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Design and Modeling of Circularly Polarized Conical Beam Antennas for Space Missions

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“The Cosmos is all that is or was or ever will be. Our feeblest contemplations of the Cosmos stir us – there is a tingling in the spine, a catch in the voice, a faint sensation, as if a distant memory, of falling from a height. We know we are approaching the greatest of mysteries.”

Carl Sagan

To
My Parents and Sister
For their undying love and support.

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Abstract

Antenna systems are integral part of the space missions as antennas act as the first point of contact for humans. Yet, space missions present antenna designers with a unique set of requirements. The space environment and the launch vehicle dynamics require the antennas to have a good balance between the electrical, thermal and mechanical performance of the antennas. This means that not only does the antenna need to perform optimally from the electromagnetic radiation point of view, but also need to survive launch conditions, perform optimally at extreme temperatures, deter ionizing radiation and provide electrostatic protection. A careful choice of the material, thus, becomes critical for all the antenna components. Moreover, without any prospect of repair, such antennas need to be robust enough to withstand the harsh environment on its own.

This thesis work touches upon such multi-physics approach to antenna design. More specifically, the antennas concern the ones for the Lander Radioscience (LaRa) instrument intended for the surface platform of the ExoMars 2020 mission to Mars. The antennas would allow direct communication with Earth irrespective of the orientation of the surface platform when it lands, and allow observation throughout the year. This is possible through the conical radiation patterns, which have uniform elevation pattern over the azimuthal plane along the surface platform. For space communications, it is a necessity to have circular polarization as well. Therefore, the

thesis proposes different topologies of circularly polarized conical beam antennas ideal for space communications, from planar antennas to 3D metallic antenna.

With regards to the planar antenna technology, a circular slot array and a circular patch antenna have been developed, capable of generating circularly polarized conical patterns. The multi-layered PCB structure requires a careful design given the need for space qualification, especially in terms of the thermoelastic effect. A proper choice of dielectric substrate will ensure that the mechanical deformations as a result of change in temperature will not greatly affect the electromagnetic performance. Metallic antennas are deemed more appropriate for such environmental conditions, since the homogeneous material properties of the metallic antennas ensure better performance under the thermoelastic stress. To that end, a novel 3D metallic antenna is proposed that generates a circularly polarized conical beam. It is based on gamma-shaped vertical posts surrounding a centrally radiating monopole antenna.

Moreover, a numerical method to analyze the near-field performance of the antenna has also been proposed in the thesis. Especially in the case of space antennas, whose level of detail requires a dense mesh, the proposed method is effective. The method aims to improve the near-interactions in the well known Method of Moments (MoM) by making the integrals separable with respect to source and observation coordinates. The Green's function in the MoM may be decomposed into static and the smooth parts, where the static part is treated with the static multipole and the smooth part with a Taylor's series expansion. Both the parts remain separable, which dramatically accelerates the near-field interactions. This approach opens new avenues for fast electromagnetic analysis of complex structures, such as antennas on space-based platforms.

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Acronyms

3D	Three Dimension
ADC	Analog-to-Digital Converter
AR	Axial Ratio
CAD	Computer Aided Design
CFRP	Carbon-Fibre Reinforced Plastic
CP	Circular Polarization
CST	Computer Simulation Technology
CTE	Coefficient of Thermal Expansion
ECSS	European Cooperation for Space Standardization
EFIE	Electric Field Integral Equation
EM	Engineering Model
EQM	Engineering Qualification Model
ESA	European Space Agency
ESD	Electrostatic Discharge
ExoMars	Exobiology on Mars
FIPWA	Fast Inhomogeneous Plane Wave Algorithm
FMM	Fast Multipole Method
GFRP	Glass-Fibre Reinforced Plastic
LaRa	Lander Radioscience
LHCP	Left Hand Circular Polarization
MBF	Macro Basis Function

MLFMA	Multi-Level Fast Multipole Algorithm
MoM	Method of Moments
MTF	Macro Testing Function
PEC	Perfect Electric Conductor
PCB	Printed Circuit Board
PFM	Proto Flight Model
PTH	Plated Through Hole
RF	Radio Frequency
RHCP	Right Hand Circular Polarization
RWG	Rao Wilton Glisson
RX	Receive
SAM	System Assembly and Modeling
SAR	Synthetic Aperture Radar
SDP	Steepest Descent Path
SMA	SubMiniature version A
SP4T	Single-Pole Four-Throw
SP5T	Single-Pole Five-Throw
TD	Time Domain
TID	Total Ionizing Dose
TM	Transverse Magnetic
TTD	True Time-Delay
TX	Transmit
VSWR	Voltage Standing Wave Ratio

Introduction

This introductory chapter deals with the motivation and structure behind this thesis work. The core of this work concerns the design and test of antennas intended for an instrument aboard the surface platform of the ExoMars 2020 mission to Mars. Moreover, a numerical method devoted to modeling the effect of detailed objects on an antenna platform is explored.

1.1 The Lander Radioscience Experiment

LaRa, or Lander Radioscience, is an instrument aboard the ExoMars 2020 mission to Mars in an endeavour to discover the signs of life in the red planet [1]. The ExoMars 2020 mission itself is a joint effort from Roscosmos and the European Space Agency (ESA), consisting of a Russian surface platform and a European rover, where LaRa is a part of the Russian platform. The LaRa instrument will fly in a Proton rocket for a nine-month long journey to Mars, eventually landing in 2021. The LaRa instrument will remain stationary on the surface of Mars while the rover continues to explore the possibilities of life with its numerous instruments.

The LaRa experiment aims to uncover the deep mystery of what constitutes the interior of Mars. Understanding the inner composition of the planet unlocks the possibility to study its origin and evolution. One useful

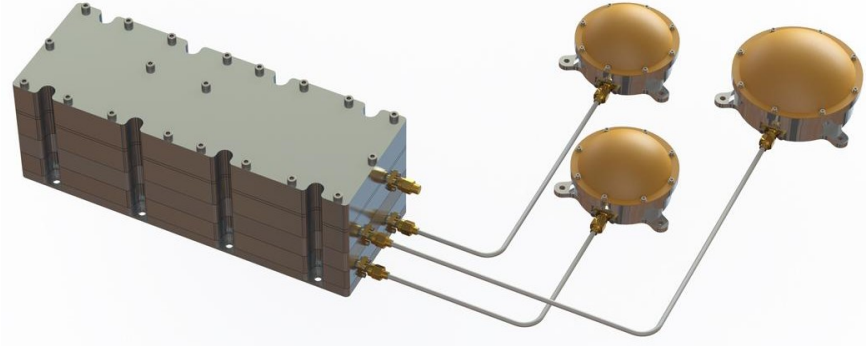


Figure 1.1: LaRa transponder and antennas [Model from AntwerpSpace].

tool through electromagnetic geodesic experiment [2] allows the possibility to remotely probe the interior by measuring accurately the rotation and orientation of the planet [3]. A useful analogy would be to observe the rotation of a raw egg as opposed to that of a boiled one. With a solid core, the boiled one rotates differently from the raw one with a liquid core [4].

An X-band signal generated by one of the masers in the Earth ground stations is transmitted and picked up by the LaRa receive antenna on Mars, coherently converted by the LaRa transponder to a higher frequency, and transmitted back to Earth. The receive (RX) antenna operates in the [7.145, 7.190] GHz band with a carrier frequency of 7.174 GHz, while the transmit (TX) antenna operates in the [8.4, 8.45] GHz band with a carrier frequency of 8.428 GHz. An assembly of the LaRa transponder with the antennas is shown in Fig. 1.1. There are two TX antennas (one being redundant) and one RX antenna. The antennas placement aboard the surface platform is shown in Fig. 1.2.

Given that the orientation of the surface platform with Earth is not known *a priori*, the LaRa antennas should allow for a full azimuthal coverage at low elevation angles. Therefore, the antennas should have maximum gain between the elevation angle of 35° to 45° from the surface platform. In terms of θ ($90^\circ - \text{elevation}$), the copolar gain should be better than 4 dB for $35^\circ \leq \theta \leq 60^\circ$ and 3 dB for $60^\circ \leq \theta \leq 70^\circ$. The axial ratio (AR) should be lower than 3 dB for $35^\circ \leq \theta \leq 70^\circ$. Table 1.1 summarizes the specifications.

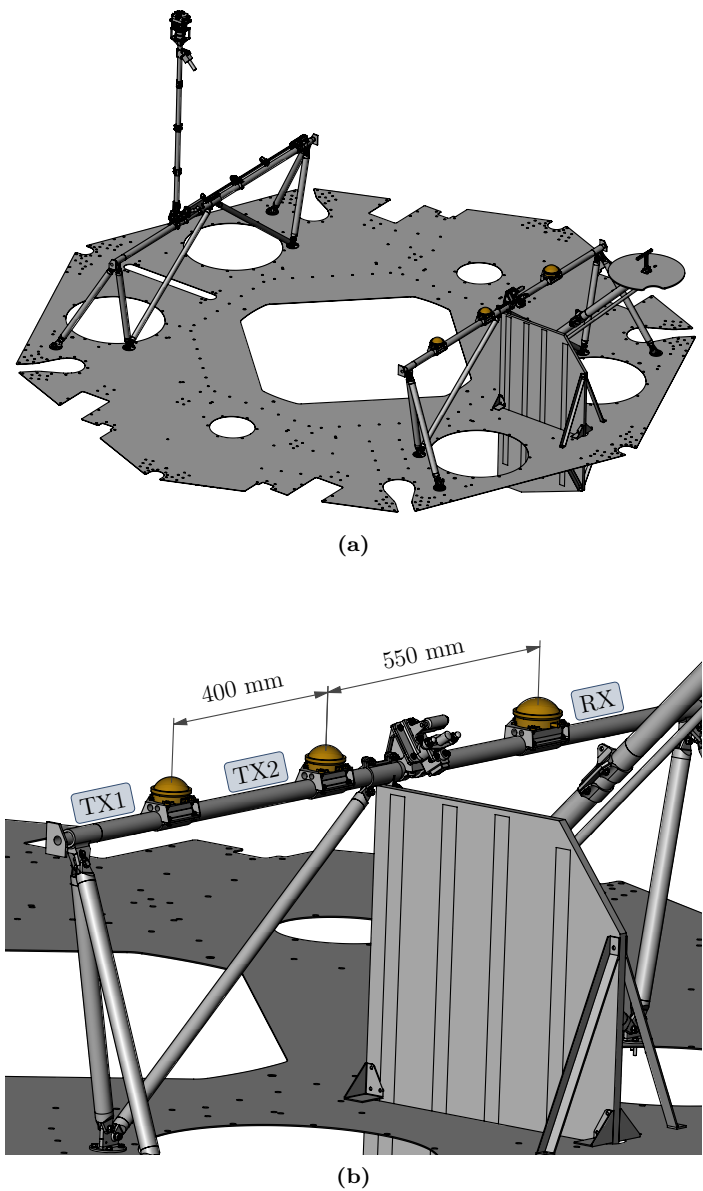


Figure 1.2: Surface platform of ExoMars 2020 mission: (a) Perspective view and (b) detailed view of the LaRa antennas placement.

Table 1.1: Brief specifications of LaRa antennas

Parameters		Values
Bandwidth		25 kHz
Carrier Frequency	TX	8.428 GHz
	RX	7.174 GHz
Copolar Gain	$20^\circ < \text{Elevation} < 30^\circ$	> 3 dB
	$30^\circ < \text{Elevation} < 55^\circ$	> 4 dB
Axial Ratio		< 2 dB
VSWR		< 1.5
Temperature		-125°C to $+60^\circ\text{C}$
Polarization		Right Circular
Mass		< 162 g (RX)
		< 132 g (TX)

1.1.1 Challenges Regarding Space Antennas

Apart from the challenge of adhering to the strict antenna pattern performance, the space antennas should also work optimally in the harsh environment. The specifications put numerous design constraints on the antennas, and a compromise between performance, cost and time is inevitable.

Table 1.1 shows some of the requirements put against the design of the LaRa antennas. What is immediately apparent is the high gain requirements at lower elevation. Furthermore, given the mass and envelope requirements, the size is no larger than twice the wavelength. The temperature requirement of -125°C to $+60^\circ\text{C}$ also imposes a huge constraint on the choice of the antenna technology and materials through thermoelastic considerations.

1.1.2 Evolution of LaRa Antennas

The initial specifications required the LaRa antennas to have a planar structure. Given that the planar technology exhibited a limitation regarding how much the patterns can be tilted to achieve sufficient gain in lower elevations, the planar technology requirement was relaxed. Moreover, the planar technology based on dielectric substrates would behave erratically at extreme

high and low temperatures without a specialized fabrication process to control the thermoelastic effect. Therefore, the design of LaRa antennas evolved from a planar slot array to a circular patch antenna, and, eventually, to a 3D monopole parasitic antenna. The all-metal 3D antenna, milled out of a single aluminum block, offered superior performance in terms of antenna patterns. Because of its 3D structure, the monopole parasitic antenna had a higher profile unlike the planar antennas. The additional height provided the much-needed higher gain at low elevation. The all-metal structure also provided resilience to extreme temperature variations, ensuring acceptable performance at extreme temperatures.

1.1.3 Novelty in LaRa Antenna Design

The LaRa antenna design evolved from a planar circular slot array [5] to the circular patch antenna [6]. Even though the antenna designs were adapted from Hubert [5] and Huang [6], the feed design was completely designed for these antennas.

The design of the all-metal 3D monopole antenna added another novel aspect to the design of the LaRa antennas. The proposed new design can generate circularly polarized conical patterns from a centrally-fed monopole antenna surrounded by the gamma-shaped parasitic elements.

1.2 Circularly Polarized Conical Beam Antennas

Circularly polarized (CP) conical beam antennas appeal to a certain class of applications. A mobile object on Earth wishing to communicate with a satellite on a given orbit is a perfect fit for the conical beam antennas, since such antennas cover the entire azimuth at given elevation range and do not require tracking of the satellite [7]. At lower frequencies (X-band and below), because of the Faraday rotation in the atmosphere, a change of polarization is unavoidable, and a CP antenna is preferred [8]. Moreover, a CP antenna helps negate the multipath effect [9], [7], and helps align the Earth-satellite frame of reference [10].

The design of such CP conical beam antennas started with planar antenna structures. Given that the antennas on substrates had sub-optimal

performance due to thermoelastic effect, the planar structure design had to be changed. This design change culminated in a novel design for a 3D monopole parasitic antenna, capable of radiating a CP conical beam.

1.2.1 Planar Antennas

In terms of planar structures, two antenna structures have been explored: a planar slot array antenna and a circular patch antenna.

Slot Array Antenna

Previous work at UCLouvain included a circular slot array for direction-finding applications [5]. The adaptation of the design proposed by Hubert and Craeye [5] required the design of a feed network integrated in the antenna so that a single port would effectively excite a CP conical beam pattern. There were eight slots in the design, and, thus, it required a 8-port feed network. Having eight-fold symmetry, the design performed quite well in terms of azimuthal symmetry. Another feature of this design was its limited back-radiation and isolation from platform since the antenna was entirely covered with the ground plane. However, the limitation of this planar structure was apparent in its inability to tilt the conical beam pattern to lower elevations.

Circular Patch Antenna

The circular patch antenna offered different possibilities to tilt the conical beam to lower elevation. One of the earliest works by Huang [6] showed that with proper excitation and choice of dielectric materials, it was possible for a circular patch antenna to achieve peak directivity from $\theta = 35^\circ$ to $\theta = 74^\circ$ based on different TM resonant modes. Two TM resonant modes were explored: TM_{21} and TM_{31} using a dielectric substrate ($\epsilon_r = 2.94$). The feed network of TM_{21} , however, proved to be tedious since the feed locations within the circular patch were offset by 45° , unlike TM_{31} for which the feed offset was 90° . Eventually, the TM_{31} circular patch antenna was designed, fabricated and tested.

1.2.2 Monopole Parasitic Antenna

The circular patch antenna, despite its long development, did not fully satisfy the high gain at low-elevation requirements. Additionally, the patch antenna had limited heritage to work in Martian conditions. This motivated the design of a novel antenna based on gamma-shaped parasitic elements surrounding a monopole feed. The antenna could be made entirely of metal, and so it could better withstand the Martian environmental conditions, whilst providing superior electromagnetic performance compared to the patch antenna. The monopole parasitic antenna design was, therefore, chosen for the LaRa instrument.

1.2.3 Use of Design Tools

For the design needs of the LaRa antennas, Microwave CST Studio Suite [11] was used. It not only offers different solutions to electromagnetic problems through its CST Microwave Studio module, but also provides solutions to multi-physics and electrostatic problems through CST MPhysics Studio and CST EM Studio modules, respectively. All these modules can be easily accessed and managed through the System Assembly and Modeling (SAM) tool available in the CST Design Studio.

The time domain (TD) solver or the transient solver of the Microwave CST is primarily based on the Finite Integration Technique (FIT) [12], which resolves Maxwell's equations (in differential form) in different grid space [13]. The TD solver can compute the broadband behavior of the antennas in one single run, and rapidly compute the frequency response with the Fourier Transform. A general overview of the numerical techniques and, in particular, the improvement in the near-field interactions by solving Maxwell's equations in integral form (based on the Method of Moments, MoM) is explored further in Section 1.3.

1.3 Numerical Methods in Electromagnetism

The present day computers are getting more and more powerful, and they can solve a huge number of problems in electromagnetism, which was oth-

erwise not possible some years ago. Yet, the real-world challenges tackled in electromagnetism continue to get bigger, and the efficient algorithms to better solve the Maxwell's equations are always evolving.

Numerous techniques currently exist to solve the problems in electromagnetism, depending on the types of problems. For problems with non-homogeneous media, using a differential equation approach, such as the Finite Difference Time Domain (FDTD) approach or the Finite Element Method (FEM) is appropriate. They require a volumetric mesh terminated by absorbing boundary conditions. For problems including radiation into the far field, such as antenna problems, integral equation methods, such as the Method of Moments formulation (MoM), are more appropriate. With MoM, however, the system of equations can become dense and large when the objects are large in terms of wavelength or comprise of fine geometrical details. The radiation and scattering problems involving large platforms are a typical example of large objects, where accelerated MoM techniques can be used. Objects with fine geometrical details come under the scope of a dense-mesh problem or near-field interactions in the order of less than a wavelength, where typically a classical but time-consuming formulation of the MoM is used.

Solution for the large objects exists since the early 1990s with the iterative techniques, such as Krylov-subspace methods, combined with fast matrix-vector multiplications, as those exploiting multipole decompositions (Fast Multipole Method or FMM) [14]. Problems with fine geometrical detail may be divided into subdomains, and solved with what is known as domain decomposition methods [15], [16], [17], such as the Macro Basis Functions (MBF) method. The interactions between the MBFs may still be time consuming, and hybrid methods such as MBF-FMM [18] improve this shortcoming to a certain extent. Nevertheless, in FMM, the density of plane waves over the unit sphere does not decrease with distance between subdomains, and, therefore, the compression rate of the MoM matrix remains the same [19]. With the Inhomogeneous Plane Waves (IPW) method [20], the number of plane waves required decreases with increasing distance, even if the patterns need to be recomputed for each pair of subdomains [19].

1.3.1 Numerical Techniques for Near Interactions

This thesis work continues the work on the Inhomogeneous Plane Waves, with a hybrid MBF-IPW approach [21] to solving radiation problems of antennas on a large platform. Typically, when the pairs of subdomains are larger than a wavelength, the method is efficient. For the cases when the subdomains are close, a classical MoM approach is used. In [22], local Taylor expansion of the Green's function is made once and for all on a regular grid over a region comparable to the wavelength. This produces a separable representation of the Green's function. Then, using precomputed moments of the elementary basis functions, a fast matrix-vector multiplication is obtained. However, given the singular behavior of the Green's function in the near field, the Taylor expansion converges slowly for very small distances and hence requires the introduction of a *dead zone* in which computations need to be done using a standard MoM impedance matrix. Given the fixed size of the dead zone, a complexity better than $\mathcal{O}(N^2)$ is difficult to achieve.

As a further development to solve the above shortcoming, when the near-field interactions need to be computed, a different approach is used by separating the free space Green's function into static and smooth parts [23]. The static part is treated with a multipole approach for the electrostatic potential [24]. When the separation is smaller than a wavelength, the variations in the smooth Green's function are very small and, thus, sampling the Green's function for each integration point is expensive [22]. A Taylor series expansion of the smooth Green's function yields high accuracy with the integrands that are separable, which lends itself easily to acceleration. Both the MBF-IPW approach and the static and smooth Green's function approach exploit the separability of the Green's function to improve the solution time of the MoM system of equations.

1.4 Goals of the Thesis

The thesis intends to basically tackle two important issues in antenna engineering: the design of a compact and robust antenna radiating a circularly polarized conical beam for a space application, and the numerical analysis of dense mesh problems devoted to antennas placed on a large platform.

For the design of robust space antenna, much work has been done from a novel antenna design concept to analyzing the multi-physics problems due to temperature variations and electrostatic effects. The design has evolved from a conical frustum array to planar structures and, finally, to an all-metal robust 3D antenna.

With regards to the numerical analysis of the effects of the platform on the antenna, two approaches have been considered: one using Inhomogeneous Plane Waves (IPW) and Macro Basis Functions (MBFs), and the other using a combination of static multipole algorithm adapted to near-field or dense mesh problems and Taylor series expansion. While the first method targets specifically the basis and test functions separated by distances larger than a wavelength, the second method targets basis and testing functions separated by much less than a wavelength, where the near-field distance has no relation anymore with the wavelength. Self interactions would still need to be computed using the brute-force implementation of the Method of Moments (MoM).

1.5 Outline of the Thesis

This thesis is organized into a larger design part and a smaller numerical part. The first part concerns the design of the antennas intended for the LaRa instrument. The numerical part expands on the methods intended for improving near-interactions between subdomains.

Chapter 1 introduces the motivations behind the thesis. This chapter essentially shows the two aspects of this thesis: one focused on the design and analysis of a robust antenna and the other one focused on the numerical analysis.

Chapter 2 discusses a novel antenna concept for beamforming based on a conical frustum array. A few facets can be selected sequentially to beamform the patterns over the entire azimuthal range. Depending on the applications, the elements in the frustum array can be either linearly or circularly polarized. Linearly polarized elements are useful for airborne applications, whereas the circularly polarized elements may be used for space applications or satellite communications.

Chapter 3 explores two planar antenna techniques to generate CP conical patterns. A design evolution from a circular slot array antenna to a circular patch antenna is explored in this chapter. The circular patch antenna was also fabricated and measured, and the test results are also briefly discussed.

Chapter 4 details the design and analysis of a novel all-metal 3D antenna with a monopole feed surrounded by parasitic elements. The flexibility of the antenna to meet various gain requirements is explored. Finally, a prototype made with additive manufacturing is experimentally validated.

Chapter 5 expands on Chapter 4 as applied to the antennas for the LaRa instrument. A further design evolution compatible with the space requirements, with multi-physics analyses, is explored. A novel feeding structure, square coaxial transmission line, is designed and analyzed. An Engineering Model (EM) is manufactured and tested, and the results are discussed.

Chapter 6 complements Chapter 2 as an evolution to a numerical analysis of the array using an Inhomogeneous Plane Wave method combined with Macro Basis Functions (IPW-MBF). The frustum array is divided into several subdomains, and the interactions between the subdomains are accelerated with the IPW-MBF technique.

Chapter 7 tries to overcome the limitations of the IPW-MBF discussed in Chapter 6, which is evident when the subdomains are separated by only one wavelength or less. The trick lies in the decomposition of the Green's function into static and smooth parts. These parts can be treated and accelerated separately to improve the computation speed.

Chapter 8 finally concludes the thesis, and provides a perspective on future work. Some advances can be made with regards to improving near-field interactions as well as improvements with the 3D monopole parasitic antenna to meet stringent design specifications.

Appendix A complements Chapter 3 and discusses the design philosophy and the assembly of the circular patch antenna.

