technology has been mainly impulsed by satellite surveillance and ballistic missile defense applications, although with the accelerated development of solid state active elements and the MIMIC (Millimeter and Microwave Monolithic Integrated Circuits) industry growth, has been possible to use this technology on ground based and airborne radar applications as well. Nowadays, new fields of interest have emerged, photonic beam forming, micro-electromechanical phase shifters and space-time adaptive processing arrays, allowing the progress of phase array technology and its adaptation to the new challenges that civilian and military industry impose [1].

A Phased Array Antenna design for a coastal surveillance radar application in X band (8.8–9.6 GHz) is presented, the design specifications were based on the IALA (International Association of Marine Aids To Navigation and Lighthouse Authorities) recommendations for VTS (Vessel Traffic Services) systems and a comparison of diverse coastal surveillance radar products available in the market. It is remarkable that most of the consulted products use mechanical scanning instead of an electronic one, therefore the importance of the present work as an appropriation of the design technology and a business opportunity. The common parameters found on the consulted systems are listed on table 1.1.
<table>
<thead>
<tr>
<th>Parameter</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\theta_{-3dB}$ Azimuth</td>
<td>0.4° - 2.3°</td>
</tr>
<tr>
<td>$\theta_{-3dB}$ Elevation</td>
<td>Max 25°</td>
</tr>
<tr>
<td>Radiation Pattern</td>
<td>Fan, CSC^2-Inv</td>
</tr>
<tr>
<td>Technology</td>
<td>Micro-strip, Waveguide</td>
</tr>
<tr>
<td>Operation Bands</td>
<td>S, X, K, Ku, Ka</td>
</tr>
<tr>
<td>Bandwidth %</td>
<td>&lt;10%</td>
</tr>
<tr>
<td>Polarization</td>
<td>Circular, Linear</td>
</tr>
<tr>
<td>Gain</td>
<td>$&gt;$ 30 dB</td>
</tr>
<tr>
<td>SLL</td>
<td>$&gt;$ 25 dB</td>
</tr>
<tr>
<td>Range</td>
<td>20 – 40 mn</td>
</tr>
<tr>
<td>Signal</td>
<td>Pulsed, Continuous Wave (CW)</td>
</tr>
<tr>
<td>Scanning</td>
<td>Mechanical, Electronic</td>
</tr>
<tr>
<td>Duty Cycle</td>
<td>10 – 13%</td>
</tr>
<tr>
<td>Radiated Power</td>
<td>$P_p \approx$ 300W, $P_m \approx$ 5W</td>
</tr>
</tbody>
</table>

Table 1.1: Summary of diverse coastal surveillance radar products.
Chapter 2

Objectives

1. General Objective

To appropriate phased array antennas design techniques for radar applications.

2. Specific Objectives

- To design a phased array antenna for a coastal surveillance radar application in X band (8.8 – 9.6 GHz).
- To accomplish the following performance parameters:

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scanning</td>
<td>Electronic</td>
</tr>
<tr>
<td>Azimuthal Coverage</td>
<td>90º</td>
</tr>
<tr>
<td>Radiation Pattern</td>
<td>Fan</td>
</tr>
<tr>
<td>$\theta_{-3dB} \text{ Azimuth}$</td>
<td>0.7º-0.8º</td>
</tr>
<tr>
<td>$\theta_{-3dB} \text{ Elevation}$</td>
<td>20º-25º</td>
</tr>
<tr>
<td>Operation Bandwidth</td>
<td>8.8 – 9.6 GHz</td>
</tr>
<tr>
<td>Polarization</td>
<td>Linear Horizontal</td>
</tr>
<tr>
<td>Gain</td>
<td>$&gt; 30dB$</td>
</tr>
<tr>
<td>Side Lobes Level (SLL)</td>
<td>$&lt;-25dB$</td>
</tr>
<tr>
<td>Max Input Power</td>
<td>5W</td>
</tr>
</tbody>
</table>

Table 2.1: Objective Performance Parameters.
Chapter 3

Conceptual Framework

The following description pretends to introduce the concept of antenna arrays and the principle of phased array operation based on the references [2], [3].

1. Antenna Arrays

It is defined as an antenna grouping with a common feeding which radiate and receive the electromagnetic wave jointly; and depending of the amplitude and phase excitation of each element, a desired radiation pattern is obtained. This radiation pattern correspond to the Fourier transformation of a continuous distribution current, if the separation between elements is near to the half wavelength $\frac{\lambda}{2}$, the sampling theorem is accomplished and the array radiation pattern (Discrete Fourier Transformation of the sampled currents) can be modeled as the continuous distribution one. The array elements could be any kind of antenna, usually electrically small. The total radiated field can be obtained applying the superposition principle due to the Maxwell equations linearity as the summation of each element radiation field, this can be interpreted also as the multiplication of the array factor $F_A$ and the element radiation pattern $E_e$ which is presented in equations 3.1 and 3.2.

$$E_A(r, \theta, \phi) = E_e(r, \theta, \phi) \cdot F_A(\theta, \phi) \quad (3.1)$$

$$F_A(\theta, \phi) = \sum_n A_n e^{jkoz_n} \quad (3.2)$$

The antenna arrays can be classified based on its geometry as linear arrays, when the elements are grouped along a straight line, planar arrays, when the elements are grouped above a plane, and three-dimensional arrays, when the elements are placed on a volume.

2. Uniform Linear Arrays

In this case, the elements are placed along the z axis as is shown in figure 3.1, the array factor has revolution symmetry and only depends of $\theta$ angle.
CHAPTER 3. CONCEPTUAL FRAMEWORK

Figure 3.1: Linear Array scheme.

\[ F_A(\theta, \phi) = \sum_n A_n e^{jnk_0d \cdot \cos \theta} = \sum_n A_n e^{jn\Psi} \]  

(3.3)

From equation 3.3, it can be demonstrated that the array factor could be interpreted as the discrete Fourier transform of the array excitations. The DFT implements a transformation from spatial domain (positions and excitations) to an angular domain (radiation pattern). The array factor is a periodic function for \( \Psi \) variable of period \( 2\pi \). If a uniform amplitude for excitation \( (A_n = 1) \) is considered, the expression 3.3 can be replaced with expression 3.4 and a periodic sinc function is obtained.

\[ F_A(\Psi) = \sum_{n=0}^{N-1} e^{jn\Psi} = \frac{e^{jN\Psi} - 1}{e^{j\Psi} - 1} = e^{j\frac{N-1}{2}\Psi} \frac{\sin \frac{N\Psi}{2}}{\sin \frac{\Psi}{2}} \]  

(3.4)

The nulls that delimit the main beam are placed on \( \Psi = k_0d \cdot \cos \theta = \pm \frac{2\pi}{N} \), but in the practice is more useful to work with the half power beam width \( (\theta_{-3dB}) \), which for a high number of elements \( (N) \) is denoted as:

\[ \theta_{-3dB} \approx 8.886 \frac{\lambda}{Nd} \text{(rad)} \]  

(3.5)

The most important conclusion from equation 3.5 is to note that the half power beam width is inversely proportional to the number of elements \( (N) \) multiplied by the separation distance between them \( (d) \), and this is the total length of the array \( L = Nd \). As it was commented before for radar applications it is important to have narrow beams in order to achieve an appropriate angular resolution, therefore it is unavoidable to use large aperture antennas. On the other hand, from equation 3.4 it is possible to find that the first side lobe is located on \( \Psi = \frac{3\pi}{N} \) and for a high number of elements \( (N) \), the side lobes level \( (SLL) \) can be approximated to \(-13.5\,\text{dB} \), which is not enough to accomplish the specifications for radar applications, a possible solution is to use a different excitation as it will be explained later.

3. Linear Arrays with Progressive Phase Shift

This configuration is defined when the phase shift between two adjacent elements is constant \( (\alpha = \text{cte}) \). It allows to control the beam direction and is the operation principle of phased array antennas. In this case, it is necessary to adjust the expression 3.3 with the new variable change:

\[ \Psi = k_0d \cdot \cos \theta + \alpha \]  

(3.6)
As it was mentioned before, the array factor is a periodic function for $\Psi$ variable of period $2\pi$. However, the radiation pattern has limits that are determinate by the possible $\theta$ values on the real space, this range is defined as the visible margin. If inside this margin is included $\Psi = 0$, the maximum value of radiation is located on $\theta_{\text{max}}$ direction and it is defined by the following expression:

$$
\theta_{\text{max}} = \cos^{-1}\left(\frac{-\alpha}{k_o d}\right)
$$

Unfortunately $d$ and $\alpha$ values cannot be chosen too high because other replicas of the main beam (Grating Lobes) appear, this concept is analog to the sample theorem for digital signal processing but in the angular space domain. In the practice $d$ is limited between $0.6\lambda$ and $0.8\lambda$ for broadside arrays ($\theta_{\text{max}} = 90^\circ$) and the maximum steering angle is limited between $\pm 60^\circ$ respect to the broadside direction for exploration arrays.

4. Tapered Amplitude Linear Arrays

A symmetric and decreasing feeding from inside to outside is used to diminish the side lobe levels ($SLL$). This concept can be interpreted as a windowing procedure of the antenna excitation weights at the expense of increasing the main beam width. Therefore, a degradation of the array directivity is produced. Several windows have been proposed in the state of the art [4], Rectangular, Hann, Hamming, Blackman, Bartlett, Kaiser...etc, however for radar applications the compromise of main beam width ($\theta_{-3dB}$) and side lobe levels ($SLL$) are directly related with the angular resolution and target masking concepts. For this reason, what it is mostly used in the practice is to fix the required $SLL$ choosing the appropriate window and configure the number of array elements in order to achieve the desired main beam width. It is common to find Taylor based feedings for radar applications due to its outstanding compromise and the capability to adjust the $SLL$ with the following expressions for a broadside array [5]:

$$
b = \frac{6(R + 12)}{155}
$$

$$
\theta_{-3dB} \approx 8.868 \frac{\lambda \cdot b}{Nd} \text{ (rad)}
$$

Note: $R$ represents the desired side lobe level in dB and absolute units, and $b$ can be interpreted as the half main beam width broadening factor.
Chapter 4

Design

1. Azimuth and Elevation Array Design

Several synthesis methods for arrays design are presented in [5], the most known of them include:

- Schelkunoff’s Zero Placement Method
- With Windowing Fourier Series Method
- Woodward-Lawson Frequency-Sampling Design
- Narrow-Beam Low-Side lobe Design Methods
- Multi-Beam Array Design

In general, these methods are used to achieve a defined radiation pattern; however, it is also possible to design through the direct synthesis method, which it will be used for this phased array design since the initial specifications do not require a defined radiation pattern.

It is important to remember that the antenna power gain is directly related to the directivity and indirectly related to the effective antenna aperture. Therefore, it will be necessary to achieve the greatest possible aperture through the combination of two parameters: the radiating elements separation, and the number of necessary elements, and taking into account the other requirements.

The half power beam width was defined in equation 3.5. Studying this function behavior, it was observed that for high values of \( N \), an asymptotic behavior is obtained and therefore, the increase of the number of elements do not reduce the array beam width significantly. On the other way, the elements separation is also influenced by the effects of spurious radiation into other undesired directions, therefore it is necessary to reach a compromise between the maximum possible distance separation between elements and the minimum quantity of them, needed to meet the design goals. In figure 4.1 the steps of the design process are presented.
In order to achieve the design goals, it is important to establish first the elements separation, so that would achieve the desired margin of angular sweep. It must be considered that the maximum spacing is limited by the occurrence of grating lobes at the ends of the scanning range, if \( d < \frac{\lambda}{2} \) and/or phase \( \alpha \) between elements is large, replications of the main beam within the visible range could be produced as mentioned in the previous chapter.

\[
d_{\text{max}} < \frac{\lambda}{1 + \sin(\theta)} \tag{4.1}
\]