CHAPTER 2

Beamforming with a Conformal Conical Frustum Array

This chapter showcases the work done in the development of a conformal conical frustum-type array. The radiation pattern should cover the entire azimuthal range at a given elevation angle. The elevation angle is strongly dependent on the tilt of the frustum. Two design applications were targeted - one for a Martian transponder for direct communication with Earth, and the other as a beamforming antenna aboard a flying platform.

Section 2.2 targets the design of a circularly polarized frustum array for the Martian transponder. This was a preliminary work carried out to explore the viability of having a beamforming antenna for the Martian transponder. The antenna would radiate at 45° elevation angle with full azimuthal coverage. This work culminated in a conference paper titled, “A conical frustum-type array devoted to a Mars-based transponder [25]”.

Realizing that the beamforming electronics would introduce additional complexity in the Martian transponder, where the antennas were already far from the transponder (see Fig. 1.2), the frustum array design was discarded for planar antennas, as detailed in Chapter 3. However, for applications where the antennas can remain close to the electronics circuitry as in nano-

satellites or in flying platforms such as drones and aircraft, the conformal frustum array retains its usefulness.

Section 2.3 expands on this airborne radar application with the focus on a simple beamforming architecture. This work has also been presented at a conference with the paper titled, “Sequentially shifted beamforming from a conformed conical array [26]”, and this has been done in collaboration with Jose Ignacio Echeveste and Thomas Pairon.

2.1 Introduction

Conformal antennas offer seamless integration capability. They find wide use in applications such as wireless sensor nodes for sensing toxic gases [27], energy harvesting [28], wearable watch [29] and textile on a space suit [30], ingestible capsule antenna [31], or over a conduit on a cylinder [32].

Applications are endless, yet, challenges persist, mainly in terms of controlling the antenna characteristics when morphed onto the structure. For applications where the shape of the structure can be maintained over time, the analysis and performance become controllable, and the shape can even be taken advantage of. For instance, it is easier to tilt the beam using planar arrays on tilted platforms, as the array need only to radiate at broadside. The tilt angle of the platform on which the planar arrays rest would then determine the angle in which the beam is tilted. Large beamforming structures can be used to transmit information from flying platforms, such as aircraft and drones to the ground, or to detect obstacles around those platforms. Being able to point beams in any azimuthal direction may require a relatively large number of antennas, distributed around the platform. Moreover, it may not be necessary to excite all the elements at the same time to form a beam; indeed, when the antennas are made conformal, their back radiation is very small and hence, contrary to the case of planar arrays, the patterns do not overlap. This pleads in favor of the use of a limited number of elements at a given time, at the expense of multiple beams. Beamforming with fewer elements of course allows the use of fewer active channels, assuming that they can be switched rapidly to any sub-array. Such classes of antennas are discussed in the following sections.

This chapter deals with two different designs, differing in the ability of the antenna elements to generate either circular or linear polarization. The circularly polarized frustum array in Section 2.2 is attractive for space or satellite communications, where circular polarization is a must and high gain a prerequisite. The tilted platform of the conformal array is useful since tracking satellites from ground always requires an optimal gain at certain elevation angles (for instance, elevation of 20° - 60° preferred at European latitudes for geostationary system [33]) with full azimuthal coverage. A crossed-slot antenna element produces the circular polarization, and its array implementation on a conical frustum is discussed. A particular attention is given to how many elements can contribute to beamforming at the same time.

Section 2.3 expands on the design of the previous section, and includes a beamforming architecture to show how a simple switching network can be employed to shift the different beams over the azimuth. A linearly polarized array is used here, since this is intended for target detection aboard a flying platform. Instead of a crossed-slot, a single slot element is used for generating linear polarization.

2.1.1 Circular Polarization and the Axial Ratio

Most of the antenna designs mentioned in this thesis work are circularly polarized (or rather elliptically polarized) and more often than not, it is usually preferred to express the quality of circular polarization in terms of axial ratio (AR).

The wave polarization is defined in terms of the locus traced by the time-varying electric field vector in a plane normal to the direction of propagation in space. The locus can be represented in terms of a polarization ellipse as shown in Fig. 2.1. Assuming the direction of propagation in $\hat{\mathbf{z}}$ direction, the instantaneous electric field vector (at $z = 0$) can be decomposed into orthogonal terms as [34, Section 2.8]

$$\mathcal{E}(z = 0; t) = \mathcal{E}_x \hat{\mathbf{x}} + \mathcal{E}_y \hat{\mathbf{y}} = E_x \cos(\omega t) \hat{\mathbf{x}} + E_y \cos(\omega t + \delta) \hat{\mathbf{y}} \quad (2.1)$$

where δ gives the phase difference between the x - and y -components.

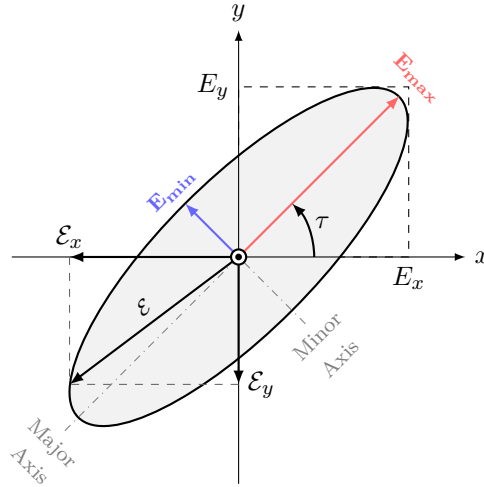


Figure 2.1: The electric-field vector resolved into linear components on a polarization ellipse.

In phasor form, (2.1) becomes

$$\mathbf{E} = E_x \hat{\mathbf{x}} + E_y e^{j\delta} \hat{\mathbf{y}} \quad (2.2)$$

Depending on the shape of the ellipse, the polarization may be linear, circular or elliptical.

Linear The polarization is linear when there is no minor axis of the ellipse ($\delta = n\pi$; $n = 0, 1, 2, \dots$ and tilt angle, $\tau = 0^\circ$ or 90°), and so the ellipse becomes a straight line.

Circular If the major and minor axes are equal ($E_x = E_y$ and $\delta = \pm(\pi/2 + 2n\pi)$; $n = 0, 1, 2, \dots$), the polarization is circular.

Elliptical If the major and minor axes are *not* equal ($E_x \neq E_y$ and $\delta \neq n\pi$ or $\delta \neq \pm(\pi/2 + 2n\pi)$; $n = 0, 1, 2, \dots$), the polarization is elliptical.

AR in terms of Two Orthogonal Linearly Polarized Components

The ratio of the major axis to the minor axis of the polarization ellipse gives the AR [35, Section 2.12], i.e.

$$\begin{aligned}
 \text{AR} &= \frac{\text{Major Axis}}{\text{Minor Axis}} = \frac{|\mathbf{E}_{\max}|}{|\mathbf{E}_{\min}|} \\
 &= \frac{\sqrt{\frac{1}{2} [E_x^2 + E_y^2 + \sqrt{E_x^4 + E_y^4 + 2E_x^2 E_y^2 \cos(2\delta)}]}}{\sqrt{\frac{1}{2} [E_x^2 + E_y^2 - \sqrt{E_x^4 + E_y^4 + 2E_x^2 E_y^2 \cos(2\delta)}]}} \\
 &= \sqrt{\frac{E_x^2 + E_y^2 + \sqrt{E_x^4 + E_y^4 + 2E_x^2 E_y^2 \cos(2\delta)}}{E_x^2 + E_y^2 - \sqrt{E_x^4 + E_y^4 + 2E_x^2 E_y^2 \cos(2\delta)}}} \quad (2.3)
 \end{aligned}$$

For circular polarization, where $\delta = \pi/2$, (2.3) simplifies to

$$\text{AR} = \frac{E_x}{E_y} \quad (2.4)$$

AR in terms of Two Orthogonal Circularly Polarized Components

The axial ratio may also be expressed in terms of the circularly polarized components, the left- and right-hand circularly polarized components. The left-hand circular polarization (LHCP) and the right-hand circular polarization (RHCP) give the *sense* of circular polarization. If the electric field rotates counterclockwise in time or $\delta = +(\pi/2 + 2n\pi)$ in (2.2), the circular polarization is LHCP. And, if the electric field rotates clockwise in time or $\delta = -(\pi/2 + 2n\pi)$ in (2.2), the circular polarization is RHCP. The RHCP follows the direction of a right-handed screw.

The electric field vector decomposed in LHCP and RHCP components is given as

$$\mathbf{E} = E_R(\hat{\mathbf{x}} + j\hat{\mathbf{y}}) + E_L(\hat{\mathbf{x}} - j\hat{\mathbf{y}}) \quad (2.5)$$

$$= \hat{\mathbf{x}}(E_R + E_L) + \hat{\mathbf{y}}j(E_R - E_L) \quad (2.6)$$

This implies,

$$E_x = E_R + E_L \quad (2.7)$$

$$E_y = j(E_R - E_L) \quad (2.8)$$

$$E_R = (E_x - jE_y)/2 \quad (2.9)$$

$$E_L = (E_x + jE_y)/2 \quad (2.10)$$

The AR then is given by [36, Chapter 3]

$$\text{AR} = \left| \frac{E_R + E_L}{E_R - E_L} \right| \quad (2.11)$$

AR is often expressed in decibels using the following relation

$$\text{AR}_{[\text{dB}]} = 20 \log \left| \frac{E_R + E_L}{E_R - E_L} \right| \quad (2.12)$$

The orthogonal polarizations may also be expressed as *co-polarization* and *cross-polarization*. Co-polarization or copol represents the intended polarization of the antenna, while cross-polarization or crosspol represents the polarization orthogonal to the copol. In this thesis, the polarizations are expressed in terms of both right- and left- polarizations as well as co- and cross-polarizations. If the antenna is designed to radiate a RHCP beam, the copol represents the RHCP beam. It is, therefore, useful to relate AR with the level of cross-polarization or *crosspolar discrimination*, XPD (see also Fig. 2.2)

$$\text{XPD} = 20 \log \left| \frac{\text{AR} + 1}{\text{AR} - 1} \right| \quad (2.13)$$

2.2 Circularly Polarized Conical Frustum Array

This section details the design of the crossed-slot element with a single feed, and the performance of the frustum array, when a few facets are selected. This section also explores the feed network for the frustum array, and some perspectives on the integration of the frustum array.

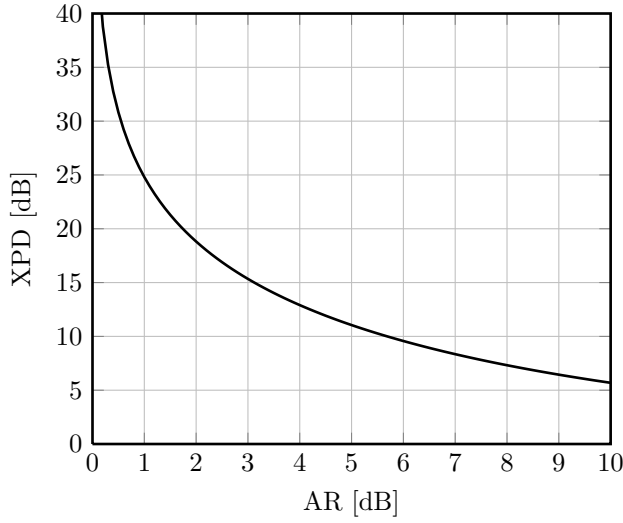


Figure 2.2: Axial ratio and cross polar discrimination.

2.2.1 Crossed-Slot Single Feed Element

The crossed-slot element (Fig. 2.3) is capable of generating circular polarization. A circularly polarized antenna may be designed by exciting two orthogonal linearly polarized antennas with a 90° phase shift. Such antenna can be aperture-coupled using a microstrip or a stripline with a crossed slot antenna [37]. However, having an off-center feed to excite orthogonally the two arms of a crossed slot antenna imposes an asymmetry in the antenna, which lowers the cross-polarization level [38]. On the other hand, feeding the crossed slot in a balanced manner reduces the induced voltage across the unfed slot, thereby decoupling the slot entirely. Ideally, a 90° hybrid network would work, however, for a compact frustum antenna, a simple single-line feed with the effect of an hybrid coupling [39] is deemed more appropriate. Katayama [40] reported an inclined crossed slot antenna fed with a microstrip line. The structure, being inherently asymmetric, had the tendency to tilt the beam and shift the axial ratio. Iwasaki [41] mentioned a crossed-slot etched on a circular patch antenna and proximity fed with microstrip, but it suffered from a poor axial ratio bandwidth. Designing a circularly polarized antenna presents not only a constraint on the impedance

bandwidth but also on the axial ratio bandwidth. The axial ratio bandwidth is also the most limiting factor in the circularly polarized antenna design. Several designs in [38], [42] include the use of a patch antenna on top of the crossed slot to increase the bandwidth of the antenna.

A single-feed circularly polarized crossed-slot antenna is designed at X-band using a stripline feed. The antenna structure is shown in Fig. 2.3. The antenna consists of a crossed-slot etched on top of a 0.254 mm thick Rogers RT6006 substrate ($\epsilon_r = 6.15$, $\tan\delta = 0.0023$) on a slab of dimension 20 mm $\times$ 20 mm. A stripline runs through the antenna sandwiched in the substrate between the top crossed-slot antenna and the bottom ground plane shown in Fig. 2.3. With a ground plane below, the back radiation is limited [43]. The stripline is fed from one of the edges of the slab with a waveguide port. The slot elements are sequentially excited by the stripline so that the phase difference between any two consecutive slot apertures is maintained close to 90° with a given quarter-wavelength stripline length. The stripline is terminated with a series stub to effectively couple with the final slot aperture and, to some extent, match the impedance seen at the excitation port. While the initial design had the same slot length and width for the crossed slot, different slot lengths and widths were found optimal. Having different slot lengths effectively changes the resonance of the antenna [44]. With the proper dimensions, the resonant frequency can be optimized.

For the slot lengths of 8.63 mm and 8.34 mm, widths of 0.83 mm and 0.66 mm, and feed width of 0.148 mm, the plots of axial ratio (AR) and reflection coefficient as a function of frequency are shown in Fig. 2.4. The broadside AR ($\theta = 0^\circ$) is below 3 dB from 8.319 GHz to 8.421 GHz. The reflection coefficient is below -10 dB from 8.3 GHz to 8.45 GHz. The resonance could be made even better with different matching networks, such as double stub matching network and quarter-wave transformers, to match the stripline feed to a given characteristic impedance. If two similar crossed-slot antennas are used in parallel, the matching network needs to match to a 100 Ω transmission line so that when both are combined, the line will be matched to a 50 Ω line.

Also, the AR observed for $\phi = 0^\circ$ at the resonant frequency of 8.365 GHz is shown in Fig. 2.5. The beam spans roughly from $\pm 20^\circ$. The asymmetry

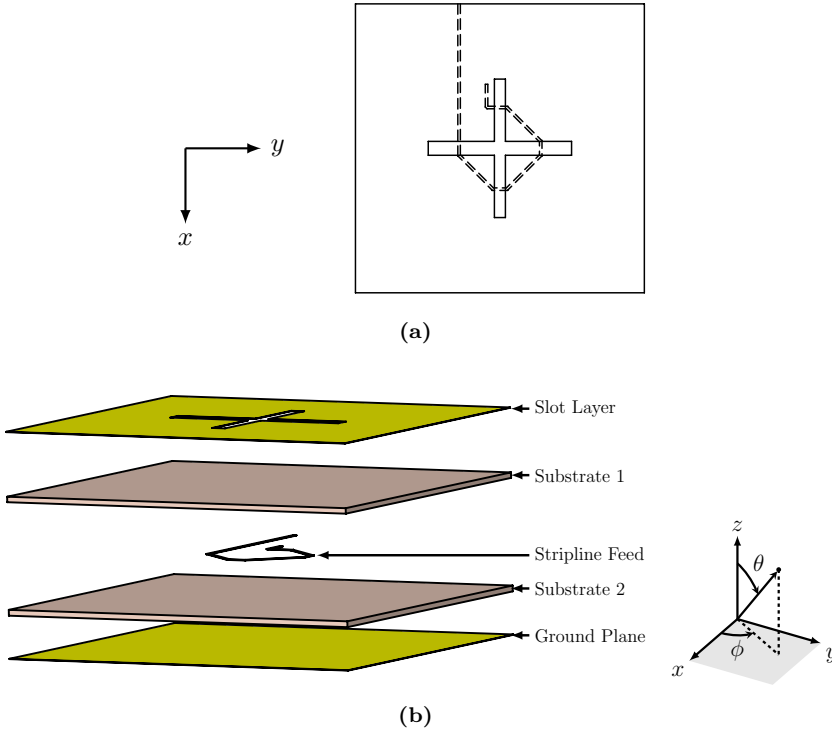


Figure 2.3: Crossed-slot antenna structure: (a) Top view with the slot and the stripline feed (b) Exploded view.

can be attributed to the stripline feed, which is slightly offset from the first slot arm.

2.2.2 Crossed-Slot Frustum Array

The crossed slot conical frustum antenna array is shown in Fig. 2.6, and bears a similar structure to the GAIA satellite phased array antenna [45] but differs in the type of antenna elements and beamforming architecture. The proposed frustum array comprises two crossed-slot elements in each of the 12 facets of the conical frustum. The composition of the array is similar to that of the single feed crossed slot element outlined in Section 2.2.1. The frustum is a compact structure, spanning a radius of 6 cm with a height of 3 cm. When placed in a platform, the embedded element patterns of all

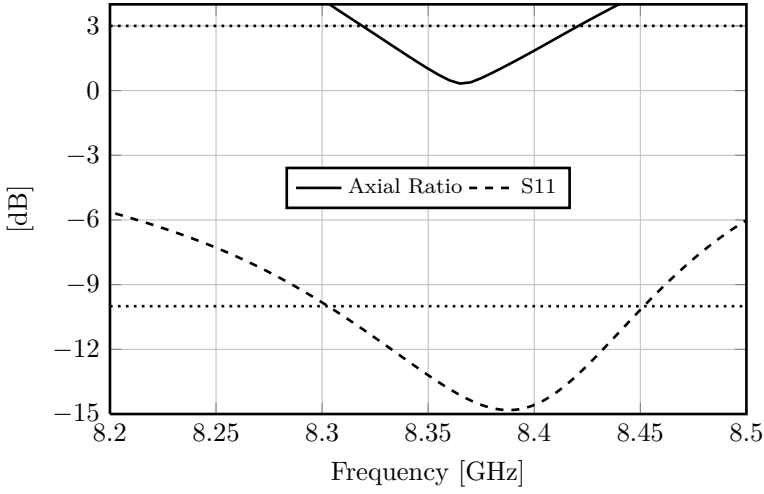


Figure 2.4: Simulated AR plot versus frequency ($\theta = 0^\circ, \phi = 0^\circ$) and reflection coefficient plot for single feed crossed-slot antenna.

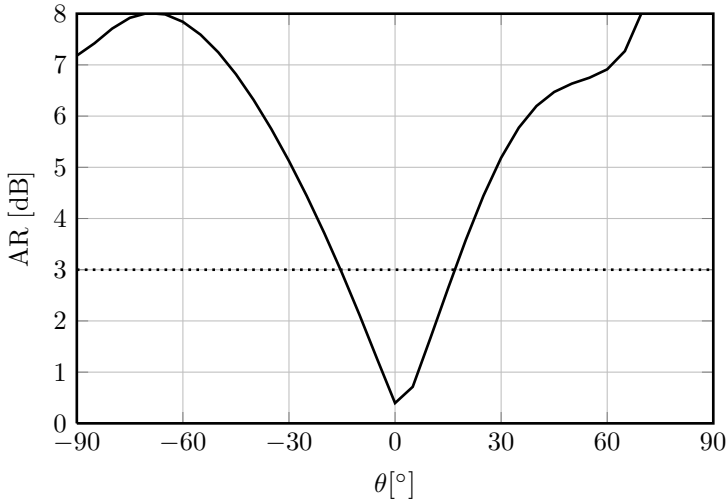


Figure 2.5: AR of single feed crossed-slot antenna for $\phi = 0$ at 8.365 GHz.

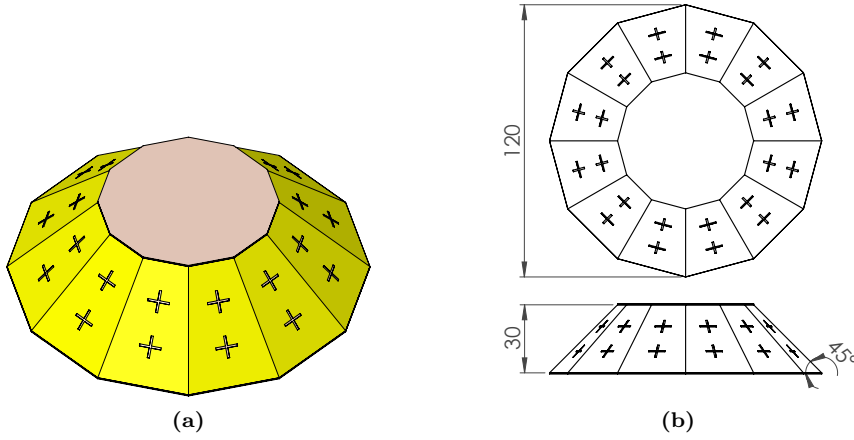


Figure 2.6: Crossed-slot conical frustum antenna array: (a) perspective view with two crossed-slot elements in each of 12 facets (b) Dimensions (in mm) of the conical frustum array.

the elements are computed, and later combined with appropriate amplitude and phase to form the desired beam while at the same time maintaining circular polarization.

For analysis, each element of the array is fed using discrete delta-gap sources, four sources for four slot arms in each crossed-slot antenna, all at 90° phase difference to induce circular polarization. Ultimately, the discrete sources can be replaced by the stripline feed, as described in Section 2.2.1 with around 12 channels to select each one of the facets. Also, it is sufficient to select only 3 facets at a time for azimuthal beam scanning given the large beamwidth. Since, for 12 facets, each facet should cover a beamwidth of $360^\circ/12 = 30^\circ$, while for 2 and 4 facets, the angular coverage should be at least 15° and 7.5° , respectively, for full azimuthal coverage. A sequentially shifted beamforming architecture is described in detail in Section 2.3.

The plot of AR over frequency when 6 crossed-slot elements in 3 facets are selected is shown in Fig. 2.7. The AR at $\theta = -45^\circ$ and $\phi = 0^\circ$ is below 3 dB from 8 GHz to 8.16 GHz. The AR bandwidth in this case is limited but can be easily optimized with stripline feeds.

The global array radiation pattern, when 6 crossed-slot elements are

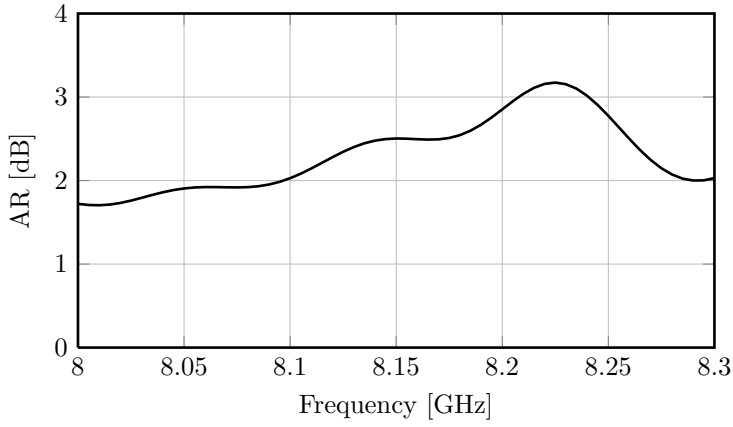


Figure 2.7: Simulated AR plot versus frequency of 6 crossed-slot elements in 3 consecutive facets of the conical frustum antenna array.

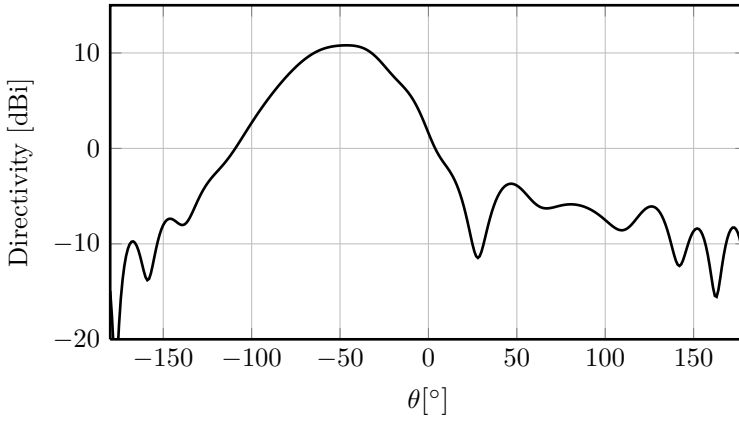


Figure 2.8: Radiation pattern of 6 crossed-slot elements in 3 consecutive facets of the conical frustum antenna array at 8.1 GHz.

excited, is shown in Fig. 2.8. The crossed-slot elements are also phase-shifted to account for their separations. Also, note that the maximum radiation is at -45° , with a directivity of 11 dBi. Depending on the applications, the angle of maximum gain can be adjusted from the tilt of the frustum.

2.3 Sequentially Shifted Beamforming

The frustum-type array in this section consists of linearly polarized slot elements. Each facet comprises two elements to obtain a more directive beam in elevation. Then several facets are beamformed to implement azimuthal scanning. The arrangement of antennas over small facets and the proximity of the whole conical structure strongly distort the antenna patterns, through coupling with the neighbors and with the platform. In order to fully include those effects, the beamforming analysis will start from the embedded element patterns [46] obtained for a realistic structure with linearly polarized antenna elements and frustum platform comprising 12 facets. This section explains how the shape of the array pattern evolves while it is scanned over a sectoral domain, for which a given set of consecutive antennas is used, and the implementation of a simple sequential beamforming architecture.

2.3.1 Antenna Design

The antenna structure has 12 facets, and each facet consists of two linearly polarized slot elements, as shown in Fig. 2.9. The antenna array is designed to operate at 24.125 GHz, the frequency of choice for radiolocation. The slot elements are 0.5λ in length and are separated by 0.6λ in the facet.

The slot elements are etched on top of a 0.508 mm Rogers RO4003C laminate. The substrate is backed by a ground plane to reduce back radiation from the slot elements. The simulations are carried out using discrete ports and, thus, each slot element is fed directly with a delta-gap source. A feed network can be realized with striplines and the beamforming network can be installed inside the frustum in the bottom.

The facets are slanted by 45° from the base of the frustum. The facets can be fabricated individually and screwed on top of a frustum holder. The number of facets determines the appropriate beam coverage and the required number of consecutive facets that need to be selected for azimuthal beam-scanning. Indeed, beamforming between consecutive facets requires a sufficient overlap between the respective embedded element patterns.

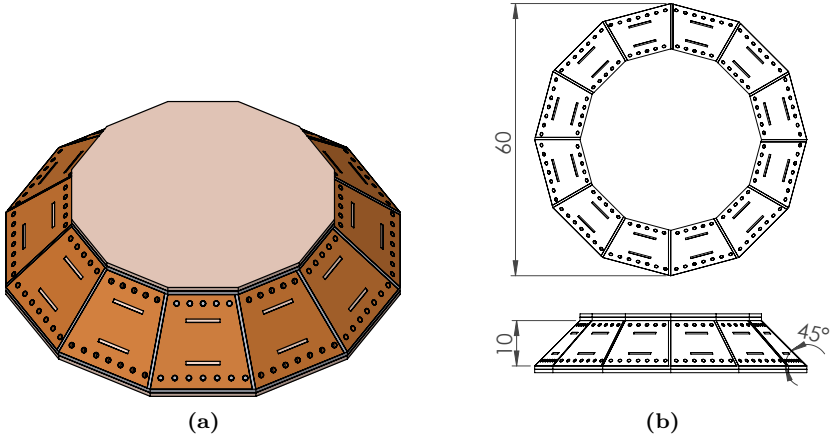


Figure 2.9: A 12-facet frustum array with linearly polarized slot elements: (a) perspective view with two slot elements in each of 12 facets (b) Dimensions (in mm) of the conical frustum array.

2.3.2 Embedded Element Patterns

The embedded element pattern of each facet is computed using the commercial software CST Microwave Studio. Fig. 2.10 shows the embedded element patterns (color scale in dB, horizontal coordinates in the (u, v) plane; $u = \sin \theta \cos \phi$, $v = \sin \theta \sin \phi$) of the two consecutive facets shown in Fig. 2.9. The embedded element patterns are combined in groups of three facets to steer the beam. The frustum is designed with overlapped subarrays [47] where two consecutive sub-arrays share two facets.

Each combination of three facets steers the beam for an angular range of 30° . Three different steered beams are obtained for each facet using five different delay phase lines (see Section 2.3.3). These angles are chosen to be $-10^\circ, 0^\circ$ and 10° , obtaining a beam every 10° in azimuth. This allows the generation of 36 equispread beams over the whole azimuthal range. In Fig. 2.11, the three steered beams towards $-10^\circ, 0^\circ$ and 10° are represented for a 44° elevation. A maximum variation of 0.26 dB is obtained among the three beams of peak gain. Fig. 2.12 shows superimposed patterns after 10° shifting.

The 3-dB beamwidth is 24° and the side lobe level is 7 dB lower than

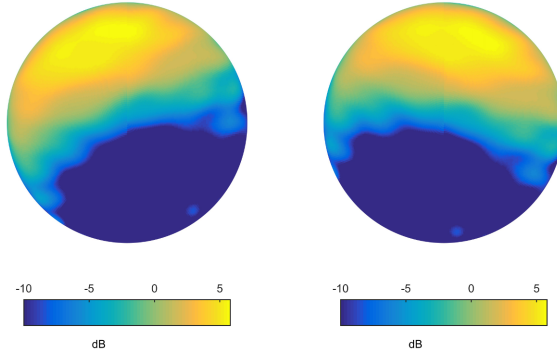


Figure 2.10: Embedded element patterns of two consecutive facets.

the main beam. The main beam is not greatly affected by the steering, the side lobe level is increased by about 1.5 dB with respect to the broadside.

2.3.3 Beamforming Architecture

This section describes the beamforming architecture developed in order to steer the beam over 360° azimuth range. This work has been carried out in close cooperation with Thomas Pairon (UCLouvain). As discussed in Section 2.3.2, for each group of 3 facets, the beam is steered towards 3 discrete angles $\psi_i + \phi$ with $\psi_i \in [-10, 0, 10]^\circ$ and ϕ corresponding to the angular position of the second facet of the given group. Owing to the large beam width, it is sufficient to perform beamforming in those three directions in order to cover the whole azimuth with overlapping patterns (see Fig. 2.11).

In order to steer the beam in direction ψ , the proper group of consecutive antennas $[n-1, n, n+1]$ is selected using Table 2.1. The n^{th} antenna is phase-shifted by an angle α_n expressed as:

$$\begin{aligned}\alpha_{n-1,i} &= -ka \cos \theta \cos (2\pi/N + \psi_i) \\ \alpha_{n,i} &= -ka \cos \theta \cos (\psi_i) \\ \alpha_{n+1,i} &= -ka \cos \theta \cos (2\pi/N - \psi_i),\end{aligned}$$

where k is the wavenumber, a is the radius of the frustum, θ is the elevation angle (in this configuration, $\theta = 44^\circ$) and N is the number of facets.

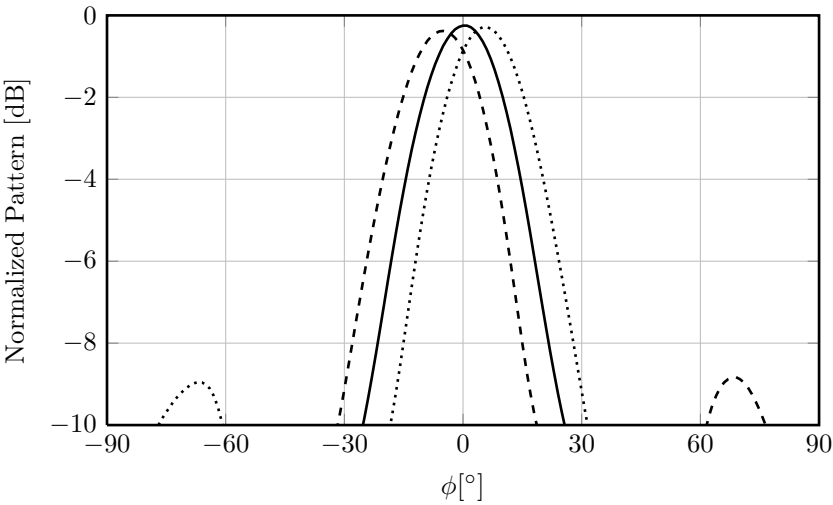


Figure 2.11: Steered beams with different phase combination for a given set of facets.

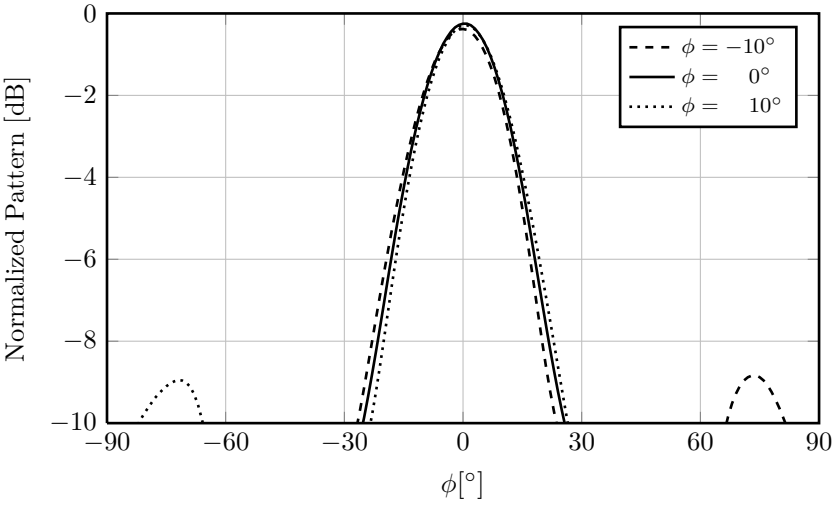


Figure 2.12: Superimposed beams created within a sector.

Table 2.1: State of the control signals S_p for a desired angular range.

ψ [°]			Antenna Facet			S_0	S_1	S_2
5	→	25	12	1	2	00	00	11
35	→	55	1	2	3	00	00	00
65	→	85	2	3	4	01	00	00
95	→	115	3	4	5	01	01	00
125	→	145	4	5	6	01	01	01
155	→	175	5	6	7	10	01	01
185	→	205	6	7	8	10	10	01
215	→	235	7	8	9	10	10	10
245	→	265	8	9	10	11	10	10
275	→	295	9	10	11	11	11	10
305	→	325	10	11	12	11	11	11
335	→	355	11	12	1	00	11	11

Since two consecutive groups have two overlapping elements (i.e. each element belongs to three different groups, see Table 2.1), 9 different values of α exist for each element. Among those 9 values, only 5 values are unique as $\psi_0 = -\psi_2 = 10^\circ$ and $\psi_1 = 0^\circ$. It is possible to steer the beam over 360° by sequentially selecting 3 adjacent elements; those received signals are phase-shifted with an angle among a set of 5 possible values only. The time-delayed signals are then combined.

An analog beamformer architecture is selected in favour of a digital implementation since the number of required time delays and the number of simultaneously active antennas is small [48]. In order to activate any subset of 3 consecutive elements among $N = 12$ elements, only 3 SP4T (Single-Pole Four-Throw) RF switches are required (see Fig. 2.13). The activated path of the p^{th} switch is selected by means of a 2-bit control signal S_p . Table 2.1 summarizes the state of the control signals S_p for a given range of steering angles.

Once the antennas are selected, the proper phase shift is applied to each output of the RF switches in order to steer the beam. These phase-

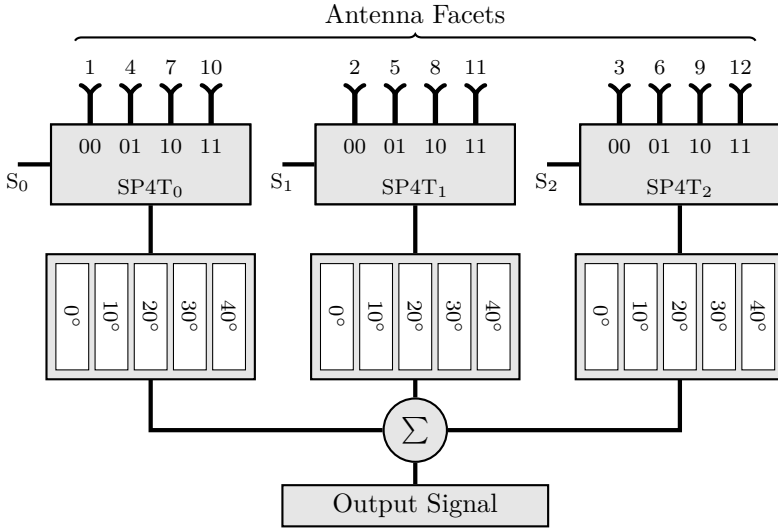


Figure 2.13: Design architecture of the analog beamforming using true-time delay transmission lines as phase shifters.

shifters are implemented by a matrix of 5 electronic true time-delay (TTD) lines. The TTD lines ensure wideband operation, given that phase shifts immediately correspond to required time shifts [49]. These TTD are either microstrip lines or coaxial cables of different electrical lengths. For each switch output, the proper TTD can be selected using RF switches (SP5T) and 3-bit control signals (not represented on Fig. 2.13).

Finally, the steered beam is obtained from a combination of the three signals with a 3-way power combiner before sampling with a single channel analog-to-digital converter (ADC). The full digital beamforming approach [50] can not be overlooked, but in regards to the small number of required phase shifts allowing a full azimuthal scanning, an analog approach has been selected. For direction-finding purposes, outputs obtained from consecutive beams can be deconvolved to achieve a resolution much finer than 10° .

2.3.4 Feed Design for the Frustum Array

The frustum array, be it with circularly or linearly polarized elements, can be electromagnetically coupled with a stripline underneath the elements. The feed network can be built as a power divider circuit in stripline, and fed from the bottom with an SMA connector. Semi-rigid cables can then connect the SMA connectors of each facet to the RF switches at the bottom of the frustum.

2.4 Conclusion

A conical frustum array with 12 facets is proposed targeting two distinct applications differing in polarization. The circularly polarized frustum array finds applications in space communications. A crossed-slot element, fed with stripline and imitating a hybrid feed network is also discussed. The linearly polarized array can be used for tracking objects from flying platforms. A slot element, electromagnetically fed with a stripline, generates the linear polarization.

A simple analog beamforming architecture was introduced using only a few RF switches. The phase-shifters are implemented using true time-delay lines. Five different phase-shifters combined with the RF switches enable scanning over azimuth with a 10° resolution, or better if some post-processing is applied.

This chapter dealt with the analysis of antenna arrays conformed to a conical structure, with a primary aim of using such antenna design for the LaRa transponder. However, the complication of using the RF circuitry in the antennas placed already far away from the transponder without appropriate thermal control in the Martian environment could not be overlooked. Therefore, a decision to shift the design from the conformal array to a planar antenna design was made, and such planar antenna design and test is further discussed in Chapter 3.

CHAPTER 3

Circularly Polarized Conical Beam Planar Antennas

This chapter explores planar antenna technology to generate a circularly polarized (CP) conical beam. Slot and circular patch antenna designs with their stripline feeding networks are also discussed. A CP conical beam antenna is required for a Martian transponder (see Section 1.1). The circular patch antenna was also fabricated and tested. The feed network design has been done in collaboration with Jose Ignacio Echeveste.

Section 3.3 deals with the design of the circular slot array and its integrated feed network. The slot elements are all linearly polarized, yet, given the rotating phase in the feeding network, the array pattern polarization is circular. This observation has been done initially with Simon Hubert.

Section 3.4 describes the design of the circular patch antenna resonating in different modes. The most appropriate design was chosen, fabricated and tested. Also discussed is the appropriateness of such technology for space application, where wide temperature variations can drastically affect antenna performance. Then, the motivation to shift towards an all-metallic design is reiterated.

3.1 Introduction

Circularly polarized antennas with conical patterns are widely studied in the literature. Such antennas find wide applications in communications between mobile devices and satellites because the conical patterns have uniform elevation pattern over the azimuthal plane [51]. Conical patterns are useful and cost-effective for ground-based communication devices utilizing satellite services. At European latitudes for a geostationary system, the antenna mostly needs to cover an elevation range of 20° to 60° (or $30^\circ < \theta < 70^\circ$ from zenith) [33]. This avoids the need for complex beamsteering in mounted devices while providing full azimuthal coverage. Such application also requires good axial ratio ($AR < 3$ dB) because of the change in polarization in the Earth's atmosphere due to Faraday rotation, especially at frequencies below X-band [8]. The conical beam antennas also find applications in indoor wireless communications [52], [53]. Thanks to the azimuthal symmetry, the conical-beam antennas provide wide coverage.

Circular polarization in an array is generated by feeding the linearly polarized elements with appropriate phase shifts [54]. A conical beam in circular patch antennas can be generated by exciting second- or higher-order modes [6]. Both designs focus on a circular patch antenna or on a circular array of patch antennas. A simple single-feed design with notches in a circular patch antenna is outlined by Nakano *et al.* [55], but it has poor axial ratio beamwidth and asymmetric pattern. A four-port feed network utilizing hybrids was reported by Fouany *et al.* [56] for an isoflux antenna.

3.2 Choice of the Dielectric Substrate

The planar antennas described in this chapter are etched on a dielectric substrate whose choice is critical for space applications. The space qualification is usually done on the finished hardware, but it is equally desirable to properly select the substrate to begin with.

Among the most important characteristics of a dielectric substrate for a space application are its outgassing and thermal properties in addition to the electromagnetic properties. Table 3.1 lists those important properties.

Table 3.1: Properties of different dielectric substrates considered.

Property	RO3003	RO4003C	RT5880	RT5880LZ	RT6002	TMM10i
¹ ϵ_r	3.00	3.55	2.20	1.96	2.94	9.9
¹ $\tan\delta$	0.0013	0.0027	0.0009	0.0019	0.0012	0.0020
² TC (ϵ_r)	+13	+40	-125	+22	+12	-43
³ CTE	17	11	31	44	16	19
	16	14	48	33	16	19
	25	46	237	41.5	24	20
⁴ TML	-	0.06	0.03	0.01	0.02	0.06
⁵ CVCM	-	0.00	0.00	0.01	0.01	0.00
⁶ WVR	-	0.02	0.02	0.02	0.01	0.04

¹@10 GHz²Thermal Coefficient of dielectric constant from -50°C to 150°C , in ppm/ $^\circ\text{C}$ ³Coefficient of Thermal Expansion (X,Y,Z) from 0°C to 100°C , in ppm/ $^\circ\text{C}$ ⁴Total Mass Loss, in %. Recommended $< 1\%$ ⁵Collected Volatile Condensable Materials, in %. Recommended $< 0.1\%$ ⁶Water Vapor Recovered, in %. Recommended $< 0.1\%$

The recommended values of the outgassing properties (TML, CVCM and WVR) are provided by the space agencies, and all the substrates mentioned in Table 3.1 meet these criteria (TML $< 1\%$, CVCM and WVR $< 0.1\%$).

Property 2 (TC) indicates the stability of the dielectric constant over a range of temperature, and property 3 (CTE) shows the mechanical deformation in X,Y and Z directions as temperature changes. These thermal properties affect the electromagnetic behavior of the antenna, since one (TC) changes the dielectric constant while the other (CTE) physically changes the dimension of the antenna as temperature varies. More critical perhaps is the CTE in Z if the integrity of the vias in the design has to be ensured.

As will become clear in the subsequent sections, the proposed planar antenna design uses several vias to connect the multi-layer PCB, and thus the RT6002 dielectric substrate has been found the most suitable for the design.

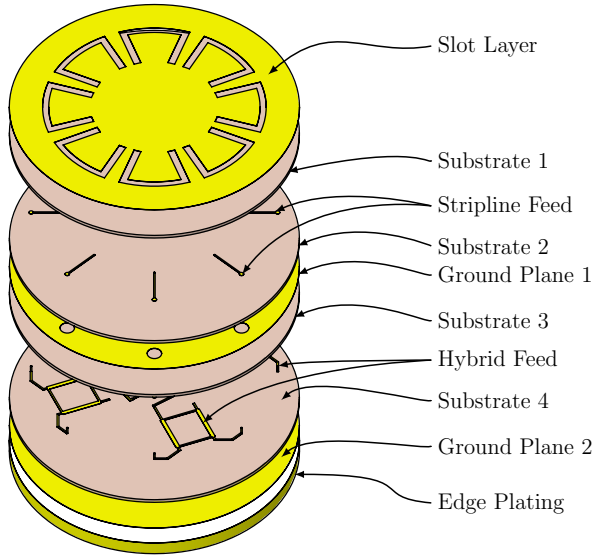


Figure 3.1: Circular slot array layer stackup exploded view.

3.3 Circular Slot Array Antenna

This section describes a single-feed circular slot array of 8 elements producing a circularly polarized conical beam. The circular slot array for a direction finding application was reported by Hubert and Craeye [5]. However, that design does not have a single feed structure nor a circularly polarized conical beam. This section proposes a single-feed design connected to a feed network to generate circularly polarized conical beam.

3.3.1 Design of Slot Array Antenna

The proposed circular slot antenna array operates with an angular spacing of 45° and a phase separation of 90° . The exploded view of the different layers of the circular slot array is shown in Fig. 3.1. The top slot layer consists of 8 U-shaped slot elements etched on top of a Rogers RT6002 substrate ($\epsilon_r = 2.94$; $\tan \delta = 0.0012$) [57]. The slot elements are electromagnetically coupled with stripline, which are then connected through vias to a feed network located underneath, as shown in Fig. 3.2(a). The feed network,

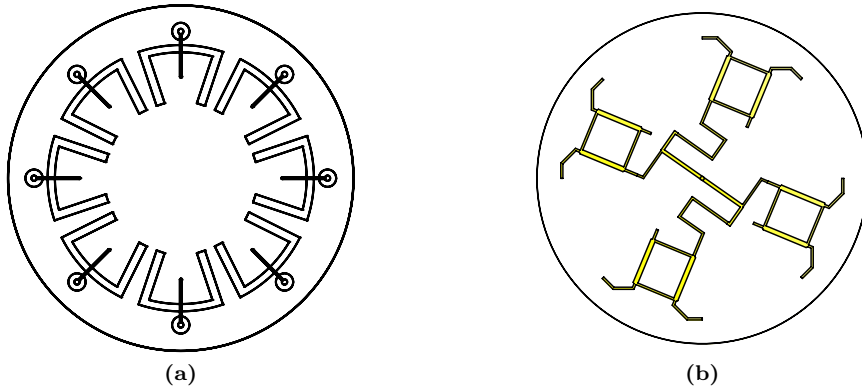


Figure 3.2: Circular slot array top views: (a) Slot elements with coupled stripline feed and (b) Hybrid feed network.

shown in Fig. 3.2(b), distributes equal power with a 90° phase difference between consecutive elements. It consists of 90° hybrids, T-junction power dividers, quarter wave transformers and delay lines to ensure appropriate excitation at each slot element. The feed network is also employed to match the impedance of the SMA connector (50Ω) to a higher impedance in the stripline that couples to the slot elements. This five-copper layer design is built on four identical substrates, each of thickness 0.508 mm (see Fig. 3.1).