The equation for maximum spacing is function of the operation wavelength and the maximum valid angle as shown in equation 4.1. In the figure 4.2 it is shown that elements separation is determined by equation 4.1, and based on the figure, a first approximation could be done in order to choose the maximum \( d \) value that allows the desired electronic scanning. For electronic scanning range equal to \( \pm 45 \) degrees, the maximum separation between elements is found was 0.58\( \lambda \). It is important to observe the behavior of the beam width depending of the elements separation to understand its effects (figure 4.3).
Previously, it was realized that the beam width of the antenna is inversely proportional to its aperture size; therefore, to obtain a narrow beam is required to increase the physical size of the antenna, although some design considerations must be taken into account. First, it is important to determine if it will be electronic exploration, because according to this, for arrays without electronically scanning, the separation between the elements, may be in most cases between $0.6\lambda \leq d \leq 0.8\lambda$, but for an electronic scan array, this separation is usually between $0.4\lambda \leq d \leq 0.6\lambda$, this separation is the highest possible without undesired effects within the visible range. It is true that in some cases for specific applications, the grating lobes are desired, but in most cases, and for this particular application, it will not be.

\[
ed_{\text{max}} = \frac{1}{1 + \sin(\theta)}
\]

Figure 4.2: Maximum separation between elements

\[
\Delta \theta = \frac{\pi \theta_{3dB}}{2d}
\]

Figure 4.3: HPBW for different values of separation between elements
Another factor that directly affects the beam width of the array radiation pattern, is the number of necessary elements to achieve the specifications. As discussed above, the half power beam width ($\theta_{-3dB}$) is inversely proportional to the length of the array: once the separation of the elements is fixed, it will be necessary to determine the smallest feasible amount of elements to achieve the design goals. This is important to note, the largest numbers of elements, the narrowest is the beam, but the feeding network design will be more complex and also the losses will be higher. Then it is necessary to find a compromise between the number of elements and complexity of the feeding network, it is important to find if the total number of elements is a multiple factor of 2, 3 or even 5, so in this way it would be easier to design the network feeding. The following table summarizes the obtained values in order to fulfill the specifications.

<table>
<thead>
<tr>
<th>Number of Elements ($N$)</th>
<th>Elevation</th>
<th>Azimuth</th>
</tr>
</thead>
<tbody>
<tr>
<td>5</td>
<td>144</td>
<td></td>
</tr>
<tr>
<td>Separation ($d$)</td>
<td>0.7(\lambda)</td>
<td>0.54(\lambda)</td>
</tr>
</tbody>
</table>

Table 4.1: Number of elements and spacing.

The effect of increasing the number of elements is presented in figure 4.4: it can be observed that as the number of elements increase over the array, the ($\theta_{-3dB}$) becomes smaller. In the figure 4.5 is presented the elevation and azimuthal array radiation pattern using the direct synthesis method, note that in the design only have been considered the ($\theta_{-3dB}$) parameter as far, even so it is important for applications in radar to have control over the side lobes level, therefore it will be important to use a technique to reduce it to acceptable levels according to design specifications.

Figure 4.4: Number of Elements Vs Half Power Beam Width.
Windowing is a possible solution in order to reduce the side lobes level, with a weighting in the output delivered to each of the elements, it is possible to control the side lobe ratio, although there is not a particular rule which should be followed to use a window or another, it is important to note that each element should be fed by a percentage of the total power, so it should not be any element without power. A well known window family is the cosine type on a pedestal, the power delivered to each element decreases progressively and gradually from the center toward the end of the array. In the figure 4.6 is presented a comparison between different windows types for a 20 elements array.

It is important to note that if the side lobes level is reduced, an enlargement of the half power beam width ($\theta_{-3dB}$) is obtained. Due to the design specifications, in order to achieve a $SLR$ (Side Lobe Ratio) higher than 25 $dB$, the Taylor’s window was chosen, this is the
one that least degrades the $\theta_{-3dB}$ parameter keeping the side lobes level low, although the power losses have not been shown in the figure due to the normalization of the radiation pattern. Further comparisons and specifications for other windows can be consulted on [4].

One of the design requirements is to have a SLR greater than $25\,dB$, but this is impossible to achieve without a side lobes reduction technique. In figure 4.7 a comparison between Rectangular and Taylor window is presented, it is important to note that as mentioned before the half power beam width effect ($\theta_{-3dB}$) has increased according to equations 3.8 and 3.9.

Figure 4.7: Tapered Array Factor.

Furthermore, a large number of elements per row was obtained (144), so it is important to reduce distribution network complexity, even losing the achieved benefits with the side lobes reduction technique. Ideally, in order to achieve a good performance with the Taylor weighting, each element should receive the appropriate power percentage and diminish gradually between adjacent elements. However, this could complicate the power dividers design, for this reason it is important to implement a power sub array configuration. The power delivered to each sub array follows a similar Taylor weighting but keeping constant the power delivered to each element of sub array as it is shown in figure 4.8.

So far, the design has focused on operating parameters specified in the requirements, though it is important to note that only has been taken into account the behavior on the center frequency of the operating band (9.2 GHz), but in the frequency band extremes and maximum steering angles, other behaviors are expected, these considerations were taken into account in order to maintain the system performance properly. In figure 4.9 the final array factor on broadside direction is presented. The array factor at the center, maximum and minimum frequency of the operating band, steering at 45º and 135º is presented in figure 4.10, it is possible to appreciate that even at the extreme frequencies and steering angles, the specifications are still accomplished.
Figure 4.8: Azimuth Radiation Pattern with Amplitude Sub Array.

Figure 4.9: Array Factor Broadside Direction.

Figure 4.10: Azimuth Maximum Steering Angles.
2. Phase Shift Quantification Error

It is important to take into account the quantification error produced by the implemented phase shifters. This effect is generated due to the limited number of phases of this devices. The author in [6] defined \( \delta^2 \) as the phase root mean square error, being \( \delta \) a random variable with uniform distribution \( [-\pi/2, \pi/2] \), \( B \) is the phase shifter number of bits.

\[
\delta^2 = \frac{\pi^2}{3(2)^B}
\] (4.2)

The first effect is perceived as a gain loss at the main lobe, as it can be noted in equation 4.2, it will depend on the phase shifter number of bits. The obtained gain can be calculated using the following expression:

\[
G = G_0 \left(1 - \delta^2\right)
\] (4.3)

However, for more than 4 bits the losses are negligible as it is shown in the following table:

<table>
<thead>
<tr>
<th>( B )</th>
<th>( G )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>( \infty )</td>
</tr>
<tr>
<td>2</td>
<td>(-1 dB)</td>
</tr>
<tr>
<td>3</td>
<td>(-0.23 dB)</td>
</tr>
<tr>
<td>4</td>
<td>(-0.056 dB)</td>
</tr>
<tr>
<td>5</td>
<td>(-0.014 dB)</td>
</tr>
<tr>
<td>6</td>
<td>(-0.0034 dB)</td>
</tr>
</tbody>
</table>

Table 4.2: Phase Shift Quantification Error Gain Losses.