The choice of the number of elements is made in order to obtain a uniform pattern over the azimuthal range (ϕ) [58]. The opposite elements should add up in phase to avoid cancellation of patterns in that particular direction. The slots are designed to resonate at 8.415 GHz, each of length $0.56 \lambda_0$, where λ_0 is the free-space wavelength. They are bent inside the array so as to accommodate all the elements in the circular array. The slot termination length and the stripline feed stub are used to tune the resonance of the slot.

3.3.2 Active Impedance of the Circular Slot Array

The resonance of the slot array is strongly dependent on the phase variations of the slot array elements, since the elements are sequentially excited with a phase difference of 90° . This requires a derivation of the impedance in

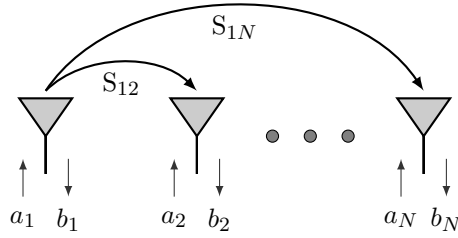


Figure 3.3: Array antenna of N elements.

terms of active impedance, as opposed to passive impedance.

Consider an array of N elements, with incident a and reflected b voltages as shown in Fig. 3.3.

The reflected voltage from element i accounting for mutual coupling is

$$b_i^A = S_{i1}a_1 + \cdots + S_{1i}a_i + \cdots + S_{iN}a_N \quad (3.1)$$

The active impedance of element i , S_{ii}^A , is the ratio between the incident and reflected voltages when the array is fully excited, i.e

$$S_{ii}^A = \frac{b_i^A}{a_i} = \frac{\sum_{n=1}^N S_{in}a_n}{a_i} = \frac{\sum_{n=1}^N S_{in}A_n e^{j\phi_n}}{A_i e^{j\phi_i}} \quad (3.2)$$

where the incoming voltages are the excitations of the array defined by $a_i = A_i e^{j\phi_i}$.

In the circular slot array design, the elements are uniformly excited in amplitude and with a phase difference of 90° between elements. Equation (3.2) becomes

$$S_{ii}^A = \frac{\sum_{n=1}^N S_{in} e^{j\phi_n}}{e^{j\phi_i}} = \frac{\sum_{n=1}^N S_{in} e^{j(n-1)\Delta\phi}}{e^{j\phi_i}} \quad (3.3)$$

where $\phi = [0^\circ, 90^\circ, 180^\circ, 270^\circ, 0^\circ, 90^\circ, 180^\circ, 270^\circ]$.

The ideal antennas have zero reflection voltages at their resonant frequency. In order to account for the interaction between elements, the re-

reflection voltages of the fully excited ($b_i^A = 0$) array should be minimized

$$b_i^A = S_{ii}^A = 0 \quad (3.4)$$

Expressed in a matrix form and with the proper excitation

$$\begin{bmatrix} b_1^A \\ b_2^A \\ \vdots \\ b_8^A \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & \cdots & S_{18} \\ S_{21} & S_{22} & \cdots & S_{28} \\ \vdots & \vdots & \ddots & \vdots \\ S_{81} & S_{82} & \cdots & S_{88} \end{bmatrix} \begin{bmatrix} 1 \\ e^{j90} \\ \vdots \\ e^{j270} \end{bmatrix} \quad (3.5)$$

Reciprocal networks satisfy $S_{ij} = S_{ji}$. Furthermore, in the case of antennas that have discrete rotational symmetry, as it is the case of the present antenna, the S parameters are $S_{ij} = S_{(i+1;j+1)}$. The excitation vector applied to the antenna has a phase difference of 90° between consecutive elements. Equation (3.5) can then be rewritten as

$$\begin{bmatrix} b_1^A \\ b_2^A \\ \vdots \\ b_8^A \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & \cdots & S_{18} \\ S_{81} & S_{11} & \cdots & S_{17} \\ \vdots & \vdots & \ddots & \vdots \\ S_{21} & S_{32} & \cdots & S_{11} \end{bmatrix} \begin{bmatrix} e^{j(0\Delta\phi)} \\ e^{j(1\Delta\phi)} \\ \vdots \\ e^{j(7\Delta\phi)} \end{bmatrix} \quad (3.6)$$

Because of the discrete rotational symmetry of the structure and the applied excitation, if the reflected voltage of any element is set to zero

$$b_1^A = S_{11}e^{j(0\Delta\phi)} + S_{12}e^{j(1\Delta\phi)} + \cdots + S_{18}e^{j(7\Delta\phi)} = 0 \quad (3.7)$$

$$b_2^A = S_{81}e^{j(0\Delta\phi)} + S_{11}e^{j(1\Delta\phi)} + \cdots + S_{17}e^{j(7\Delta\phi)} = 0 \quad (3.8)$$

Active matching of the antenna

The antenna elements and the feed chain are designed and optimized separately. The incident and the reflected voltages of the antenna elements in Fig. 3.3 are the transmitted and reflected voltages respectively on the feed chain as shown in Fig. 3.4.

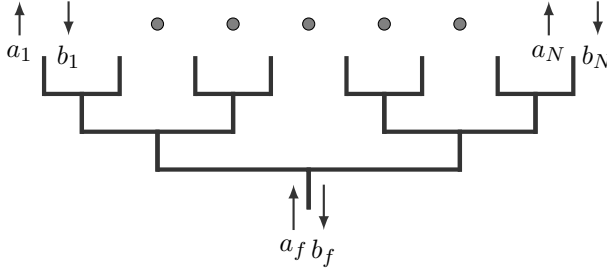


Figure 3.4: Scheme of the feed chain.

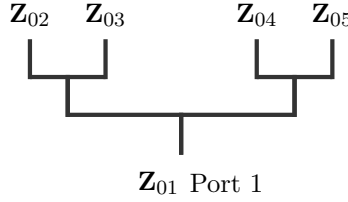


Figure 3.5: A microwave network with ports having different characteristic impedance.

Considering an ideal and lossless feed chain, there is no reflection voltage ($b_f = 0$), the power of the incident voltage (A_f) is equally divided into the different ports with the proper phase shift according to ϕ .

Generalized scattering matrix

A microwave network can be defined with ports having different characteristic impedance, as shown in Fig. 3.5.

In this case, the generalized scattering parameters should be used. These are defined as

$$S_{ij} = \frac{b_i}{a_j} = \frac{V_{0i}^- \sqrt{Z_{0j}}}{V_{0j}^+ \sqrt{Z_{0i}}} \quad (3.9)$$

These generalized scattering parameters are verified with the simulation results of Microwave CST. The circuit is a 5-port feed network with one excitation entrance and four ports as shown in Fig. 3.5. The scattering matrix is obtained twice. First, the scattering matrix is computed with normalized parameters in every port with respect to $80 \, \Omega$.

The scattering matrix, S , when $Z_{01} = Z_{02} = 80 \Omega$ is

$$S|_{\substack{Z_{01}=80 \\ Z_{02}=80}} = \begin{bmatrix} 0.22 + j0.13 & 0.29 - j0.36 & 0.37 - j0.29 & 0.34 - j0.32 & 0.26 - j0.39 \\ 0.39 - j0.25 & 0.25 - j0.20 & 0.26 + j0.12 & 0.17 + j0.20 & 0.40 - j0.52 \\ 0.31 - j0.35 & 0.26 + j0.09 & 0.22 - j0.14 & 0.39 - j0.53 & -0.33 + j0.05 \\ 0.37 - j0.28 & 0.16 + j0.21 & 0.47 - j0.46 & 0.26 - j0.20 & 0.26 + j0.11 \\ 0.28 - j0.37 & 0.33 - j0.56 & -0.33 + j0.05 & 0.28 + j0.04 & 0.15 - j0.21 \end{bmatrix} \quad (3.10)$$

In the second case, the generalized scattering matrix is computed with different characteristic impedance in their ports ($Z_{01} = 50 \Omega$, $Z_{02} = 80 \Omega$). The scattering matrix is then

$$S|_{\substack{Z_{01}=50 \\ Z_{02}=80}} = \begin{bmatrix} -0.03 - j0.00 & 0.20 - j0.44 & 0.30 - j0.38 & 0.27 - j0.40 & 0.17 - j0.45 \\ 0.34 - j0.34 & 0.11 - j0.26 & -0.21 - j0.14 & 0.21 - j0.04 & -0.35 + j0.60 \\ 0.23 - j0.43 & -0.22 - j0.12 & 0.18 + j0.11 & -0.31 + j0.62 & -0.10 + j0.26 \\ 0.31 - j0.37 & 0.21 - j0.05 & -0.41 + j0.56 & 0.14 - j0.25 & -0.22 - j0.12 \\ 0.20 - j0.44 & -0.26 + j0.64 & -0.10 + j0.26 & -0.23 - j0.09 & 0.21 + j0.02 \end{bmatrix} \quad (3.11)$$

Now, the voltage in Port 2 is computed as

$$\begin{aligned} V_2 &= \sqrt{Z_{02}}(a_2 + b_2) \\ &= \sqrt{Z_{02}}(a_2 + S_{21}a_1 + S_{22}a_2 + S_{31}a_3 + S_{41}a_4 + S_{51}a_5) \end{aligned} \quad (3.12)$$

Considering that only Port 1 is excited, the voltage is computed as

$$V_2 = \sqrt{Z_{02}}(a_2 + b_2) = \sqrt{Z_{02}}S_{21}a_1 = \frac{\sqrt{Z_{02}}}{\sqrt{Z_{01}}}S_{21} \quad (3.13)$$

From the two different matrices (3.10) and (3.11), the voltage is

$$V_2 = \frac{\sqrt{80}}{\sqrt{50}}S_{21}|_{\substack{Z_{01}=50 \\ Z_{02}=80}} = 0.3981 - 0.2506j \quad (3.14)$$

$$V_2 = \frac{\sqrt{80}}{\sqrt{80}}S_{21}|_{\substack{Z_{01}=80 \\ Z_{02}=80}} = 0.3981 - 0.2506j \quad (3.15)$$

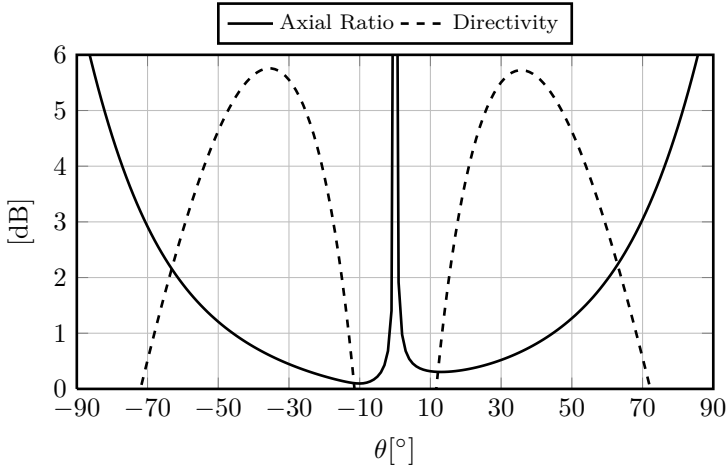


Figure 3.6: Simulated directivity and AR at $\phi = 90^\circ$.

The characteristic impedance of $80 \, \Omega$ is chosen because the slot antenna has very high input impedance, and thus it would be easier to match with the $80 \, \Omega$ characteristic impedance. Higher impedance line could have been chosen but the stripline would then need to be very narrow, which is not desired for design tolerance.

3.3.3 Results

The radiation pattern and the AR for $\phi = 90^\circ$ are given in Fig. 3.6. The directivity is maximum at $\theta = 35^\circ$, and is at least 3 dB for $20^\circ < \theta < 60^\circ$. The AR is less than 3 dB for $5^\circ < \theta < 70^\circ$. Fig. 3.7 also shows the radiation patterns for different θ , with excellent azimuthal symmetry.

The magnitude and phase of the scattering parameters of the feed network at 8.415 GHz are shown in Table 3.2. The power is divided equally in the 8 different feeds (or -9 dB at each feeding port). The power attenuates as the wave traverses a longer line length, which leads to a reduced efficiency. The phase difference at each consecutive port is within $\pm 8^\circ$ from 90° . The axial ratio and pattern symmetry are extremely sensitive to even minor variations in excitation values.

Finally, the slot array is integrated with the feed network. The directiv-

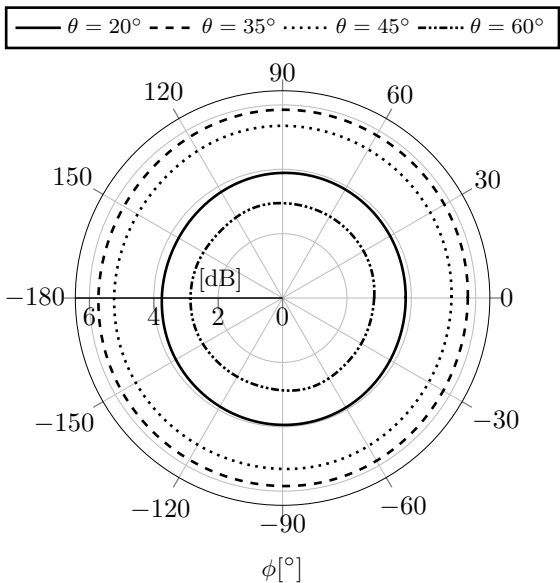


Figure 3.7: Simulated radiation pattern cuts for different angles (θ).

Table 3.2: Excitation of the hybrid feed network.

Excitation	Amplitude [dB]	Phase [°]
1	-9.060	0
2	-9.197	-82.45
3	-9.742	187.51
4	-9.267	89.78
5	-9.100	-0.29
6	-9.240	-82.80
7	-9.660	186.90
8	-9.089	89.40

ity of the integrated slot array for different values of θ is given in Fig. 3.8. There is a marked difference in terms of azimuthal symmetry as compared to Fig. 3.7. This is mainly due to imperfect power distribution at each feeding port. Further optimization of the feed network by adjusting the width and length of the quarter wave transformers can yield a symmetric pattern. Fig. 3.9 shows the active reflection coefficient of the integrated feed design. The design has a -14 dB bandwidth from 8.35 GHz to 8.43 GHz.

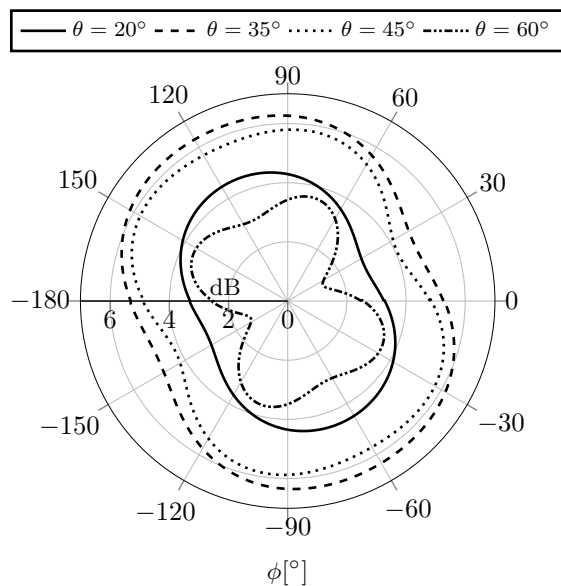


Figure 3.8: Simulated radiation pattern cuts of the integrated design at different angles from broadside.

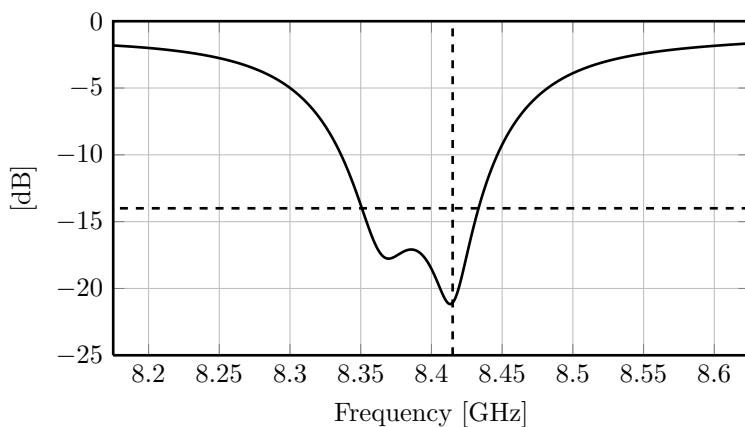


Figure 3.9: Simulated active reflection coefficient of the circular array of slot antennas with integrated hybrid feed network.

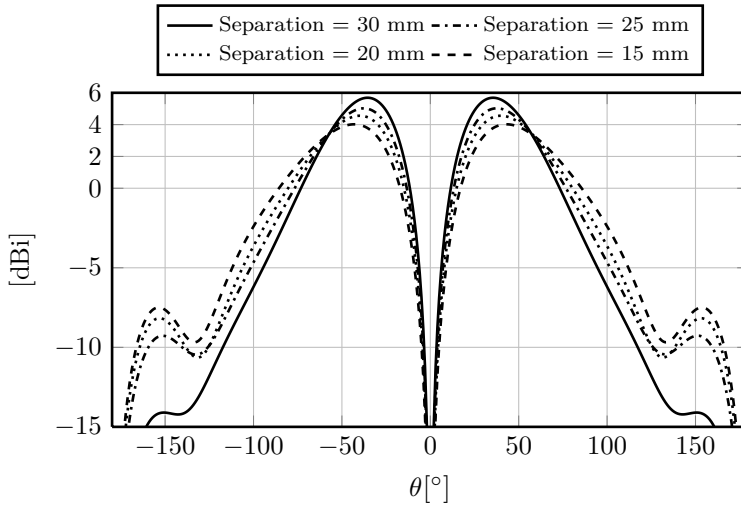


Figure 3.10: Simulated directivity plots at $\phi = 45^\circ$ for different separations of slot elements.

3.4 Circular Patch Antenna

The circular slot array had the potential to have excellent azimuthal symmetry (see Fig. 3.7 and Fig. 3.8) given the possibility to feed it at multiple locations, which subsequently suppressed the non-resonant modes. However, this required the slot elements to remain further apart, which then raised the pattern to higher elevation, which can be seen in Fig. 3.10 for eight slot elements in a circular array. Decreasing the slot separation indeed increases the gain at lower elevation but suffers from substantial decrease in gain at higher elevations. As can be seen from Fig. 3.10, there is a significant energy spillage at the backlobes when slot separation is smaller. This can also be verified from the analytical model of Fig. 3.11(a), where a current source is rotating in a circular array. Fig. 3.11(b) shows the normalized power pattern and axial ratio for different separation of the source.

Despite the possibilities to have excellent azimuthal symmetry and axial ratio as seen in Fig. 3.11(b), the circular slot array required higher number of feed points, which makes the feeding network complex and large. In pursuit of conforming to the LaRa specifications (see Table 1.1) of high-

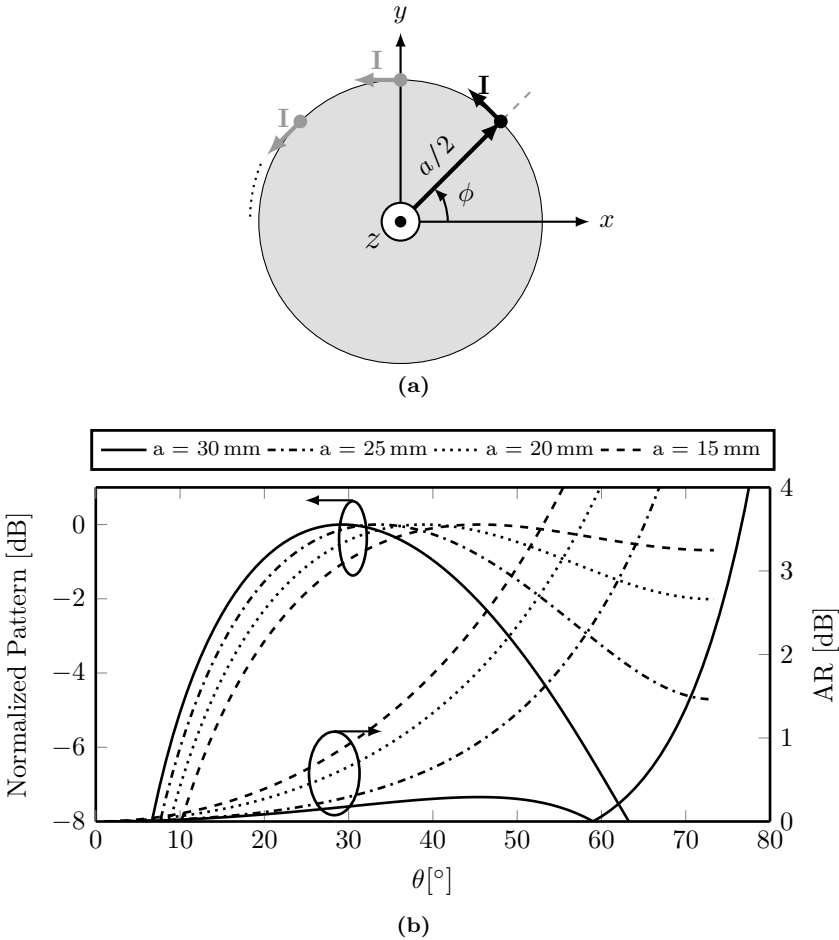


Figure 3.11: Patterns in a circular array: (a) representation of the sources in the array, (b) normalized power patterns and axial ratio.

gain patterns at low elevation and size requirements, the design concept had to be changed.

In fact, circularly polarized conical beam antennas had been widely studied in the literature but not extensively for the low-elevation applications. Huang [6] provided one of the earliest design guides to generate circularly polarized conical beam with a maximum at different elevation angles. To that end, he proposed the use of two or four feed points at various locations around the patch on dielectric substrates with different permittivities to excite higher-order resonant modes in the patch. For instance, a patch antenna etched on a dielectric substrate of permittivity $\epsilon_r = 4.2$ exciting a TM_{41} mode would theoretically generate a conical beam at $\theta = 69^\circ$ with a peak directivity of 4.0 dBi. A theoretical treatment of the transverse magnetic (TM) resonant modes of a microstrip patch antenna modeled as a cylindrical cavity with magnetic walls is given by Derneryd [59]. Vaughan [60] has extended the analysis to higher order modes using modal expansion technique.

This section explores the full design implementation and test of a circular patch antenna operating at TM_{31} resonant mode based on the design criteria put forth by Huang [6]. This design includes an integrated feed network, which is built directly underneath the patch antenna. Furthermore, the LaRa antenna specifications, for which the patch design was initially envisaged, required a probe-fed design. This imposed a complex fabrication process since the design also had to be qualified for space application.

3.4.1 Parametric Study

The different TM modes of a patch antenna exhibit peculiar pattern characteristics. For instance, Fig. 3.12 shows the elevation patterns of the patch antenna resonating at different TM modes. Exciting a higher order mode lowers the elevation patterns (Table 3.3).

All these patches were made to resonate at 8.415 GHz, with good axial ratio (see Fig. 3.13) and finite ground planes. Only two substrates (RT6002 and RT5880LZ, see Table 3.1) were considered for this study since they were found to be the most appropriate for use in space.

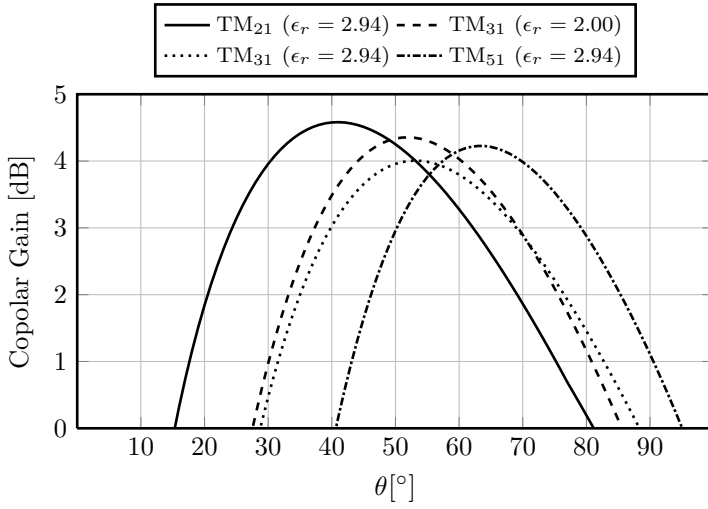


Figure 3.12: Best-case directivity of CP patch antennas resonating at different TM_{mn} modes.

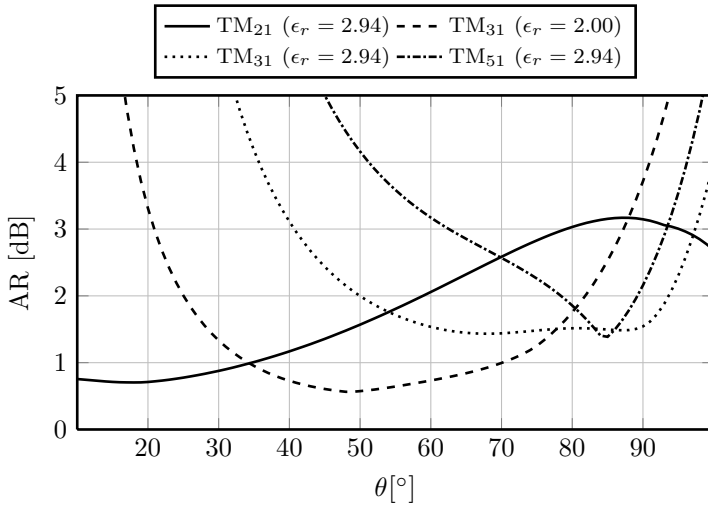


Figure 3.13: Worst-case AR of CP patch antennas resonating at different TM_{mn} modes.

Table 3.3: Gains of CP patch antennas operating at different modes.

Mode	θ_{max}	Directivity	Patch Diameter
TM ₂₁ ($\epsilon_r = 2.94$)	41°	4.58 dB	0.54 λ_0
TM ₃₁ ($\epsilon_r = 2.00$)	52°	4.36 dB	0.94 λ_0
TM ₃₁ ($\epsilon_r = 2.94$)	53°	4.01 dB	0.76 λ_0
TM ₅₁ ($\epsilon_r = 2.94$)	63°	4.23 dB	1.19 λ_0

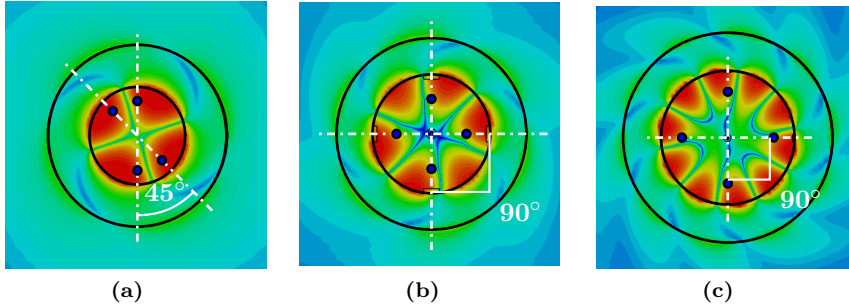
**Figure 3.14:** Electric field distribution in (a) TM₂₁ (b) TM₃₁ and (c) TM₅₁ resonant patch antennas.

Fig. 3.14 shows the E-field distribution on the patch antenna for the aforementioned TM modes. Note that the angular spacing of the feed locations of TM₂₁ mode is 45°, while that of TM₃₁ and TM₃₁ is 90°. The phase distribution at the four feed locations for TM₂₁ is $[0^\circ, 90^\circ, 0^\circ, 90^\circ]$, and that for TM₃₁ and TM₅₁ is $[0^\circ, 90^\circ, 180^\circ, 270^\circ]$.

Finally, the TM₃₁ mode design was chosen, because it had a good balance between the high and low elevation angles according to the requirements for LaRa antennas (see Table 1.1). Also, the feeds for TM₃₁ mode design are physically separated by 90°, improving the pattern symmetry in azimuth.

3.4.2 Design of Patch Antenna

The fully assembled antenna consists of an antenna PCB mounted on a choke-ring housing, and fed with a semi-rigid cable assembly. The assembly with the housing and the cable assembly is detailed in Appendix A. This subsection details the design of the PCB, its stackup and the feed network.

Microstrip Patch Antenna

The exploded view of the antenna layer stackup is shown in Fig. 3.15. It consists of four substrate layers of Rogers RT/Duroid 6002 [57] and four copper layers. The thickness of the two bottom substrate layers is each 0.508 mm while the thickness of the top layer is 1.016 mm. Because of the long lead time and high cost of 1.016 mm substrate, two layers of 0.508 mm RT 6002 were stacked. Having a thicker substrate layer at the antenna level increases the radiation efficiency and the impedance bandwidth [38]. The thicknesses of the top and bottom copper layers are 35 μm initially, but via-metallization after all the layers are stacked adds up about 7 μm for a total of 42 μm . The different substrates are bonded with pre-preg layers of Rogers RO 2929 [61]. The finished thicknesses of the pre-preg layers vary, from 38 μm to 152 μm . A central contact is soldered to the *feed network* of the PCB as it is sequentially processed. This is done to avoid surface mounting an SMA connector. With the soldered central contact, a field-replaceable SMA connector can easily mate with the PCB. The overall thickness of the PCB is about 2.4 mm. The final PCB stackup with vias and feed port integration is shown in Fig. 3.16.

The top part in Fig. 3.15 corresponds to the microstrip circular patch antenna. The little dots in the patch correspond to the feed points, where pins cross the *Ground Plane* 1. The little notches in the patch enable the azimuthal smoothing of the radiation pattern, and also change the impedance of the antenna. Having only four feeding points concentrates the fields around the feeding locations (Fig. 3.17). The notches then help to reduce the field concentration around the feed location and distribute the fields evenly throughout the patch. The changed field distribution influences the overall impedance seen at the patch. The patch diameter directly impacts the resonance of the antenna, whereas the radius of the feed points impacts the depth of resonance.

The patch antenna is excited at four different locations with a 90° angular separation. They are excited with the same amplitude and 90° phase difference. This excitation provides a circular distribution of the current on the surface of the patch.

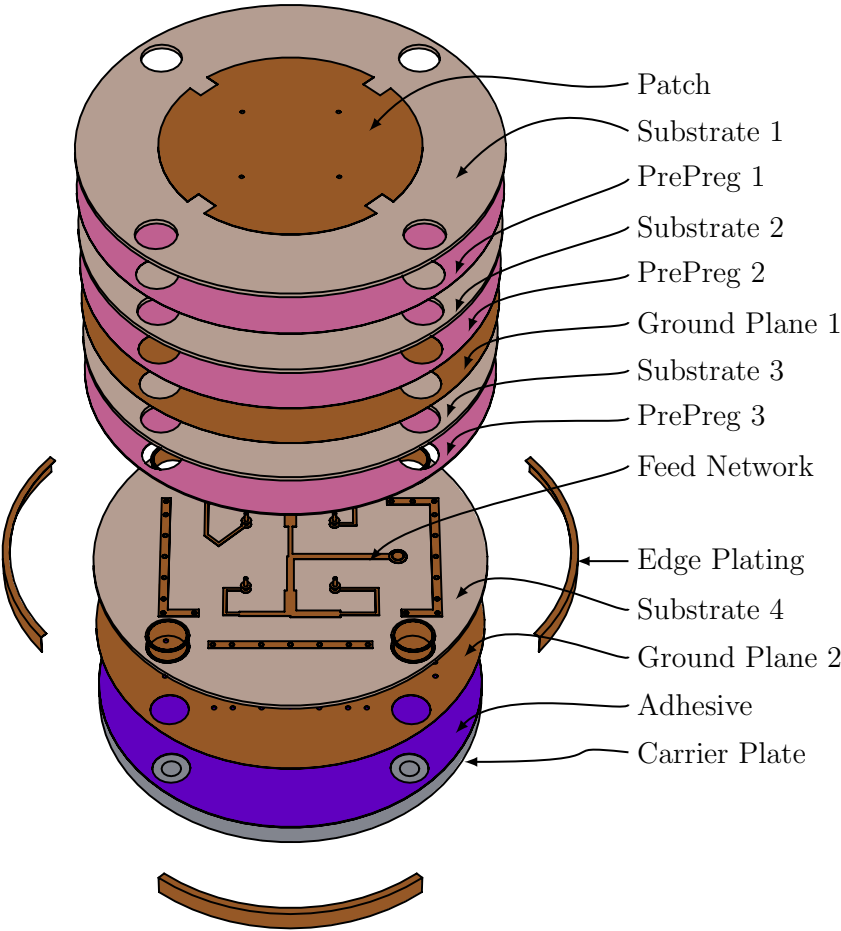


Figure 3.15: Patch antenna layer stackup exploded view.

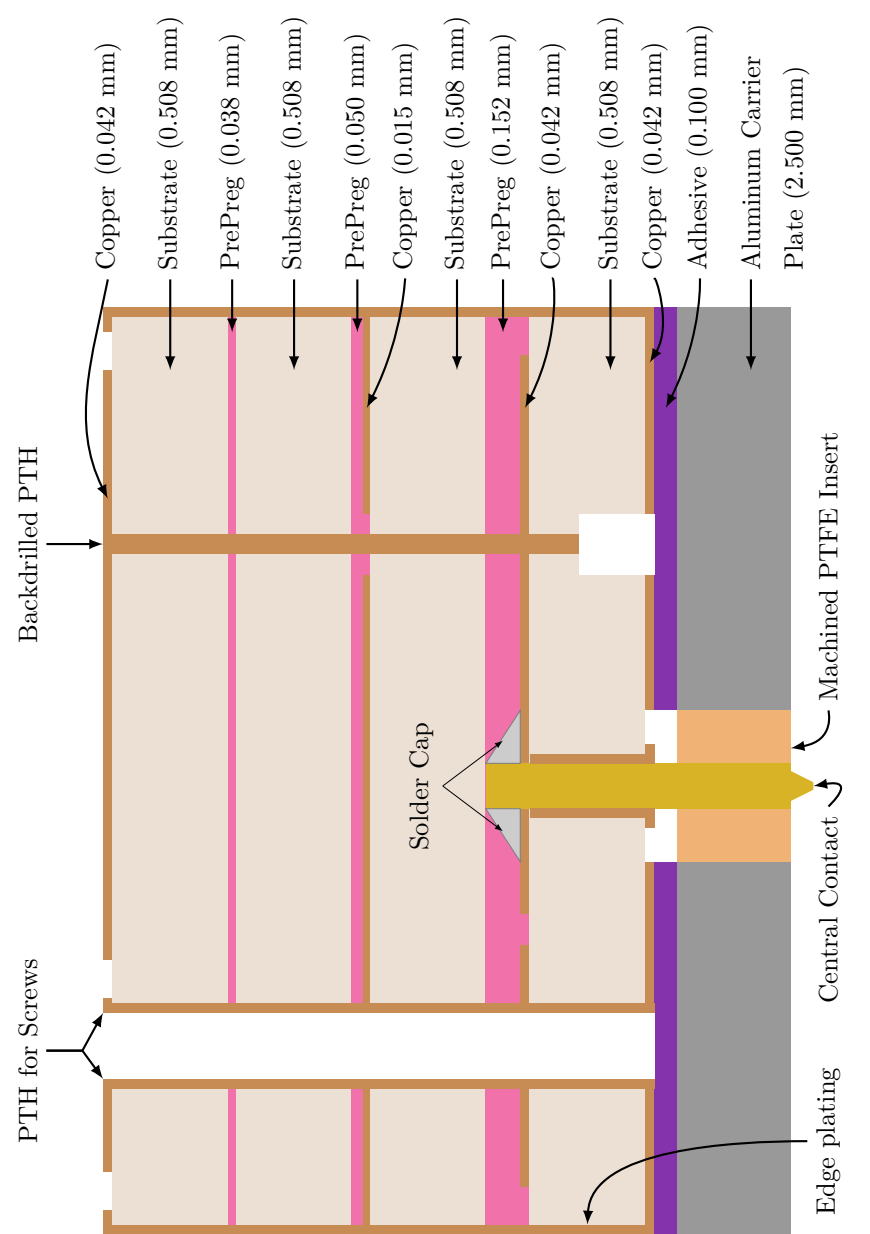


Figure 3.16: PCB Stackup of the CP patch antenna.

Feed Network

Fig. 3.18 shows the distribution network, which is realized in stripline technology, in the layer beneath the *Ground Plane* 1 layer. Stripline technology has been chosen to avoid back radiation and improve isolation. Given the desired phase shifts between consecutive ports, hybrid circuits [62, Section 7.5] [56] could have been used. Hybrids often offer proper phase shift and bandwidth over a relatively large bandwidth (order of 5%). However, those have been found difficult to house in the limited space available for the stripline circuits; this is why the feeding circuit has been based on splitters, delay lines and matching sections. The desired bandwidth being only of the order of one percent, this approach has been found acceptable. Special attention has been given to avoid very thin traces and to keep a sufficient distance from the screws that hold the PCB. Optimization has also been performed versus the radial position of the four feed points, which mainly impacts the antenna impedance, though it also influences other performance parameters.

The bars in Fig. 3.18 help to suppress the standing waves in the cavity formed in the feed network. This can be seen in Fig. 3.19(a), when the bars have not been used. When the bars are used, the field distribution becomes more uniform (see Fig. 3.19(a)), and the phase rotation within the feed network becomes sequential. The sequential phase rotation is required to ensure that the phases of the currents at the four ports of the feed network are separated by 90° . The inclusion of the bars seems sufficient to shift the resonance well outside the frequency band of interest. The surface integrated waveguide (SIW) technique was also explored but it was not possible to fit the many thin vias near the feed network. Such vias rather played a role of coupling one line with the other and, thus, was discarded in favor of less intrusive bars.

The design of this patch antenna is discussed further in Appendix A. It also includes the design parameters and the methodology employed. The antenna and feed network are first designed separately and then integrated. The automatic optimizer available from Microwave CST is used for the full integrated design.

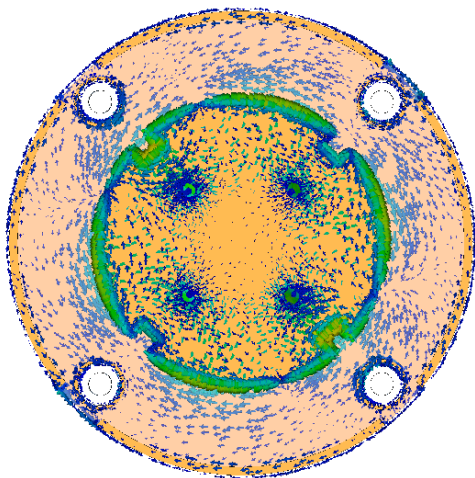


Figure 3.17: Current distribution on the CP patch antenna.

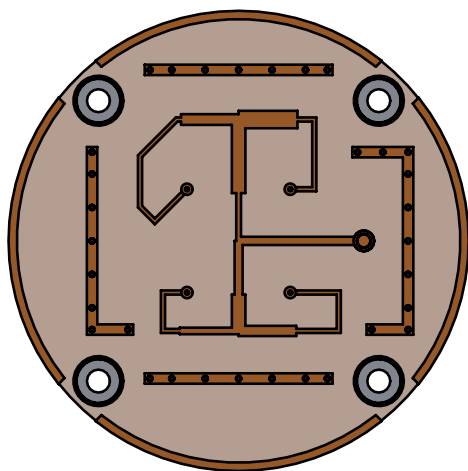


Figure 3.18: Stripline feed network (top view) of the CP patch antenna.

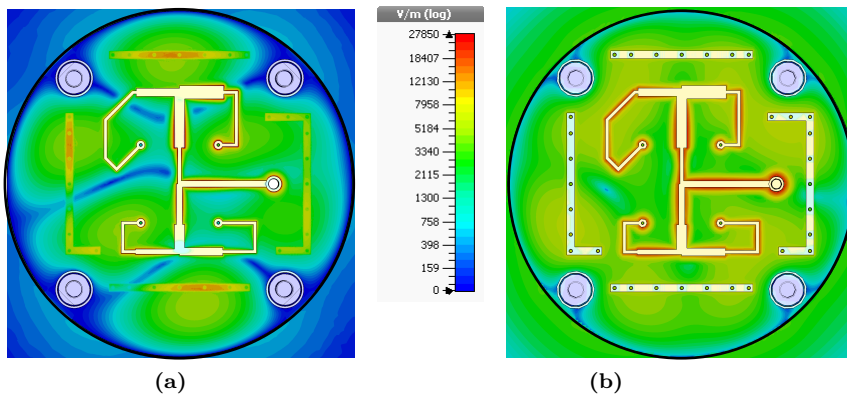


Figure 3.19: Electric field distribution (a) without and (b) with the bars in the feed network.

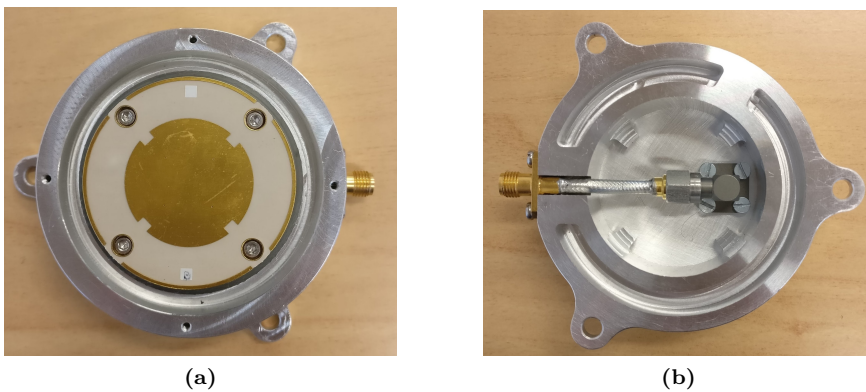


Figure 3.20: Fully assembled CP patch antenna: (a) Top view with the patch antenna and (b) Bottom view with the cable assembly.

3.4.3 Measurement Results

The patch antenna was fabricated and tested. The fully assembled and manufactured patch antenna is shown in Fig. 3.20. Note that the cable assembly in Fig. 3.20(b) has a straight semi-rigid cable instead of a helical loop as designed (see Fig. A.3). The helical loop was not possible to manufacture without envisaging a specific design by the cable manufacturer. Instead, for measurement purpose, a straight assembly was deemed sufficient.

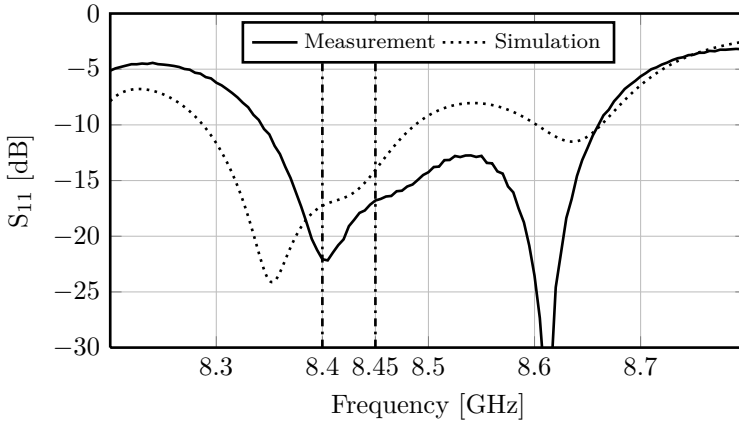


Figure 3.21: Reflection coefficient of the circular patch antenna.

Reflection Coefficient

The reflection coefficient of the measured and simulated CP patch antenna is shown in Fig. 3.21. The CP patch antenna was designed for operation in the [8.4, 8.45] GHz band, with a carrier frequency at 8.415 GHz. At the band of interest, the S_{11} values remain below the required -14 dB.

Copolar Gain

The best- and worst-case copolar gains at the central frequency, 8.415 GHz, and at 8.45 GHz are given in Fig. 3.22. The best-case copolar gain refers to the maximum copolar gain values at each elevation angle (or θ) for all azimuth angles (ϕ), while the worst-case copolar gain refers to the minimum copolar gain at those angles. The insertion losses of the connectors and the cable loss of the semi-rigid cable assembly add up to 0.4 dB, which is included posteriori to the simulation results, since the simulation is carried out without the cable assembly. As can be seen from the figures, the best-case copolar gain shows good correspondence with the measured results. However, the worst-case copolar gain is significantly lower due to the asymmetric nature of the pattern, especially at 8.415 GHz. Interestingly, the measured copolar gains at 8.450 GHz show good correspondence with the simulated ones.

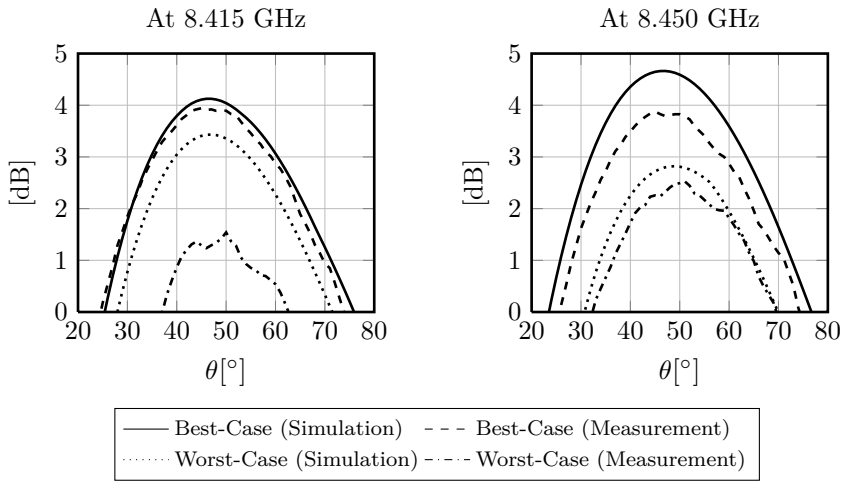


Figure 3.22: Best- and worst-case copolar gains of the CP patch antenna at 8.415 GHz and 8.450 GHz.

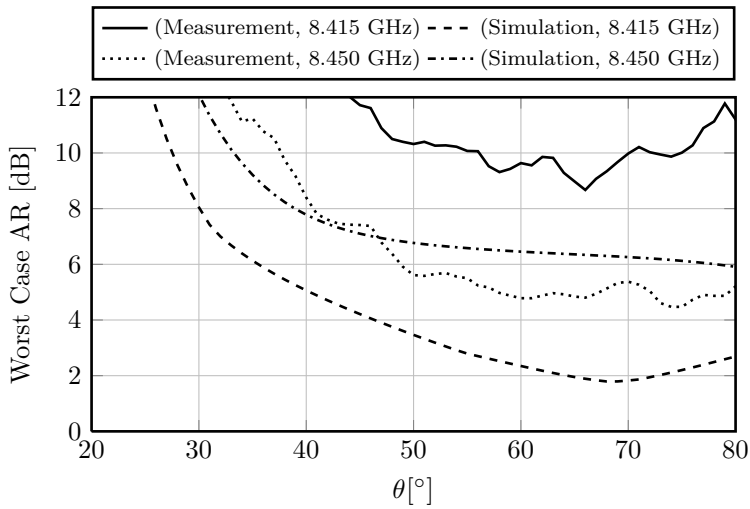


Figure 3.23: Worst case AR over θ of the CP patch antenna at 8.415 GHz and 8.450 GHz.

Axial Ratio

The worst-case axial ratios over different operational bands are shown in Fig. 3.23. The worst-case again takes the maximum value of the AR at a given elevation angle for all azimuth angles. The AR is much worse in measurement, compared to the simulation. This could be due to fabrication tolerances of the feed network. As in the case of copolar gain, the measured AR at 8.450 GHz shows good correspondence with the simulated one, even though the absolute values are much higher than the required 3 dB.

3.5 Conclusion

Chapter 3 detailed the design of circularly polarized conical beam planar antennas. A circular slot array and a circular patch antenna were explored. Even though the gain of the circular slot array was optimal at higher elevations, the patch antenna provided a simpler design (fabrication and feed network) with superior gain performance at low elevation. This provided a motivation to build and test the patch antenna, since it matched closely to the LaRa specifications.