In the practice, the real quantification error distribution is not random, it is periodic, the consequence is the rise of undesirable quantification side lobes. The peak value of these quantification lobes is determined by 4.4 when they are pointing near the broadside direction:

\[
V_{peakQSL} \approx \frac{1}{2^{2B}}
\] (4.4)

For phase shifters with more than 4 bits, the peak value of these quantification lobes are suitable for most of the applications as it is shown in table 4.2.

<table>
<thead>
<tr>
<th>( B )</th>
<th>( V_{peakQSL dB} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>(-12)</td>
</tr>
<tr>
<td>3</td>
<td>(-18)</td>
</tr>
<tr>
<td>4</td>
<td>(-24)</td>
</tr>
<tr>
<td>5</td>
<td>(-30)</td>
</tr>
<tr>
<td>6</td>
<td>(-36)</td>
</tr>
</tbody>
</table>

Table 4.3: Peak Value of Quantification Side Lobes.

The phase quantification produces also a maximum pointing deviation in the main beam. This error depends on the phase shifter number of bits \( (B) \) and the half power beam width \( (\theta_{-3dB}) \).
CHAPTER 4. DESIGN

\[ \delta_\theta = \theta_{-3dB} \cdot \frac{\pi}{4} \cdot \frac{1}{2^B} \]  

(4.5)

<table>
<thead>
<tr>
<th>( B )</th>
<th>( \delta_\theta )</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>19.6%</td>
</tr>
<tr>
<td>3</td>
<td>9.8%</td>
</tr>
<tr>
<td>4</td>
<td>4.9%</td>
</tr>
<tr>
<td>5</td>
<td>2.45%</td>
</tr>
<tr>
<td>6</td>
<td>1.22%</td>
</tr>
</tbody>
</table>

Table 4.4: Maximum Pointing Deviation Error.

Finally, a simulation was performed to evidence the mentioned effects. The main lobe is steered near 45\(^\circ\) and as mentioned before, with 4 bits phase shifter the side lobe level is below \(-25dB\), necessary condition in order to accomplish the design specifications.

![Azimuth Radiation Pattern with Phase Shift Quantification Error](image)

Figure 4.11: Azimuth Radiation Pattern with Phase Shift Quantification Error.

3. Phase Sub Array Techniques and Limitations

The conventional phased array architecture contemplates one phase shifter per radiating element as is depicted in figure 4.12a, thereby the steering is controlled by changing the phase shift between adjacent radiating elements. However, this model could be expensive for large arrays, increasing the total price of the antenna. An alternative architecture have been proposed in the literature [7], it uses one phase shifter per radiating sub-array, preserving the distance \( d \) between adjacent radiating elements as it is shown in figure 4.12b. The phase in each sub-array is constant and the steering is controlled by changing the phase shift between adjacent sub-arrays. Unfortunately, this technique drastically increase the side lobes level, a sensitive parameter for radar applications.
Diverse authors have proposed different solutions to overcome this problem, the first one presented a solution based on overlapping sub-arrays in order to improve the side lobes level performance, therefore the distribution of each sub-array is extended to the adjacent sub-arrays, at the expense of complicating the distribution network design, in addition, with the use of digital time delays on each sub-array an increase in the instantaneous bandwidth was achieved [8]. The second one, presented a combination of tapered elements and sub-array excitations, although the sub-array phase error was not corrected at all [9]. Finally a different random sub-array size was proposed, however the side lobes level did decrease below $-20\,\text{dB}$ with the implemented technique [10].

Nowadays, this situation has not been solved in its totality and the academy continues looking for a solution suitable for sensitive applications. In order to accomplish the side lobe levels requirement for the present design, one phase shifter per radiating element will be used as it is shown in figure 4.12a. Finally, a simulation was performed in order to show the mentioned side lobes level increase effect, the results are presented in figure 4.13.
4. Array Element

4.1. Folded Dipole Design

The dipole is almost the most known antennas array element due to its feasible behavior and fabrication facilities, it can be manufactured by wire or printed technologies and through the strips width increase a higher frequency bandwidth could be achieved. A folded dipole was chosen as the radiating element for this design, it consists on two interconnected arms, one of them is directly connected to the feeding. The input impedance of the folded dipole is four times higher than a conventional dipole, the arms width control the impedance adjustment, however the folded dipole presents higher frequency bandwidth than the conventional one. The design process is presented in figure 4.14.

![Element Design Process](image)
Figure 4.15: Folded Dipole Geometry.

The folded dipole geometry was implemented as shown in figure 4.15 over a Rogers substrate (RO4350B), the element model was implemented, simulated and optimized on CST Studio Suite software and the following results were achieved.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Impedance (Zin)</td>
<td>$104.5 \Omega$</td>
</tr>
<tr>
<td>Dipole Length (L)</td>
<td>$10.8 , mm$</td>
</tr>
<tr>
<td>Arms Width (W)</td>
<td>$1 , mm$</td>
</tr>
<tr>
<td>Arms Separation (S)</td>
<td>$0.6 , mm$</td>
</tr>
<tr>
<td>Input Line Width (a)</td>
<td>$0.6 , mm$</td>
</tr>
<tr>
<td>Input Line Separation (Gap)</td>
<td>$0.6 , mm$</td>
</tr>
<tr>
<td>Input Line Length (L_alim)</td>
<td>$5 , mm$</td>
</tr>
<tr>
<td>Dielectric Constant (Er)</td>
<td>3.66</td>
</tr>
<tr>
<td>Substrate Height (H)</td>
<td>$0.5 , mm$</td>
</tr>
<tr>
<td>Copper Thickness (T)</td>
<td>$0.018 , mm$</td>
</tr>
<tr>
<td>Loss Tangent</td>
<td>$0.0037 , @10,GHz$</td>
</tr>
</tbody>
</table>

Table 4.5: Folded Dipole Design Parameters.