However, the fabrication process of the patch antenna had to adhere to the strict design rules of the PCB laid out by the ECSS [63] guidelines for a space product, which complicated the PCB fabrication process particularly because of using vias in a multi-layered PCB structure. This required a full qualification of the PCB instead of a *delta* qualification because of the large temperature variation. Given this complexity, the measured results still have a meaningful relation with the simulated ones. Nevertheless, corrections would be quite long and costly since some iterations will be required with the PCB manufacturer. This *a posteriori* confirms the validity of the technological shift to adhere to the LaRa specifications, from patch to 3D monopole parasitic antenna. The latter will be the subject of Chapters 4 and 5.

CHAPTER 4

Circularly Polarized Conical Beam 3D Metallic Antenna

This chapter discusses the 3D metallic monopole parasitic antenna capable of producing conical patterns with circular polarization. This design is an evolution of the CP conical beam antennas explored in Chapter 3. The circular patch antenna had a complex fabrication process and a lack of heritage for space qualification, and an alternative had to be found. The all-metal radiator, manufactured through additive manufacturing or milled out of a single block of metal, has superior RF performance as well as good mechanical stability. It remains impervious to large temperature variations, which makes it ideal for use in space applications.

Section 4.2 gives the detailed description of the antenna, its working principle, and the design methodology. Section 4.3 provides a parametric study. The antenna gain optimization and comparison with the existing designs in literature are discussed in Section 4.4. Section 4.5 provides a comparison with the measured results of the 3D printed prototype.

4.1 Introduction

This chapter presents a novel and robust antenna design based on parasitic elements placed around the centrally-fed monopole, capable of generating excellent circularly polarized (CP) conical patterns at low elevation. Made largely of metal, the 3D monopole parasitic antenna can be used in harsh climatic conditions.

Huang [6] proposed the use of a circular patch fed with two or four probes to excite the higher order modes of the circular waveguide. This work is a definitive, if not comprehensive, guide to generating CP conical-beam patterns for different peak directivities using different substrates and resonant modes. The use of the four probes, however, requires a feeding network, adding losses, increasing the overall size, and possibly requiring multiple layers of dielectric substrates. Kawakami *et al.* [64] proposed another CP conical beam antenna relying on the introduction of four inclined monopole elements placed in a circular array. The inclination angle and the separation between the elements would help generate CP in the desired direction. Even though no phase shifters were required, the antenna still needed a feed network. Lau and Luk [7] presented a wideband CP conical beam patch antenna using L-probes and apertures connected through a hybrid feed network. The antenna has a complex microstrip feeding network, which is prone to back radiation because of exposed feed network.

Recent works show the trend towards simplifying the feed network, with a central pin exciting the antenna system [65] [66] [67] [68] [69] [70] [71]. Nakano *et al.* [65] proposed a wideband (3.3 – 9.6 GHz) CP conical beam low-profile two-arm spiral antenna. Pan and Leung [66] investigated a centrally fed wideband antenna surrounded by eight parasitic dielectric parallelepipeds. The structure is complex [68] and not very robust, making the integration of the dielectric material tedious. Besides, the Axial Ratio (AR) bandwidth is easily influenced by the size of the ground plane [69]. Qi *et al.* [67] proposed a single-feed circular aperture CP antenna by simultaneously exciting the TM_{01} and TE_{01} orthogonal modes in a circular waveguide to produce CP.

Lin *et al.* [68], Yu *et al.* [69], and Xu *et al.* [71] showed similar designs for

a centrally fed printed patch, contributing primarily to the horizontal polarization, with loop stubs [68], modified elliptical-ring slots [69], or annular ring slots [71], contributing mainly to the vertical polarization. They employ different resonant modes (mainly TM_{01} and TM_{02} of the patch) to improve the wideband performance of the antennas. The AR performance of the loop stubs antenna is easily affected by the size of the ground plane, which is not the case for the slot array designs. All of these designs [68], [69], [71], however, produce high-elevation patterns (θ between 28° and 35°). Reducing the number of loop stubs or elliptical-ring slots effectively lowers the elevation angle but the gain and AR are then sharply degraded [68]. Kumar and Harish [70] proposed a torus knot radiator fed with a central monopole, but the torus knot is complicated to manufacture although it might be easier with additive manufacturing. Also, the gain value is less than 3 dB at $\theta = 30^\circ$. Ohmine *et al.* [72], [73] reported annular-ring microstrip antennas for mobile satellite communication with high gain at low-elevation angles (gain of at least 3 dB for $30^\circ \leq \theta \leq 60^\circ$).

It is apparent from the literature review that a true CP conical beam antenna for low elevation application is lacking. High gain at low elevation is difficult to achieve with planar structures. Chenhu *et al.* [74] proposed a slot array on a truncated cone to generate CP conical beams, whose angle of maximum gain would depend on the tilt angle of the cone. While this design is easier to adapt to any beam-pointing angle and the overall gain can reach 5.8 dB, it still needs dielectric support conforming to the structure.

4.2 Antenna Description

The proposed antenna corresponds to a monopole above a ground plane surrounded by N parasitic posts (Fig. 4.1, where $N = 8$). The latter are symmetrically arranged around the monopole (rotation by $2\pi/N$), and are protruding from the ground plane. They have a curled (or gamma) shape in order to exhibit horizontal parts that generate horizontally polarized fields, and create, together with the radiation from the vertical monopole, conical patterns with circular polarization. The ground plane is supporting the radiating elements, which is itself supported by a cylindrical aluminum

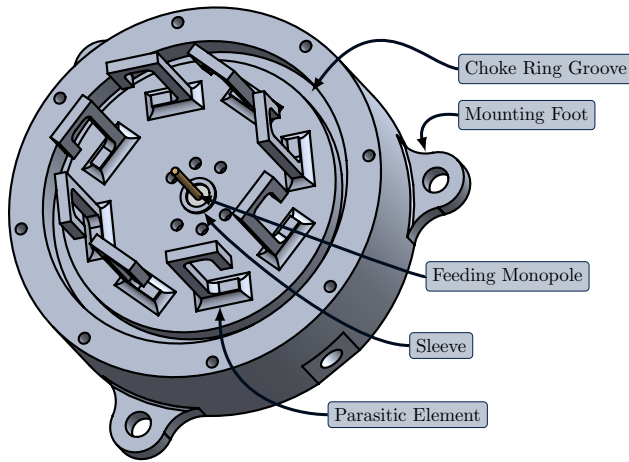


Figure 4.1: Perspective view of the monopole parasitic antenna.

housing, in which one or several grooves are made. Such grooves can act as choke rings, which tend to stop the current, and so limit back-radiation [75]. The outer wall of the choke ring is made relatively thick, so as to support a radome.

To attach the monopole feed, several holes are made around the housing (the holes in the center support both a 4- or 2-hole square flange receptacle). In addition, a tube surrounding the monopole, called monopole sleeve, provides a degree of freedom to tune the antenna impedance [76]. It creates an additional capacitance in parallel to the antenna impedance, and allows matching over a wider bandwidth.

To ease fabrication with precision machining and to make the structure robust, the antenna has fillets around the base and the inner corners of the parasitic elements. The three feet at the base of the antenna make it easy to mount the antenna atop a platform with nuts and bolts.

4.2.1 Working Principle

When only the monopole is present, given the symmetry of the structure, only vertical polarization is obtained. However, a relatively broad conical pattern is expected, since monopoles with axial symmetry cannot radiate

(in the far field) along their axis. The parasitic elements present some horizontal parts, given their curled shape. This forces horizontal currents and, hence, fields with horizontal polarization (see Fig. 4.2). If those fields have the same magnitude as the vertically-polarized ones, and if they are phase shifted by 90° with respect to them, then the polarization is circular.

The horizontal segments of the parasitic elements cannot be placed at arbitrary position, since they need to be mechanically attached to the housing. This connection is ensured by a vertical post, which also produces fields with vertical polarization. However, they may also interfere negatively with the fields radiated by the central monopole, such that a good balance between the vertically and horizontally polarized fields remains possible.

When just a vertical and a horizontal arm are considered, the induced current on the latter remains quite small simply because it is near the end of the wire, at which the longitudinal current is necessarily equal to zero. To allow substantial current, the structure is extended by a vertical arm that proceeds downward. The additional vertical arm cancels the vertical fields coming up from the original vertical post [77] such that there is a balance in the horizontal and vertical field distributions. The current at the end of the parasitics along that structure must tend toward zero. Therefore, a little vertical extension has been added, such that the current on the horizontal part remains substantial. It may also be viewed as a purely reactive extension allowing better radiation, as often found in folded monopoles [77]. Finally, to a lesser extent, it may also contribute to cancel the fields from the main vertical post.

Hence, the main characteristic dimensions of the parasitic elements correspond to their distance from the monopole, the lengths of the two vertical parts and the length of the horizontal parts. The cross-section of the parasitic elements plays a secondary but a non-negligible role. That cross-section is mainly governed by mechanical strength. The little horizontal bar at the foot of the parasitic element can marginally improve the antenna radiation performance. This also contributes to the strength of the structure. Given the relatively simple shape of the parasitic elements, the whole conducting structure can be milled out of a single block of aluminum or can be produced through an additive manufacturing process.

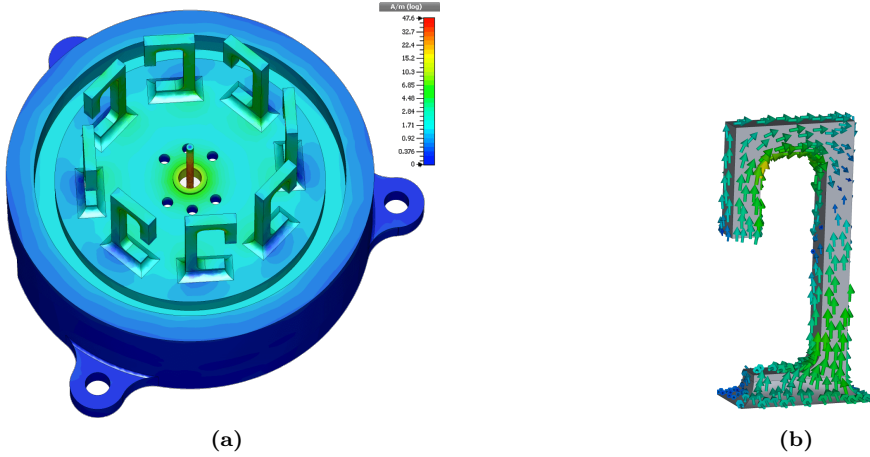


Figure 4.2: Current distribution on the (a) whole antenna, and (b) single element, as obtained with CST Microwave Studio.

4.2.2 Design Parameters

The proposed design is intended for operation in the X-band (8.4 GHz to 8.45 GHz) with a central frequency of 8.428 GHz. Fig. 4.3 shows the dimensions of the monopole parasitic antenna along with those of a single parasitic element. Here, r gives the separation between the parasitic elements and the monopole feed, and $(d2 - d1)/2$ gives the gap in the choke ring. All the fillets in the parasitic element (see Fig. 4.3(b)) have a radius of 1.5 mm. Fig. 4.4 shows the detailed view of the vertical section of the choke ring and that of the central monopole feed. The choke ring parameters, dt and dc , are measured from the base of the parasitic elements. The monopole height is given by lm , and the sleeve length, computed from the base of the parasitic elements, is given by ls . The thickness of the base plate, on which the parasitic elements rest and the monopole feed can be screwed, is given by tg . A dielectric tube acts as a spacer to hold the monopole in place. In practice, an SMA connector with extended central contact can replace the monopole and dielectric, and the dielectric portion of the connector may be cut to the height of the base plate (tg). The diameter of the central connector is 1.3 mm and the thickness of each of the parasitic elements is 2 mm. Table 4.1 summarizes the values used in the antenna design.

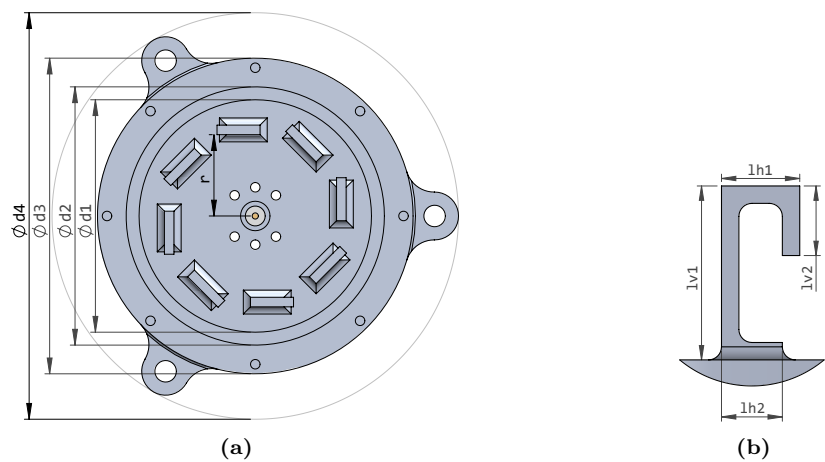


Figure 4.3: Dimensions of the monopole parasitic antenna: (a) top view and (b) parasitic element.

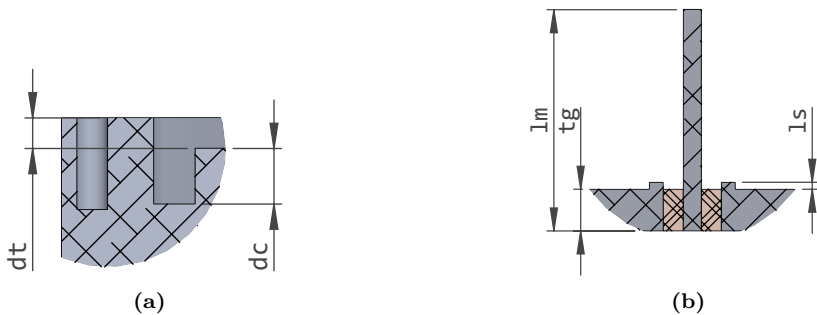


Figure 4.4: Dimensions of (a) the choke ring and (b) the monopole feed.

Table 4.1: Design Parameters of the Antenna

Parameters	Value (mm)	Parameters	Value (mm)
r	17	$lh1$	9
$d1$	48.6	$lh2$	7
$d2$	56	$lv1$	20
$d3$	66	$lv2$	8
$d4$	85	lm	15.9
dc	3.64	ls	0.5
dt	2	tg	3

4.2.3 Design Methodology

The monopole parasitic antenna can be tuned to operate at different frequencies. Following steps can prove useful in the initial design of the antenna:

1. For design frequency, f , and wavelength, λ , choose $r = \lambda/2$. Fine tune r to change resonance.
2. Choose the angle of maximum gain, θ_{max} , such that $lv1 = r \cdot \tan \theta_{max}$
3. The lengths of the arms of a parasitic element equal a half-wave resonant dipole, i.e. $lh1 + lv2 = \lambda/2$

With these initial values, it is possible to get a good initial performance. To improve the performance, it is required to further fine-tune or optimize the the geometrical parameters.

For cases when angle of maximum gain is increased, $\tan(\theta_{max})$ rapidly increases and, thus, $lv1$ also increases. From a practical point of view, such large $lv1$ may not be possible to fabricate. Therefore, for optimum gain at low elevation, the parameters in Table 4.3 maybe scaled for design at any desired frequency.

The parametric study in Section 4.3 can prove useful to understand the effect of various design parameters to achieve good CP conical patterns.

4.3 Parametric Study

Several parameters of the antenna play important roles in generating CP conical beam patterns. Understanding these different parameters is important to help design the antenna to meet any stringent specifications of CP conical beam patterns at any desired frequency.

The parameters under study are: separation between the monopole and the parasitic element, effect of the larger ground plane, as well as the different dimensions of the parasitic element. The length of the monopole is fixed at 15.9 mm, which is the length of a standard SMA connector with extended central contact. The length protruding out of the base plate is 12.9 mm.

Without the parasitic elements, the antenna would resonate at 5.8 GHz. For the parametric study, the mounting feet have been removed, and in some cases, also the fillets around the monopole elements. The mounting feet introduce some small ripple in the azimuthal gain whereas the fillets have negligible impact on antenna electromagnetic performance.

4.3.1 Monopole Height

A standard SMA connector can be directly screwed to feed the antenna, which makes the feed system extremely simple. However, it is possible to tune the antenna or improve its bandwidth with different monopole heights (lm). The -10 dB impedance bandwidth is, on average, around 7.6%.

4.3.2 Monopole to Parasitic Element Separation

The monopole to parasitic element separation (r) is another important parameter for tuning the antenna (Fig. 4.5). However, the antenna patterns need to be re-optimized at the shifted frequency by modifying all the parameters of the parasitic elements. In terms of impedance matching, roughly speaking, the frequency is inversely proportional to r .

4.3.3 Parasitic Element Parameters

The number of parasitic elements, N , mainly affects the azimuthal gain symmetry (see Fig. 4.6).

Even though the vertical length, $lv1$, does not affect the reflection coefficient very much, it plays a role in improving gain at different elevation angles (Fig. 4.7). A higher element favors increased gain at lower elevation angles. The other vertical length, $lv2$, however, affects the reflection coefficient (Fig. 4.8) but does not affect the overall gain at any elevation angles. This length is crucial to create a balance in the vertical polarization, and, thus, impacts more the axial ratio.

The horizontal length, $lh1$, contributes again in balancing the horizontal and vertical polarizations (reducing the AR), and also affects the reflection coefficient (Fig. 4.9).

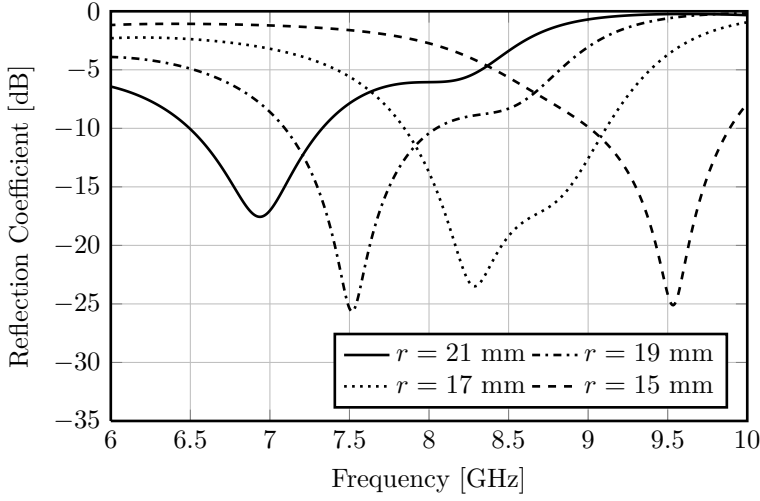


Figure 4.5: Reflection coefficient for different separation of the monopole and the parasitic elements.

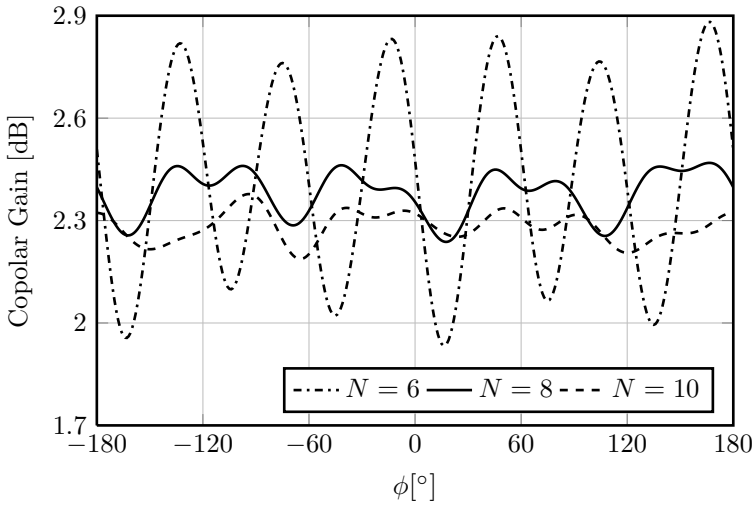


Figure 4.6: Azimuthal gain variations at $\theta = 70^\circ$ for different numbers of parasitic elements.

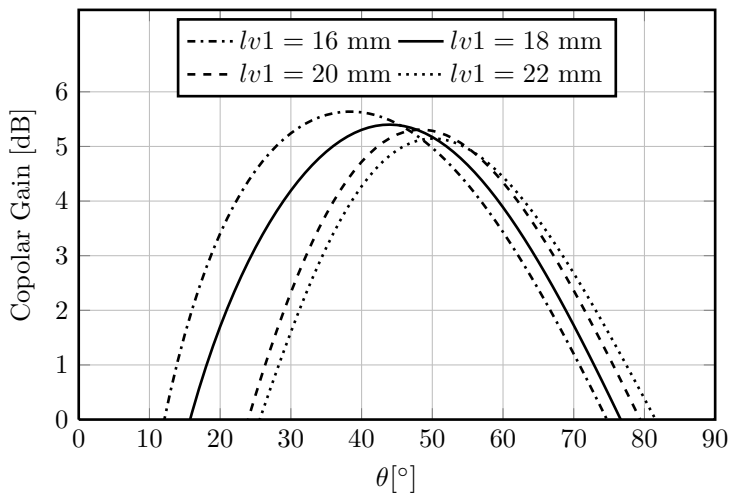


Figure 4.7: Copolar gain at $\phi = 0^\circ$ for different heights (l_{v1}) of parasitic elements.

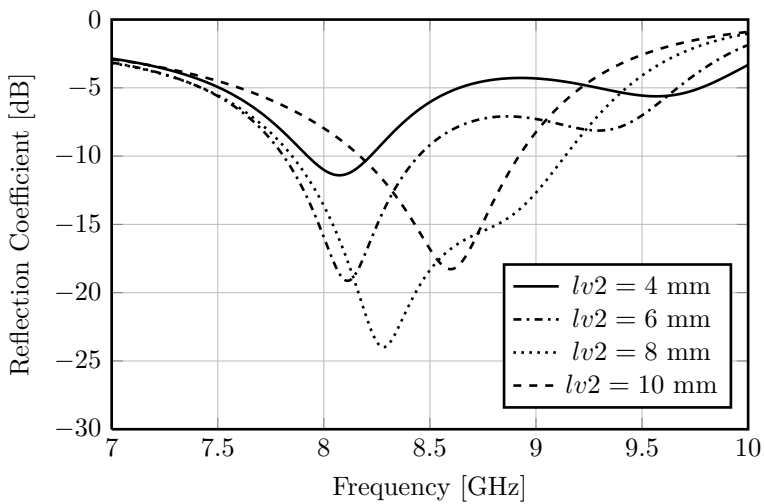


Figure 4.8: Reflection coefficient for different heights (l_{v2}) of parasitic elements.

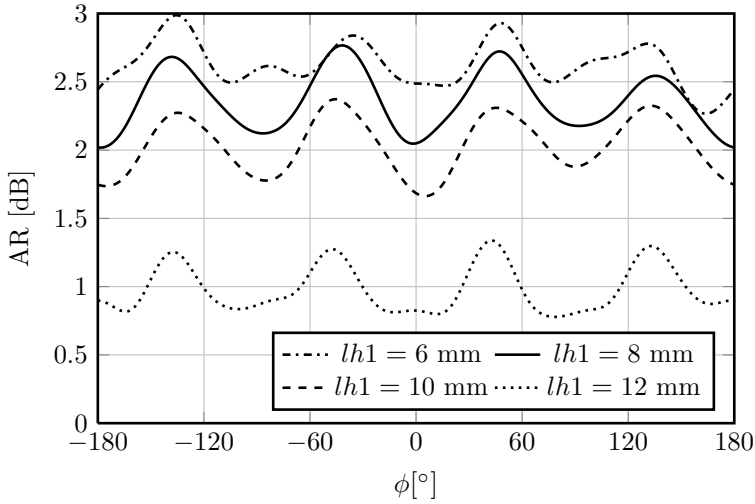


Figure 4.9: AR at $\theta = 70^\circ$ for different horizontal lengths ($lh1$) of parasitic elements.

4.4 Gain Optimization and Comparisons

One of the advantages of the proposed design is its flexibility to change the angle of maximum gain. Depending on the applications, the antenna may be optimized to perform at a range of elevation angles, $\theta_{max} = 35^\circ$ to 70° .

Table 4.2 summarizes the most important performance metrics at the intermediate angles between 35° and 70° at 8.428 GHz, and compares these metrics to those obtained in the literature. Note that the values given in Huang [6] are peak directivity. As such, they do not consider the dielectric and feed network losses. Some modes proposed by [6] require excitation pins that are unevenly distributed along the patch azimuth. This may lead to large azimuth variations of the radiation pattern and difficulties to balance and match the feeding network [78]. In this respect, the peak directivity provided for the TM_{41} mode should be considered as quite theoretical [79].

Table 4.3 shows the parameter values used to obtain the performance metrics of Table 4.2. Some parameters such as lm , ls , and tg have the same values as in Table 4.1.

The simulated copolar gain, axial ratio and reflection coefficient of the

Table 4.2: Comparison between proposed antenna optimized for different angles of maximum gain and the literature.

θ_{max}	Gain (dB) [min, max]	AR (dB)	S11 (dB)	3-dB BW*	Reference Value (dB)	[Reference] Antenna Type
35°	[7.02, 7.21]	< 1.7	−32.0	36°	7.2 [†]	[69] Elliptical slot
50°	[6.47, 6.70]	< 1.3	−27.7	30°	6.2 [‡]	[6] Circular patch
60°	[5.13, 5.35]	< 0.5	−17.5	39°	4.8 [†]	[80] Drooping dipole
70°	[4.06, 4.30]	< 2.3	−17.5	25°	4.0 [‡]	[6] Circular patch

*Beamwidth, [†]Gain, [‡]Directivity**Table 4.3:** Parameter values used in proposed antenna optimized for different angles of maximum gain.

Parameters (mm)	Angle of Maximum Gain (θ_{max})			
	35°	50°	60°	70°
r	18.42	17.02	33.31	33.39
$d1$	70.84	70.84	104.62	110.77
$d2$	92.54	97.0	125.37	120.77
$d3$	106.0	120.0	137.0	144.0
dc	12.0	12.82	6.0	6.0
dt	0.0	−0.94	3.0	−3.0
$lh1$	8.55	7.37	11.5	12.0
$lh2$	11.0	4.1	12.0	8.0
$lv1$	12.55	20.5	23.0	24.0
$lv2$	5.78	7.0	2.0	2.0
N^*	10	10	13	14

*Dimensionless

antennas optimized for maximum gain at different angles are shown in Figs. 4.10, 4.11, and 4.12.

While the gain optimization at higher elevation is easier, the gain at lower elevation suffers from additional energy spilled over due to sidelobes at higher elevations (Fig. 4.10). The antenna structure is relatively large (almost 4λ), for the design optimized for low elevation angles.

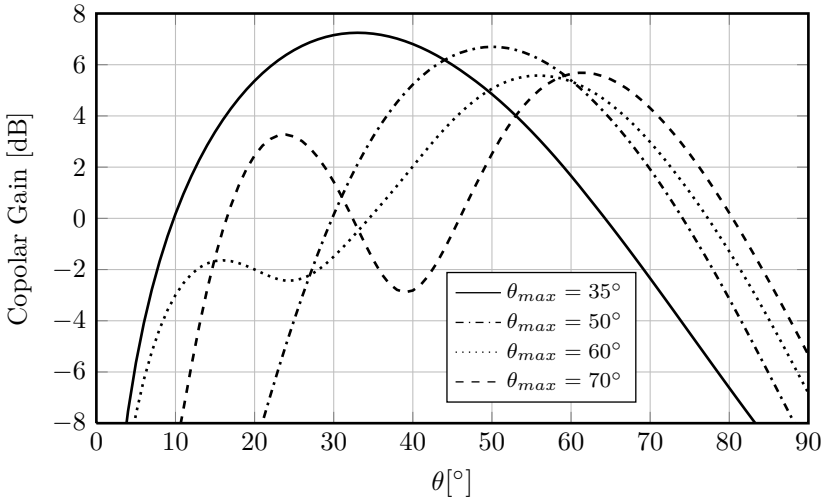


Figure 4.10: Best case copolar gains of the gain optimized antennas at 8.428 GHz.

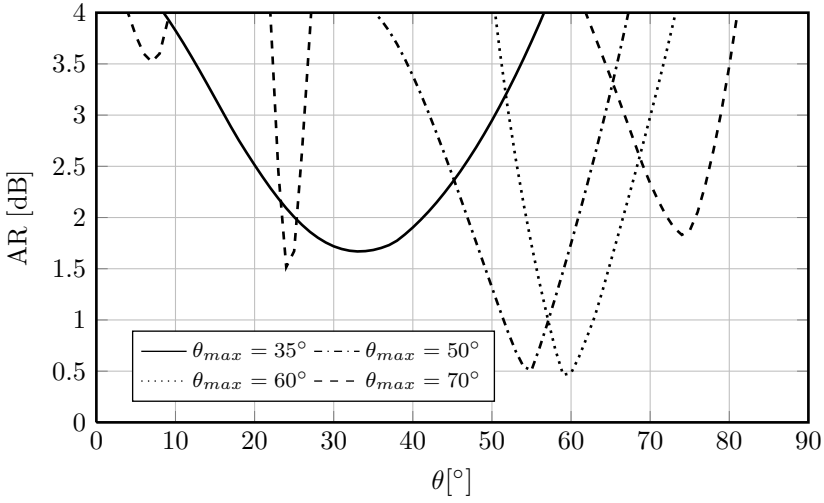


Figure 4.11: Worst case AR of the gain optimized antennas at 8.428 GHz.

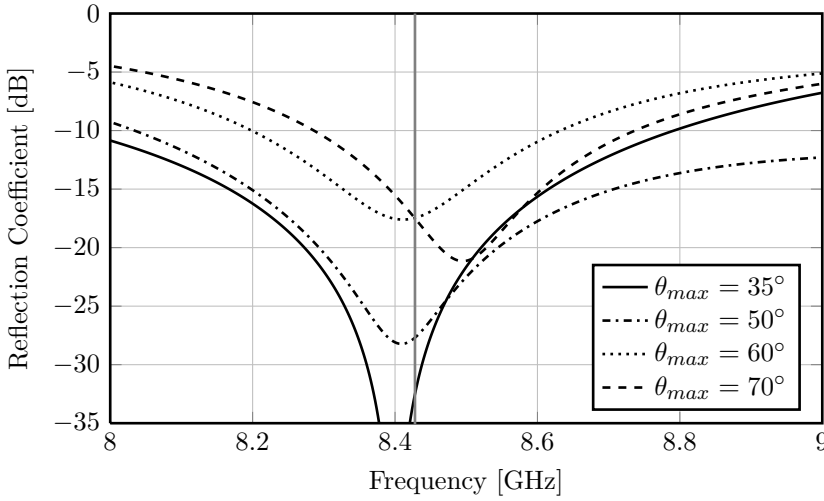


Figure 4.12: Reflection coefficients of the gain optimized antennas.

4.5 Experimental Validation on a Prototype

For rapid prototyping, the antenna was fabricated using metal additive manufacturing (Fig. 4.13). While a complex feed structure is not necessary for this type of antenna, a semi-rigid cable assembly was built (Fig. 4.13(b)) to make it possible to feed the antenna from a lateral connector, and to make calibration easier during the measurement activity.

For the measurement of the reflection coefficient (Fig. 4.14), two feed setups are considered: a direct bottom feed using a straight SMA connector with extended central contact and a side feed using a semi-rigid cable assembly. The semi-rigid cable assembly consists of a right-angle plug, a 0.085" RG405 semi-rigid cable and a 2-hole flange jack connected to a right-angle SMA connector with extended central contact acting as the feeding monopole. The curved semi-rigid cable was required to accommodate the assembly, and also to meet the cable length requirements for space application. The measurement result with the bottom feed shows close correspondence with the simulated result (Fig. 4.14). The difference is probably due to the manufacturing imperfections. The measurement result with the semi-rigid cable assembly is quite different from the simulated result because of

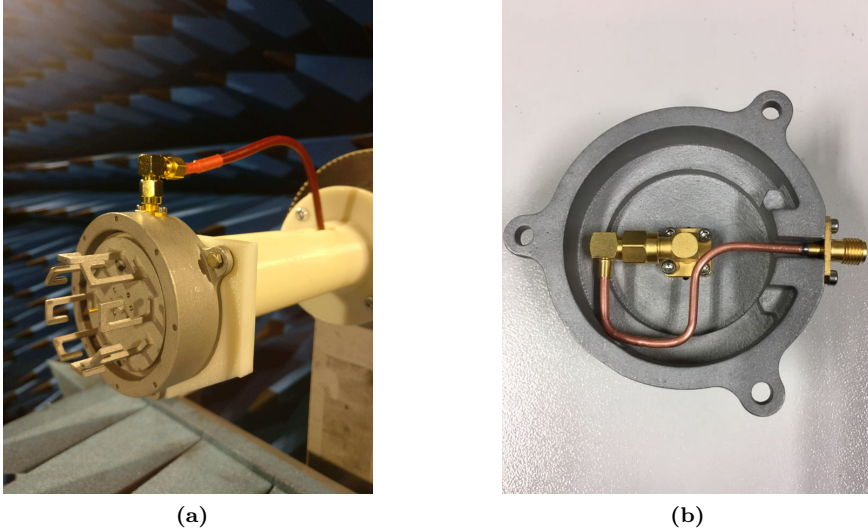


Figure 4.13: The 3D printed prototype antenna (a) in the anechoic chamber, and (b) with the semi-rigid cable feed assembly.

the change in the cable properties resulting from the sharp bends. In this case, the effect of the cable assembly is not considered in the simulation. Fig. 4.15 shows the best- and worst-case copolar (right hand circular polarization, RHCP) gains at each elevation angle, where gain is positive, over the whole azimuthal plane. Best- and worst-case gain mean the largest and the smallest value over all azimuth angles, for all angle θ of interest, respectively. This plot gives the best and worst values of copolar gain (RHCP) for the elevation angles in the upper hemisphere $[0^\circ, 90^\circ]$ at 8.428 GHz. The measurement results show good correspondence with the simulation results (Fig. 4.15). The azimuthal gain variations are poorer in measurement but are still smaller than 1 dB for $30^\circ \leq \theta \leq 70^\circ$.

Fig. 4.16 shows the worst-case AR of the simulated and 3D printed prototype. Worst-case AR means the largest value over all azimuthal angles for the angle θ of interest, where $AR < 3$ dB for $39^\circ < \theta < 68^\circ$. Also, the simulated AR bandwidth for the prototype is given in Fig. 4.17, which comes about 8.8%.

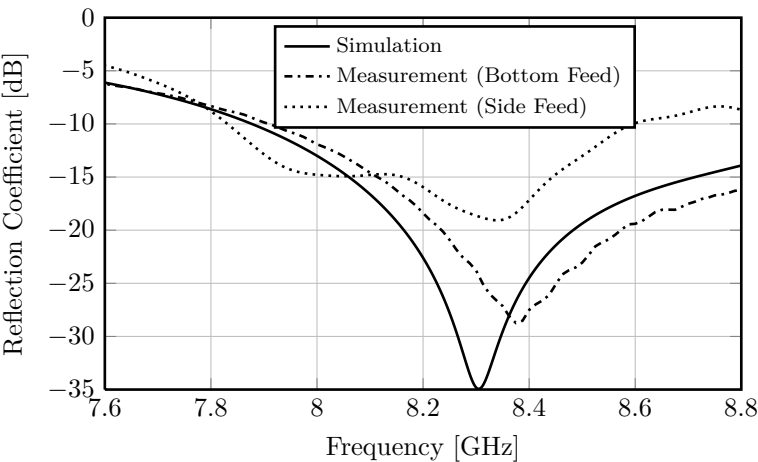


Figure 4.14: Reflection coefficient of the measured prototype.

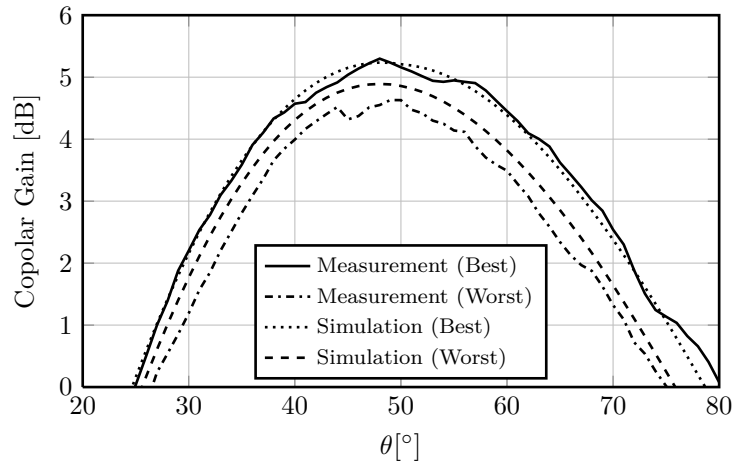


Figure 4.15: Best and worst case copolar gains at 8.428 GHz.

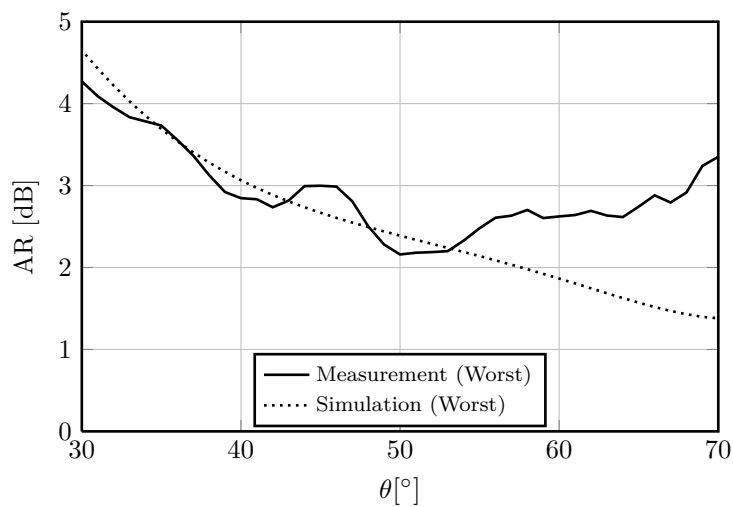


Figure 4.16: Worst case AR at 8.428 GHz.

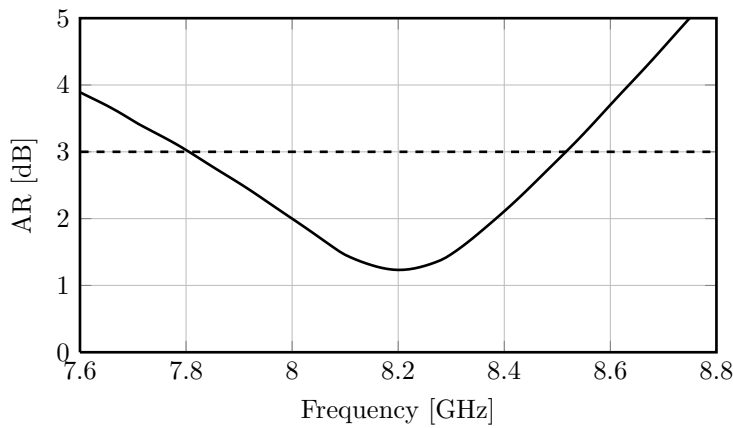


Figure 4.17: AR bandwidth of the prototype at $\theta = 50^\circ$.

4.6 Conclusion

This chapter presented a novel metallic antenna based on a monopole feed surrounded by several gamma shaped metallic parasitic elements, which produces a circularly polarized conical-beam. The distinct shape of the parasitic element with its horizontal and vertical arms help create a balance in the horizontal and vertical currents responsible for generating circular polarization. The antenna is suitable for use in harsh environmental conditions, and is easily optimized for high performance at different elevation angles. Because of its 3D structure, the antenna can offer high gain at low elevation. Simulation results show that gain above 4 dB at $\theta = 70^\circ$ is possible. A prototype was manufactured using additive manufacturing, which showed encouraging results as compared to the simulation results.

Chapter 5 builds on the design philosophy explored in Chapter 4, and applies it to the LaRa instrument. More specifically, the chapter will expand on the design adaptations to the space environment, in terms of protection from the environment and the low-loss feed network. Multi-physics aspects like the thermoelastic and the electrostatic effects on the design will also be discussed.

CHAPTER 5

Monopole Parasitic Antenna for the LaRa Instrument

Chapter 5 serves as an extension to the 3D monopole parasitic antenna described in Chapter 4. Because of the superior RF performance, mechanical stability, and resistance to the thermoelastic and electrostatic effects, the monopole parasitic antenna has been chosen to fly on the ExoMars 2020 mission to Mars as part of the LaRa instrument. This chapter, therefore, expands on the performance criteria, and design adaptations pertaining to the LaRa instrument.

Section 5.1 introduces the complete LaRa design specifications and explains the design evolution from a conical frustum array to the 3D monopole parasitic antenna. It will also touch upon the design adaptations required for making the antenna flight-worthy. Section 5.2 expands on the further improvements to the antenna design like the square coaxial transmission line and the radome. Section 5.3 explores the multi-physics aspects of the design from the RF electromagnetic to thermoelastic effects. As a space product, the LaRa antennas need to undergo several qualification tests, mentioned in Section 5.4. Finally, Section 5.5 outlines the RF electromagnetic measurement results after the environmental qualification tests.

5.1 Introduction

The RF measurements carried out on the 3D-printed prototype (Section 4.5) validated the design concept. The antenna was also manufactured with precision machining for the Engineering Qualification Model (EQM).

5.1.1 LaRa Design Specifications

Table 1.1 provides the brief specifications for the LaRa antennas. Some additional requirements are:

- Two TX antennas operate at [8.4, 8.45] GHz band, and one RX antenna operates at [7.145, 7.190] GHz
- Lifetime of 6 years, including 2 Earth years on the Martian surface
- Withstand nominal RF power of 8 W
- Operation in the Martian atmosphere (low pressure and dust accumulation)
- Ensure thermal stability and identify uncertainties over wide range of temperature variations
- External dielectric surface materials to have volume resistivity of up to $10^9 \Omega \cdot \text{m}$ and surface resistivity of up to $10^8 \Omega$
- Resistance to the ESD discharge
- Mass of the TX antenna less than 132 grams, and that of the RX antenna less than 162 grams
- Withstand TID radiation levels of less than 500 krad
- Withstand stationary vibration exposures in flight
- Protect radio astronomy frequency bands
- Design vents to account for sudden pressure changes
- Support non-operational storage temperature of at least $+125^\circ\text{C}$ for thermal sterilization

These additional requirements imposed some design adaptations that will be discussed in detail in the following sections.

5.1.2 Evolution and Design Adaptations

The LaRa antenna design started off with a conformal frustum array with a simple analog beamforming capability (see Chapter 2). Because of the need of RF switches in the antenna, the frustum array was deemed impractical for the LaRa instrument even though it had superior RF performance. Since the instrument was at a large distance from the antennas, the idea of having yet another signal cable on top of the RF cables made the design unfeasible.

This motivated a compact design with single feed. The circular patch antenna of Chapter 5 showed early promise. But given the fabrication complexity, and the lack of heritage for such space product, the design had to be changed.

Eventually, this gave rise to the work on a simpler yet superior performing 3D metallic antenna, which is robust enough to withstand the harsh Martian environment. The choice of material, both metallic and non-metallic, has been crucial and taken into consideration as the design matured [81]. Materials with proven heritage in space have been chosen so as to reduce the qualification procedure and speed up the development of the antenna. Wherein applicable, the antennas have been manufactured, integrated and tested in accordance to the existing space standards to ensure utmost adherence and compliance to the specifications and requirements.

5.2 Design Description

There are two different LaRa antennas (TX and RX) operating at different frequency bands. The TX and RX antennas operate at the carrier frequencies of around 8428 MHz and 7174 MHz, respectively. They radiate conical patterns with a maximum in the $35^\circ - 45^\circ$ range of elevation (or $45^\circ - 55^\circ$ in θ), with right-hand circular polarization (RHCP). The TX and RX antenna diameters are 66 mm and 80 mm, respectively.

A radome also covers the whole structure, unlike in Chapter 4. The radiator itself is made of a monopole feed through a square coaxial transmission line, while the feed design in Chapter 4 included a semi-rigid cable assembly. A Duratron 1000 [82] dielectric material is used inside the square

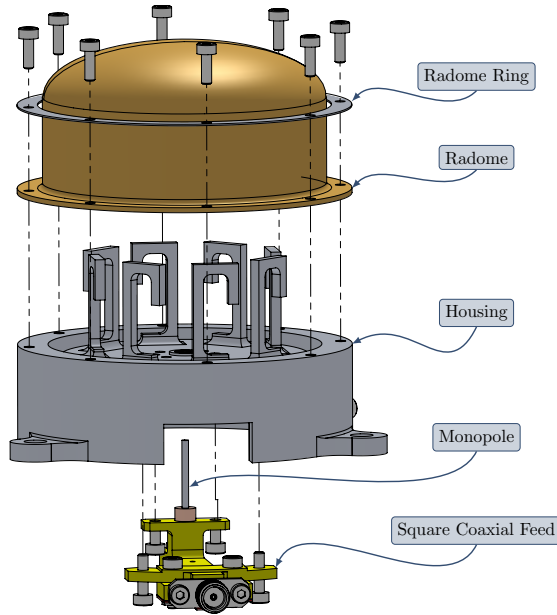


Figure 5.1: TX LaRa antenna full assembly.

coaxial transmission line to hold the long central contact and the monopole in the square coaxial transmission line. An epoxy adhesive, Scotch-Weld 2216 [83], fixes the monopole dielectric, to the antenna housing. The square coaxial transmission line is then fed from the side of the aluminum housing. The complete LaRa antenna assembly is shown in Fig. 5.1.

5.2.1 Antenna Housing

The antenna housing is the same as described in Section 4.2. The dimensions of the TX and RX vary, and the ones given in Table 4.1 correspond to the TX antenna. The design of the RX antenna started off as the scaled-up version of the TX antenna and shares the same design features as the TX antenna. But since the interface requirements of the RX antenna were different (in envelope and mass requirements) from that of the TX antenna, the optimized parameters of the RX antenna are not the exact scaled-up parameters of the TX antenna.

The parasitic elements are arranged symmetrically around the monopole, protruding from the ground plane. The aluminum housing supports the ground plane and the radiating elements. The groove in the housing acts as a choke ring. The outer wall of the choke ring is made relatively thick to support a radome.

The housing also allows the square coaxial feed transmission line to feed the antenna from the side. The other side of the feed transmission line contains the monopole. To attach the feed, there are several holes made around the center of the housing. Once the radome is fixed and staked to the housing, the enclosure can trap air inside and, thus, the holes in the center of the housing can act as a venting hole to depressurize the enclosure.

5.2.2 Square Coaxial Feed Network

A square coaxial transmission line [84] mostly filled with air feeds the antenna. This feeding line makes the connection between the monopole and the connector on the side of the housing and is located inside the antenna housing.

Fig. 5.2 shows the different parts of the square coaxial feeding line. The *top lid*, *long central contact* and the *bottom lid* are manufactured separately and assembled. The vertical *monopole* is soldered to the long central contact. The *short central contact* is commercially available and also soldered to the long central contact. All the parts are made of aluminum and are silver-plated to allow soldering. A *dielectric spacer*, Duratron 1000, supports the long central contact, which is also held tightly at the other end with a screw and a nut. The same dielectric material encapsulates the monopole within the assembly. A 2-hole field replaceable jack mating with the short central contact completes the assembly. The top surface of the dielectric material is at the same height as the top surface of the aluminum housing. The top lid also consists of two *venting holes* to allow depressurization from the square coaxial unit.