Figure 4.16: Folded Dipole Implemented Model.
As it can be noted, the element presents a directive pattern, about $7.7\, \text{dB}_i$ of gain on the maximum radiation direction and the $S_{11}$ coefficient is below $-20\, \text{dB}$ over all the operation frequency band ($8.8 - 9.6\, \text{GHz}$), the length of the dipole arms (10.8 mm) is near to $0.33\lambda_0$, facilitating the spatial distribution over the array. The presented results demonstrate it is an appropriate radiation element for the array.

4.2. Balun Design

In order to achieve $50\, \Omega$ impedance coupling at the antenna input it is necessary to use a balun element. It consists of a transformation of a balanced transmission line into an unbalanced one. The objective is to achieve $180^\circ$ of phase difference between the parallel transmission lines and then use a quarter wavelength impedance transformer as is depicted on figure 4.19.
The corner bend optimization was performed on CST Studio Suite software, it was found that transmission coefficient ($S_{21}$) is maximum for $B = 1.25 \cdot W$ as is demonstrated in figure 4.20, these bends help to reduce the reflected and diffracted waves over the corners [11].

The balun and the folded dipole were jointly optimized in order to obtain an appropriate coupling over the frequency bandwidth keeping a stable radiation pattern.
The S11 coefficient is below $-20 \, dB$ over all the frequency bandwidth as is depicted in figure 4.22, both resonances can be noted as is expected.

The radiation pattern is still directive and the maximum power gain is near $5.45 \, dBi$. However, the spurious radiation of the balun transmission lines produced a shift in the elevation main beam angle which is shown in figure 4.23c. A possible solution to overcome this issue is to implement the model on stripline technology, unfortunately it could complicate the manufacture and the elements soldering.
Actually, this angle shift on the elevation pattern has no transcendence over the total array elevation radiation pattern; it is important to remember that the array radiation pattern is the multiplication of the array factor and the element radiation pattern. In this case, the half power beam width of the element elevation radiation pattern (Balun + Folded Dipole) is about 188°; therefore, the total radiation pattern is principally affected by the array factor as it is demonstrated on figure 4.24, where five elements with uniform excitation and spaced 0.7λ along the Z axis were simulated, obtaining the 90° pointing, as it was
expected. Finally, the designed balun and folded dipole complete dimensions are presented in figure 4.25.

![Farfield (Array) Directivity Abs (Phi=90)](image)

**Figure 4.24:** Balun + Folded Dipole Elevation Radiation Pattern ($N = 5$).

![Designed Balun + Folded Dipole Dimensions [mm]](image)

**Figure 4.25:** Designed Balun + Folded Dipole Dimensions [mm].
5. Distribution Network

The distribution network was designed in order to achieve the chosen Taylor weighting with the purpose of side lobes reduction. First, the number of elements was established, the array consists of 5 rows with 144 elements in each one, this values were chosen because it was a sufficient number to accomplish the design requirements and the distribution network was easy of fabricate. Therefore, it was necessary to design the complete distribution network for one row, 144 elements, and the complete distribution network to interconnect the five rows. Then, the distribution architecture was established, this is how will be interconnected the different elements using 2-way and 3-way power dividers. Next, the power distribution was calculated and each power divider was designed, simulated and optimized using microstrip technology over the Rogers substrate (RO4350B) (\( \epsilon_r = 3.66 \)), also used for the element design. Finally, the power dividers were integrated and the distribution network was simulated and optimized in order to achieve the specifications. The design process is summarized in figure 4.26.

![Figure 4.26: Distribution Network Design Process.](image-url)
5.1. Elements Distribution Network

The distribution network for each row was designed using 12 uniform power sub arrays, 4 asymmetric 3-way power dividers, 2 asymmetric 2-way power dividers and 1 symmetric power divider as it is presented in figure 4.27.

![Elements Distribution Network Architecture](image)

The Taylor amplitude distribution for 12 elements grouping \((12 \cdot 12 = 144)\) is presented in table 4.6.

<table>
<thead>
<tr>
<th>Amplitude</th>
<th>Amplitude [dB]</th>
<th>Power Percentage %</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.264</td>
<td>-11.6</td>
<td>3</td>
</tr>
<tr>
<td>0.381</td>
<td>-8.4</td>
<td>5</td>
</tr>
<tr>
<td>0.564</td>
<td>-5</td>
<td>7</td>
</tr>
<tr>
<td>0.753</td>
<td>-2.5</td>
<td>10</td>
</tr>
<tr>
<td>0.902</td>
<td>-0.9</td>
<td>12</td>
</tr>
<tr>
<td>0.985</td>
<td>-0.1</td>
<td>13</td>
</tr>
<tr>
<td>0.985</td>
<td>-0.1</td>
<td>13</td>
</tr>
<tr>
<td>0.902</td>
<td>-0.9</td>
<td>12</td>
</tr>
<tr>
<td>0.753</td>
<td>-2.5</td>
<td>10</td>
</tr>
<tr>
<td>0.564</td>
<td>-5</td>
<td>7</td>
</tr>
<tr>
<td>0.381</td>
<td>-8.4</td>
<td>5</td>
</tr>
<tr>
<td>0.264</td>
<td>-11.6</td>
<td>3</td>
</tr>
</tbody>
</table>

Table 4.6: Taylor Amplitude Distribution for 12 Elements.

The power dividers were calculated based on the Wilkinson dividers architecture as is depicted in figure 4.28. The characteristic impedance of each transmission line was calculated.
using the following expressions taken from [12].

\[ Z_{0A} = Z_0 \cdot \left( \left( \frac{P_A}{P_B} \right)^{-1.5} + \left( \frac{P_A}{P_B} \right)^{-0.5} \right)^{0.5} \]  
(4.6)

\[ Z_{0B} = Z_0 \cdot \left( 1 + \frac{P_A}{P_B} \right)^{0.5} \cdot \left( \frac{P_A}{P_B} \right)^{-0.25} \]  
(4.7)

\[ Z_{0C} = Z_0 \cdot \left( \frac{P_A}{P_B} \right)^{-0.25} \]  
(4.8)

\[ Z_{0D} = Z_0 \cdot \left( \left( \frac{P_A}{P_B} \right)^{0.5} + \left( \frac{P_A}{P_B} \right)^{-0.5} \right) \]  
(4.9)

The resistor was not used in order to simplify the physical distribution network design, although this produce an impedance mismatch in the output ports. For that reason is necessary to optimize and readjust the physical parameters of the microstrip transmission lines (width and length) to get the appropriate impedance value at the input and output ports 50\(\Omega\). Starting on a 4 elements sub array design which will be presented in the next chapter, each power divider design will be described with the respective results.

### Symmetric 3-Way Power Divider (33.3%-33.3%-33.3%)

The power divider was modeled, simulated and optimized on CST Studio Design tool, each branch consist of a quarter wave length transformer, \(Z_0 = 50\Omega\), \(Z_n = \sqrt{N} \cdot Z_0\). In this case we have 3 ports so \(N = 3\) then \(Z_n = \sqrt{3} \cdot 50 \approx 86.6\Omega\). The implemented model is presented in figure 4.29 and the relevant parameters are depicted in table 4.7.
Finally, the simulation results are shown in figure 4.30. The ports impedance is 52.2\( \Omega \) at 9.2\( GHz \), the S11 coefficient is below \(-20 dB\) over all the frequency band and the transmission coefficients are balanced at 9.2\( GHz \).
Asymmetric 3-Way Power Divider (22%-31%-47%)

The power divider was modeled, simulated and optimized on CST Studio Design tool, it is based on the 3-way Wilkinson divider architecture. The implemented model is presented in figure 4.31 and the relevant parameters are depicted in table 4.8.
Finally, the simulation results are shown in figure 4.32. The ports impedance are equal to 50.125Ω and 50.001Ω at 9.2GHz, the S11 coefficient is below −25dB over all the frequency band and the transmission coefficients are around 0.46, 0.56 and 0.68 at 9.2GHz respectively.
Asymmetric 3-Way Power Divider (29%-34%-37%)

The power divider was modeled, simulated and optimized on CST Studio Design tool, it is based on the 3-way Wilkinson divider architecture. The implemented model is presented in figure 4.33 and the relevant parameters are depicted in table 4.9.
Finally, the simulation results are shown in figure 4.34. The ports impedance are equal to 50.125 Ω and 50 Ω at 9.2 GHz, the S11 coefficient is below -20 dB over all the frequency band and the transmission coefficients are around 0.54, 0.58 and 0.61 at 9.2 GHz respectively.
Asymmetric 2-Way Power Divider (31%-69%)

The power divider was modeled, simulated and optimized on CST Studio Design tool, it is based on the 2-way Wilkinson divider architecture. The implemented model is presented in figure 4.35 and the relevant parameters are depicted in table 4.10.
### Table 4.10: Asymmetric 2-Way Power Divider Parameters (31%-69%).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>H</td>
<td>0.5 mm</td>
<td>Substrate Height</td>
</tr>
<tr>
<td>T</td>
<td>0.018 mm</td>
<td>Copper Thickness</td>
</tr>
<tr>
<td>T_Ground</td>
<td>2 mm</td>
<td>Ground Thickness</td>
</tr>
<tr>
<td>WZ0</td>
<td>1.07 mm</td>
<td>Z0 Line Width</td>
</tr>
<tr>
<td>WZ0A</td>
<td>0.1 mm</td>
<td>Z0A Line Width</td>
</tr>
<tr>
<td>WZ0B</td>
<td>0.68 mm</td>
<td>Z0B Line Width</td>
</tr>
<tr>
<td>WZ0C</td>
<td>1.07 mm</td>
<td>Z0C Line Width</td>
</tr>
<tr>
<td>WZ0D</td>
<td>1.07 mm</td>
<td>Z0D Line Width</td>
</tr>
<tr>
<td>LZ0</td>
<td>5.06 mm</td>
<td>Z0 Line Length</td>
</tr>
<tr>
<td>LZ0A</td>
<td>7.77 mm</td>
<td>Z0A Line Length</td>
</tr>
<tr>
<td>LZ0B</td>
<td>2.97 mm</td>
<td>Z0B Line Length</td>
</tr>
</tbody>
</table>

Finally, the simulation results are shown in figure 4.36. The ports impedance are equal to 50.125 Ω at 9.2 GHz, the S11 coefficient is below −20 dB over all the frequency band and the transmission coefficients are around 0.56 and 0.83 at 9.2 GHz respectively.

![Figure 4.36: Asymmetric 2-Way Power Divider Design Results (31%-69%).](image)

### Symmetric 2-Way Power Divider (50%-50%)

The power divider was modeled, simulated and optimized on CST Studio Design tool, it is based on the 2-way Wilkinson divider architecture. The implemented model is presented in figure 4.37 and the relevant parameters are depicted in table 4.11.
Finally, the simulation results are shown in figure 4.38. The ports impedance are equal to 50.125Ω at 9.2 GHz, the S11 coefficient is below −25 dB over all the frequency band and the transmission coefficients are around 0.706 at 9.2 GHz respectively.

Table 4.11: Symmetric 2-Way Power Divider Parameters (50%-50%).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>H</td>
<td>0.5 mm</td>
<td>Substrate Height</td>
</tr>
<tr>
<td>T</td>
<td>0.018 mm</td>
<td>Copper Thickness</td>
</tr>
<tr>
<td>T_Ground</td>
<td>2 mm</td>
<td>Ground Thickness</td>
</tr>
<tr>
<td>WZ0</td>
<td>1.07 mm</td>
<td>Z0 Line Width</td>
</tr>
<tr>
<td>WZ0A</td>
<td>0.55 mm</td>
<td>Z0A Line Width</td>
</tr>
<tr>
<td>WZ0B</td>
<td>0.07 mm</td>
<td>Z0B Line Width</td>
</tr>
<tr>
<td>WZ0C</td>
<td>1.07 mm</td>
<td>Z0C Line Width</td>
</tr>
<tr>
<td>WZ0D</td>
<td>1.07 mm</td>
<td>Z0D Line Width</td>
</tr>
<tr>
<td>LZ0</td>
<td>5.1 mm</td>
<td>Z0 Line Length</td>
</tr>
<tr>
<td>LZ0A</td>
<td>4.82 mm</td>
<td>Z0A Line Length</td>
</tr>
<tr>
<td>LZ0B</td>
<td>4.82 mm</td>
<td>Z0B Line Length</td>
</tr>
</tbody>
</table>
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Network Integration

The designed power dividers were integrated with the respective radiating elements on CST Studio Design tool. The model is shown in figure 4.39, the row has 17.78 cm length, 1.31 m width and 2.5 mm height. The simulation results are shown in figure 4.40, the port impedance is 50.125 Ω at 9.2 GHz and the S11 coefficient is below −20 dB over all the frequency band. Finally, the designed 12 element subarray and the complete row are presented in figures 4.41 and 4.42, dimensions are shown in millimeters.
(a) Input Port Impedance.  
(b) S11 Coefficient

Figure 4.40: Row Model Results.

Figure 4.41: 12 Element Designed Sub Array.
Figure 4.42: Designed Complete Row.
5.2. Rows Distribution Network

The process design is the same as previous section, the numbers of rows was established as 5 in order to achieve the elevation half power beam width ($\theta_{-3dB}$) specification ($20^\circ - 25^\circ$). It was necessary to use Taylor amplitude weighting in order to decrease the side lobes level as well. The Taylor distribution power for 5 elements is presented in table 4.12 and the proposed distribution network architecture is presented in figure 4.43. A 3-way power divider and two 2-way power dividers were used.

<table>
<thead>
<tr>
<th># Row</th>
<th>Power Percentage %</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>10.36%</td>
</tr>
<tr>
<td>2</td>
<td>24.06%</td>
</tr>
<tr>
<td>3</td>
<td>31.16%</td>
</tr>
<tr>
<td>4</td>
<td>24.06%</td>
</tr>
<tr>
<td>5</td>
<td>10.36%</td>
</tr>
</tbody>
</table>

Table 4.12: Rows Power Percentage.

The distribution network was modeled, simulated and optimized on CST Studio Design tool over the same substrate, it is based on the 2-way and 3-way Wilkinson dividers architecture. The implemented model is presented in figure 4.44, the simulation results are shown in figure 4.45, the ports impedance are equal to $50\Omega$ at $9.2\,GHz$, the S11 coefficient is below $-20\,dB$ over all the frequency band and the transmission coefficients accomplish the power distribution percentage. Finally, the designed rows distribution network complete dimensions are presented in figure 4.46.
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Figure 4.45: Rows Distribution Network Results.

Figure 4.46: Rows Distribution Network Dimensions [mm].
Chapter 5

Results

A four elements sub array prototype was designed and fabricated in order to verify and validate the correct modeling of the complete phased array. Four elements were chosen due to its fabrication facilities, affordable cost, reduced size and it is enough to appreciate a relevant steering control. The radiation element (Folded Dipole) was the same designed in section 4. and the distribution network was designed using the same method depicted on section 5. over the same substrate (RO4350B) ($\epsilon_r = 3.66$). The distribution network is based on three 2-way T power dividers, the model was simulated and optimized first on CST Studio design and then on CST Microwave in order to take in account the electromagnetic couplings and effects. The distribution network model is presented in figure 5.1 and relevant parameters are depicted in table 5.1. Finally the simulation results are shown in figure 5.2.

![Figure 5.1: Prototype Distribution Network Model.](image)
### Table 5.1: Prototype Distribution Network Parameters.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>( H )</td>
<td>0.5 ( mm )</td>
<td>Substrate Height</td>
</tr>
<tr>
<td>( T )</td>
<td>0.018 ( mm )</td>
<td>Copper Thickness</td>
</tr>
<tr>
<td>( T_{\text{Ground}} )</td>
<td>2 ( mm )</td>
<td>Ground Thickness</td>
</tr>
<tr>
<td>( W_{50} )</td>
<td>1.07 ( mm )</td>
<td>( Z_{50} ) Line Width</td>
</tr>
<tr>
<td>( W_{70} )</td>
<td>0.64 ( mm )</td>
<td>( Z_{70} ) Line Width</td>
</tr>
<tr>
<td>( L_{70} )</td>
<td>4 ( mm )</td>
<td>( Z_{50} ) Line Length</td>
</tr>
</tbody>
</table>

The \( S_{11} \) coefficient is below \(-20 \, dB\) over all the frequency band, the ports impedance are near to 53.95\( \Omega \) and 53.65\( \Omega \) at 9.2 \( GHz \) and the transmission coefficients are \(-6.37 \, dB\) and \(-6.51 \, dB\). As noted, these results are appropriate in order to integrate the distribution network with the radiation elements.

The integration of the radiation elements and the distribution network was done using CST Studio Design, this tool computes the \( S \) parameters of the different modules and calculate the expected total \( S \) parameters. The circuital integrated model is depicted on figure 5.3 and the simulation results are shown in figure 5.4.
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Figure 5.3: Prototype Circuit Integrated Model.

Furthermore, the physical structure was modeled on CST Microwave in order to do an electromagnetic simulation and a parameters optimization. The simulations time increased drastically as well the results accuracy. The whole structure was assembled on a Rogers substrate (RO4350B) ($\epsilon_r = 3.66$) with 0.5 mm of thickness, the implemented model is presented in figure 5.5.
An appropriate impedance matching was achieved as depicted on figure 5.6, the S11 coefficient is below $-15\, dB$ over all the frequency bandwidth.
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The radiation pattern is presented in figures 5.7 and 5.8, the array presents a directive pattern, about 9.14 dBi of gain on the maximum radiation direction and 21.2° of azimuthal half power beam width. As is expected for uniform amplitude distribution the side lobes level are near −13.4 dB.

Figure 5.6: Prototype Integrated Physical Model Impedance Matching.

Figure 5.7: Prototype Integrated Physical Model 3D Radiation Pattern.

Figure 5.8: Prototype Integrated Physical Model 2D Radiation Pattern.
After completing the simulation and optimization in CST, the corresponding copper layer must be exported to a DXF file, it is important that the physical design dimensions be controlled throughout the PCB manufacturing design process using a CAD software, it is also necessary to generate a file format commonly used in industry, such as gerber files. The complete PCB design process is shown in figure 5.9 and the complete PCB dimensions are depicted in figure 5.10.

Figure 5.9: PCB Design Process.

Figure 5.10: Prototype PCB Dimensions [mm].
Manufactured PCB is presented in figure 5.11, total dimensions are 89\textit{mm} x 64\textit{mm} made by photo lithography process. The PCB manufacturer imposed some design restrictions as for example the lines width be greater than 0.2\textit{mm}, two layers, separation between lines 0.2\textit{mm}, ISO A4 maximum size, and vias diameter 0.4\textit{mm}.

![Manufactured PCB without components](image1)

(a) Front View  
(b) Rear View.

Figure 5.11: Manufactured PCB without components.

Afterward, the surface mounted devices were soldered to the manufactured PCB, 4 phase shifters (MAPS 010146), 8 capacitors (0.1\textit{µF}), 16 resistors (10\textit{kΩ}), 27 resistors (0Ω) and 4 switches in order to control the phase shifters. The 0Ω resistors were used to overcome some of the mentioned manufactured restrictions. A metallic support was made in order to facilitate the measurement process as depicted in figure 5.13.

![Manufactured PCB with components](image2)

Figure 5.12: Manufactured PCB with components.
The prototype impedance matching was measured with a vector network analyzer (VNA) and the results are shown in figure 5.14.

Figure 5.14: Prototype Measured Impedance Matching.
CHAPTER 5. RESULTS

Figure 5.15: Prototype Measured Detailed S11 Coefficient.

Poor impedance matching was observed, therefore in order to verify the simulation model depicted in figure 5.5, a passive prototype was manufactured as well, the results are presented in figure 5.17. Poor impedance matching was also achieved, demonstrating an inconsistency between simulation model results and prototype measurements, the summary is condensed in table 5.2.

Figure 5.16: Passive Manufactured Prototype.
CHAPTER 5. RESULTS

(a) Measured S11 Coefficient.  
(b) Measured Impedance Smith Chart.

Figure 5.17: Passive Prototype Measured Impedance Matching.

Table 5.2: Measurement Results.

<table>
<thead>
<tr>
<th>S11 [dB]</th>
<th>8.8 GHz</th>
<th>9.2 GHz</th>
<th>9.6 GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Measurement Active Prototype</td>
<td>-3.48</td>
<td>-4.04</td>
<td>-3.84</td>
</tr>
<tr>
<td>Measurement Passive Prototype</td>
<td>-5.04</td>
<td>-4.64</td>
<td>-3.86</td>
</tr>
</tbody>
</table>

The initial CST prototype model simulation was made with the time domain solver and the impedance matching results were previously presented in figure 5.6, then the same CST model simulation was made with the frequency domain solver and the following results were obtained.

Figure 5.18: Passive Prototype Measure and Simulation Comparison S11.
CHAPTER 5. RESULTS

The measured S11 result is much closer to the frequency domain solver solution, however it can be noted a large difference with the time domain one. The same comparison was done with the folded dipole model previously depicted in figure 4.21 and the distribution network model previously depicted in figure 4.44. The results are presented in figures 5.19 and 5.20.

![Figure 5.19: Dipole + Balun Simulation Model Comparison S11.](image1)

![Figure 5.20: Distribution Network Simulation Model Comparison S11.](image2)

As is observed the difference is appreciable in the frequency bandwidth (8.8 – 9.6 GHz), this inconsistency between both solvers solution is currently under studying and the case will be elevated to CST staff. It is expected to measure the radiation pattern and steering behavior by the time this work is presented.
Chapter 6

Conclusions

This master thesis aims to establish a phased array design methodology, taking into account the different possibilities and advantages of these types of arrays for radar applications, detailing the desired characteristics for such applications as the basic requirements as high angular resolution, high side lobe ratio, high gain, high directivity and low losses. Early studies for addressing this work showed that for this type of application is common to work in S, X, Ku, K, and Ka frequency bands, however it is important to achieve relatively large instrumental range, small physical dimensions and operating under different weather conditions, for this reason it was decided to work in X band. This operation band allows to work with large electrical apertures, reduced physical dimensions and achieve substantial instrumental ranges.

Most of the encountered commercial products for these applications are designed over microstrip or waveguide technology with azimuthal mechanical exploration. For that reason this work focused on the importance of electronic scanning arrays design.

The design of radiating systems with electronic scanning is well documented and there are different sources of information available. However, the principal purpose of this work is to establish an appropriate design methodology arrays with electronic scanning, although the design process did not follow a fixed sequence and required an iterative design process, investigating a suitable design methodology that could reduce design processes and lead to appropriate results.

In the first design phase was taken into consideration the specified parameters to be performed by the radiant system, in this line of thinking, it is important to note that there are design parameters which are more restrictive than others, so as a result of the design process was established that the elements separation is more restrictive that the number of them, so it is necessary to fix it as first and then continue with the others parameters.

To achieve the required side lobe ratio (SLR) design specifications, it is necessary to use some side lobes level reducing technique as the weighting in amplitude. In this work different window types were analyzed, their advantages and disadvantages in each case, it is important to note that window selection depends only on the particular application, so in this work the windows comparison was part of the design process methodology, a first approach to the amplitude weighting selection method should obey to the least characteristics degradation in previous design phases in order to meet the design requirements.

When carrying out the amplitude weighting selection, the feeding network complexity can be reduced implementing amplitude sub arrays techniques keeping stable parameters performance in order to achieve the required specifications. The possibility of phase sub arrays grouping was explored as a strategy to reduce the required phase shifters number
for the complete array, however, the problems associated with these techniques are not resolved yet and remains under worldwide investigation.

A printed folded dipole was chosen as the array radiating element due to its simplicity design, low cost, fabrication facilities and feasible integration with electronic circuits.

Finally, two prototypes were designed and fabricated in order to verify and validate the correct modeling of the complete phased array, however, poor impedance matching was achieved, demonstrating an inconsistency between time domain solver simulation model results and frequency domain solver simulation model results, the inconsistency between both solvers solution and measured results is currently under studying and the case will be elevated to CST staff.

Future works

• The exploration of different feeding topologies and techniques in order to reduce the design complexity.

• Analysis and research of sophisticated phased sub array grouping techniques as a possible solution to reduce the economic cost of these kind of arrays, without performance degradation.

• The study and implementation of diverse radiating elements alternatives as for example patches or slotted dipoles and its array performance influence.

• The consideration of different fabrication technologies as waveguide and stripline.

• The exploration of distinct amplitude weighting methods like variable attenuators or beam forming at the expense of increasing the array cost.

• A study of electromagnetic mutual coupling between RF and DC lines in single layer structures.

• A design of 3D exploration phased array for multifunction radar applications.

• Electronic control architecture implementation for entire phased array.
Bibliography