There are several parameters of interests in the design of the square coaxial feed. Most important are the cross-section parameters of the square coaxial line, which determine the impedance along the line. Also, included

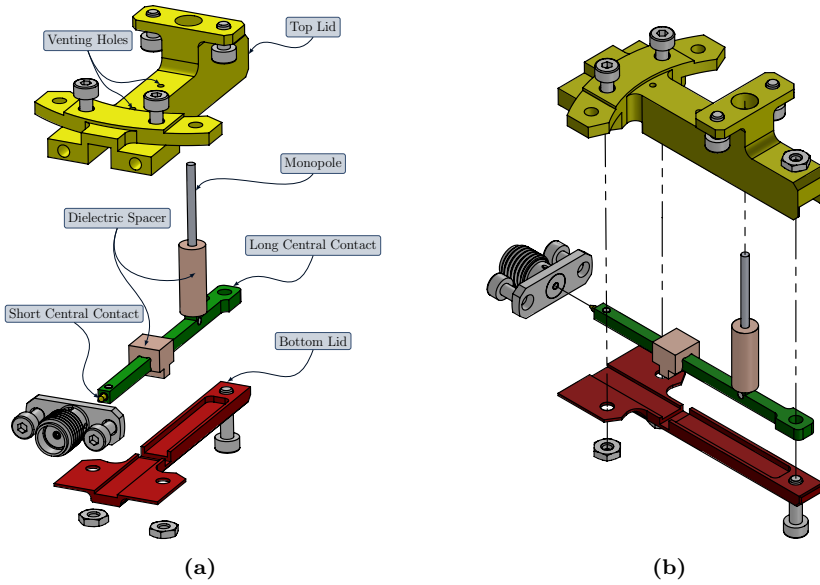


Figure 5.2: Different parts of the square coaxial transmission line.

in the design is a shorted quarter wave transformer at the end of the *long central contact*, which helps to tune the feed to the desired frequency as well as provide mechanical strength to the structure.

The impact of the transformer on the tuning and the insertion loss can be seen in Fig. 5.3 for different lengths of the transformer. A second port is added by cutting the vertical monopole and the dielectric spacer by a plane parallel to the 2-hole flange of the square coaxial top lid. For a length of 8.5 mm, the insertion loss is about 0.1 dB at 8.428 GHz. Another advantage of the shorted transformer is that it acts as a grounding for electrostatic charge that may be induced in the monopole feed.

5.2.3 Radome

A radome is necessary to protect the antenna housing and the monopole from the harsh Martian atmosphere. It also helps to avoid dust accumulation in the choke ring. HexPly 8552 [85] glass fibre was chosen as the material for the radome. The radome is fixed to the antenna housing with

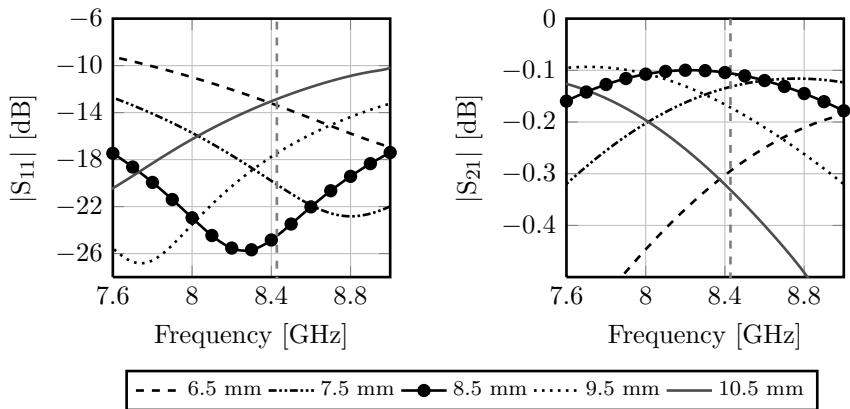


Figure 5.3: Effect of the quarter wave transformer on the square coaxial transmission line: reflection coefficient and transmission coefficient.

eight screws. Since the washers were too big to fit on top of the base of the radome, a ring was used to protect the radome material from the screws. Fig. 5.1 also shows the TX radome and the radome ring.

The radome affects the antenna performance, more specifically in terms of reflection coefficient, axial ratio and azimuthal symmetry. Fig. 5.4 shows the impact on reflection coefficient with the radome. The antenna is designed considering the radome, but the radome material dielectric properties were not known precisely a priori. A general value for the fibreglass material was used for design. That was why there was some difference in the measured and the simulated values.

5.3 Performance Analysis

5.3.1 Electromagnetic Performance

For electromagnetic analysis, the Frequency Domain Solver of CST Microwave Studio is used. The analysis is done on the full LaRa antenna assembly consisting of the antenna housing with the parasitic elements, the square coaxial feeding line and the radome. All the finer details like the fasteners, Scotch-Weld, radome ring, grounding stud are also included.

Fig. 5.5 shows the best- and worst-case simulated copolar gains of the

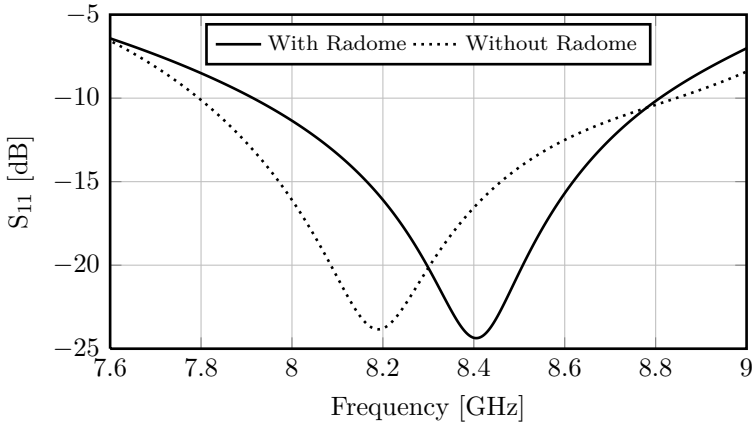


Figure 5.4: Reflection coefficient of the TX LaRa antenna with and without radome.

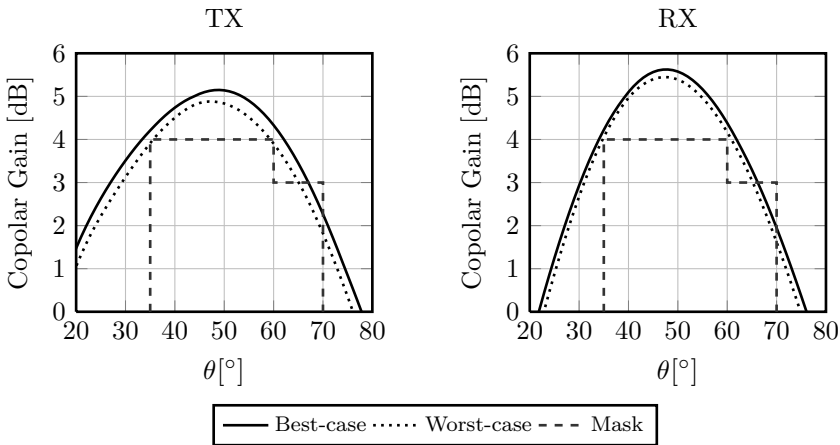


Figure 5.5: Best- and worst-case simulated copolar gains of the LaRa antennas over the TX and RX operational band.

antenna over both the TX and RX operational band with the mask. The antenna is compliant in all but the low elevation angles. Given the envelope requirements, it is difficult to fully comply with the specification.

Fig. 5.6 shows the axial ratio of both the TX and RX antennas. They are within an acceptable 3 dB band for $[35^\circ \leq \theta \leq 70^\circ]$. Only at low elevation

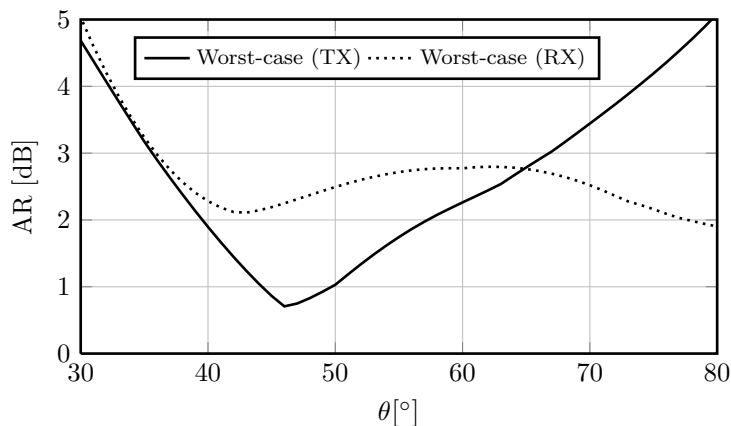


Figure 5.6: Worst-case simulated AR of the LaRa antennas over the TX and RX operational band.

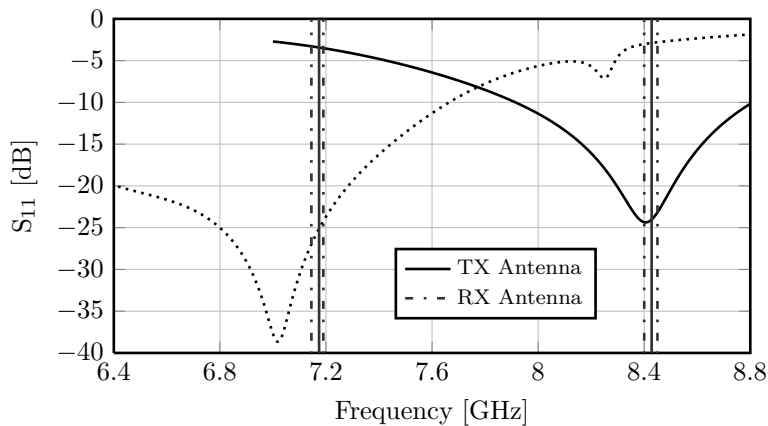


Figure 5.7: Simulated reflection coefficients of the TX and RX antennas.

for the TX is the AR higher than 3 dB.

Fig. 5.7 shows the reflection coefficients for both the TX and RX antennas. The graphs indicate excellent matching for both the antennas.

5.3.2 Electrostatic Discharge

Being an exterior surface, the radome material should have surface resistivity of up to $10^8 \Omega/\text{sq}$ to avoid electrostatic discharge (ESD). Fibreglass materials provide good RF performance and mechanical strength but have a too high resistivity to meet the ESD requirements.

Mitigation Techniques

To mitigate ESD, several alternatives were considered for the radome:

- Carbon-fibre instead of glass-fibre composite
- Conductive paint on radome
- Germanium sputtering on radome
- Discharge through radome

Adding a conductive paint would decrease the overall resistivity [86] of the radome material at a small cost in loss of RF performance. Another option is to coat Hexply 8552 with Germanium (Ge). Also, widely reported in the literature is the coating of Kapton with Indium Tin Oxide and Ge [87]. In [87], the ITO coated Kapton is found to be acceptable for RF-transparent antenna sunshield. In [88], the oxidation of Ge is shown to be problematic for storage, and a thin alumina coating is applied on top of Ge coated black Kapton to protect from environmental degradation in ground [89]. Therefore, sputtering of Germanium (Ge) on the radome material seems like a viable option. However, there is a risk of erosion of the paint or sputtered Ge in the Martian sand storms [90].

Using a carbon-fibre reinforced plastic (CFRP) instead of glass-fibre reinforced plastic (GFRP) would greatly enhance the ESD protection as CFRP materials have very low DC resistivity. The Hexply 8552 material is also supplied with carbon fibers. The various polyimide substrates like Kapton HN, 100 CB Black Kapton which are non-conductive are available from DuPont. Also available are 100XC, 160XC and 275XC Black Kapton, which are filled with conductive carbon. However, the 100XC Kapton has RF loss of about 0.3 dB at X-band compared to the Ge/100CB Kapton [91]. A comprehensive list of the materials for mitigating charging effects in spacecraft and antenna is given in [92, Section 3.2].

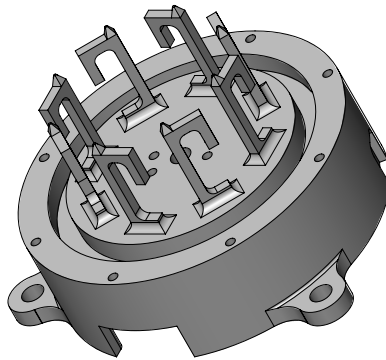


Figure 5.8: Lightning rods (little pyramids) atop the parasitic elements in the TX antenna housing.

ESD Analysis

In the absence of a suitable material, it is possible to allow the discharge to take place through the housing atop the parasitic elements. The antenna housing has a grounding stud to allow good Ohmic contact with the surface platform. Therefore, the effects of electrostatic discharge could be mitigated by adding another feature to the antenna, corresponding to *lightning rods*. These *lightning rods* have been designed as vertical extensions (pyramids) of the parasitic elements (see Fig. 5.8). With those extra features in place, the ESD discharge, if occurring, would take place along the parasitic elements through the antenna housing and not along the monopole.

The ESD effect on the antenna was studied using CST Studio Suite. The top part of the radome is covered with a Perfect Electric Conductor (PEC) sheet to indicate a charged sheet. The top sheet is at a potential of 1V and the rest of the housing is grounded. This setup is one of many cases that may arise depending on how the electrically charged Martian dust (represented as a PEC sheet on the radome) settles on the radome. When the wind is blowing, the potential distribution in the charged radome would differ according to the wind direction.

Fig. 5.9 shows a cross-section view of the electrostatic field distribution in the TX antenna. The field is higher at the tip of the parasitic element

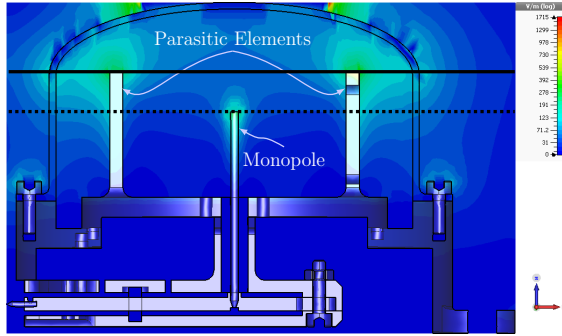


Figure 5.9: Electrostatic field distribution in the TX antenna (without the lightning rod).

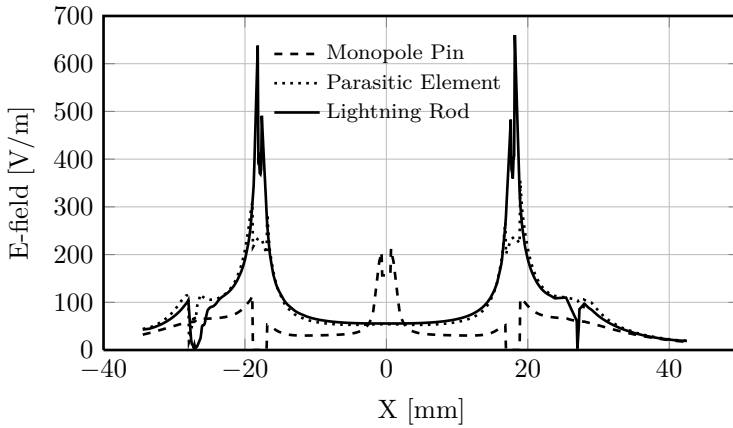


Figure 5.10: Electrostatic field distribution at the tips of the monopole and parasitic elements.

(355 V/m) than that at the tip of the monopole (212 V/m) (Fig. 5.10), therefore, any ESD discharge is highly likely to go through the parasitic elements onto the housing. With the lightning rod atop the parasitic element (Fig. 5.8), the field can reach even higher at the tip (660 V/m) (Fig. 5.10).

Should there be any electrostatic discharge at the monopole, the electric charge can also discharge along the monopole in the square coaxial transmission line to the quarter wave transformer, which is also grounded to the housing at one end.

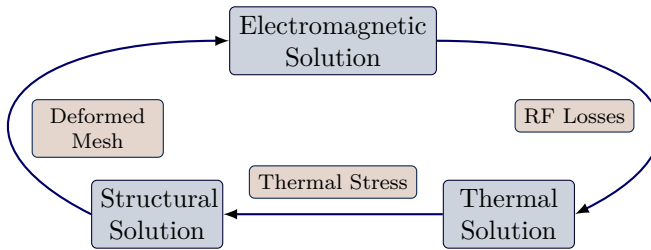


Figure 5.11: Thermoelastic analysis workflow.

5.3.3 Thermoelastic Analysis

The large temperature variations induce thermal stress on the antenna, which consequently affects its RF performance. The thermal stress includes the effects of varying ambient temperature and internal heat production because of RF losses in the conductors and dielectric materials.

Analysis Workflow

The workflow for the thermoelastic analysis is shown in Fig. 5.11. The RF losses are computed from the high-frequency electromagnetic simulation results. The input RF power of 8 W is considered for the calculation of these losses. The thermal analysis gives the temperature distribution and the thermal stress over the entire structure. The temperature distribution over the structure is used as input to the structural mechanical analysis to compute the mechanical deformations. The deformed mesh is then used to compute the antenna patterns and the reflection coefficient. The antenna is subjected to extreme temperatures from -125°C to $+60^{\circ}$ in Mars. This analysis is done at the worst-case temperature of $+60^{\circ}\text{C}$.

Thermal Analysis

The RF power loss is given as an input to the thermal solver. The thermal analysis also considers the thermal conductivity and the heat capacity of the Martian atmosphere at ambient temperature. At $+60^{\circ}$, the thermal conductivity and the heat capacity of the Martian atmosphere (mostly CO_2) are 0.0175 W/K/m and 0.883 kJ/K/kg . The thermal conductiv-

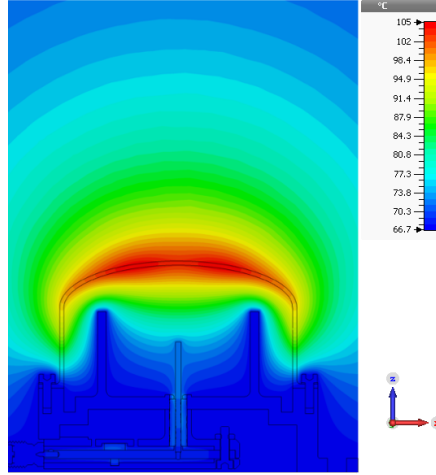


Figure 5.12: Temperature distribution across the TX antenna at $+60^{\circ}$ ambient temperature.

ity of HexPly 8552 quasi-isotropic E-Glass fibre is typically between 0.16-0.33 W/K/m [93].

The thermal analysis is done with the Thermal Steady State Solver of the CST MPHYICS Studio to allow easy coupling of EM-multiphysics simulation. The temperature distribution is shown in Fig. 5.12. The surface of the radome remains the hottest part while the majority of the antenna structure remains at the ambient temperature. Because the thermal conductivity of the fiberglass is low compared to the aluminum housing, the heat is concentrated around the radome.

Structural Analysis

For the mechanical analysis, the antenna port is fixed mechanically with a displacement boundary. Because of different coefficient of thermal expansion (CTE), the materials in the structure expand or contract at different rates over the large temperature variation. The mechanical deformations (10x) at different ambient temperatures of -125°C and $+60^{\circ}\text{C}$ are shown in Fig. 5.13. The expansion of the structure in the former and its contraction in the latter can be seen in the figure.

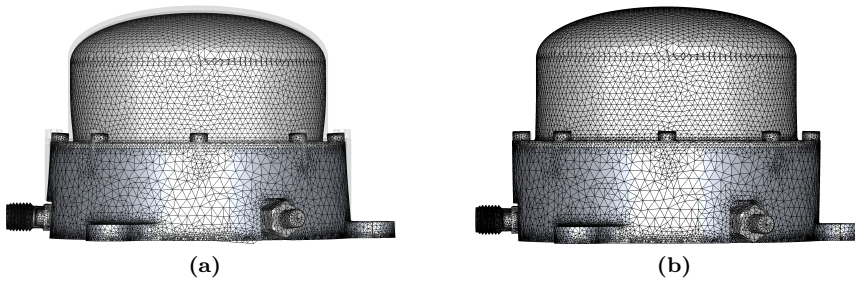


Figure 5.13: Deformed mesh of the TX antenna (10 $\times$) at (a) -125°C and (b) $+60^{\circ}\text{C}$.

Electromagnetic Analysis

The structural mechanics solver provides the deformed mesh, which is then used as an input to the electromagnetic solver for assessing the antenna RF performance. The electromagnetic simulation uses the deformed mesh exported after the structural mechanics simulation. The deformed mesh, exported in STL format, is a triangulated approximation of the structure.

Reflection Coefficient

Fig. 5.14 shows the reflection coefficient of the TX antenna at different ambient temperatures. The temperature variation is greatest on the top surface of the radome. Consequently, the mechanical deformation is largest on the top part of the radome but this is not significant enough to change the reflection coefficients of the antenna at various ambient temperatures. This shows that the design is quite robust to temperature variations.

Radiation Pattern

Fig. 5.15 compares the worst-case AR at different ambient temperatures for the TX frequency bands. The thermoelastic analysis shows minimal effect on AR. Fig. 5.16 shows the best- and worst-copolar gains at -125°C and $+60^{\circ}\text{C}$ ambient temperatures. The gains are taken at 8.4 GHz, 8.428 GHz and 8.45 GHz. In the worst-case scenario, the design is non-compliant at the edges of the mask. At $+60^{\circ}\text{C}$, the performance is close to the reference value.

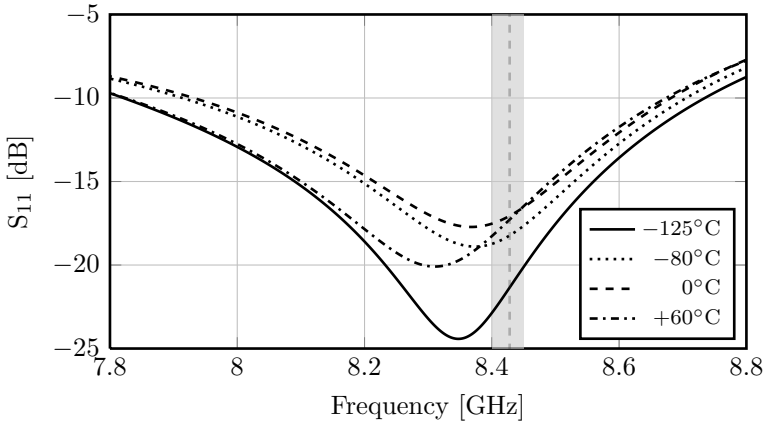


Figure 5.14: Simulated reflection coefficients of the TX antenna at different ambient temperatures.

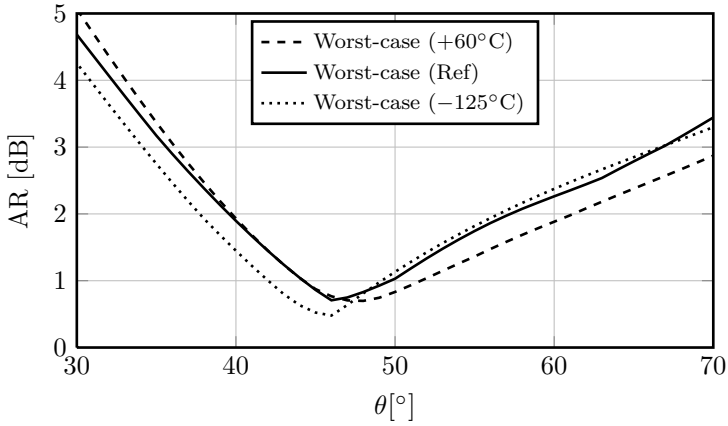


Figure 5.15: Worst-case AR at (a) -125°C and (b) $+60^{\circ}\text{C}$ ambient temperatures.

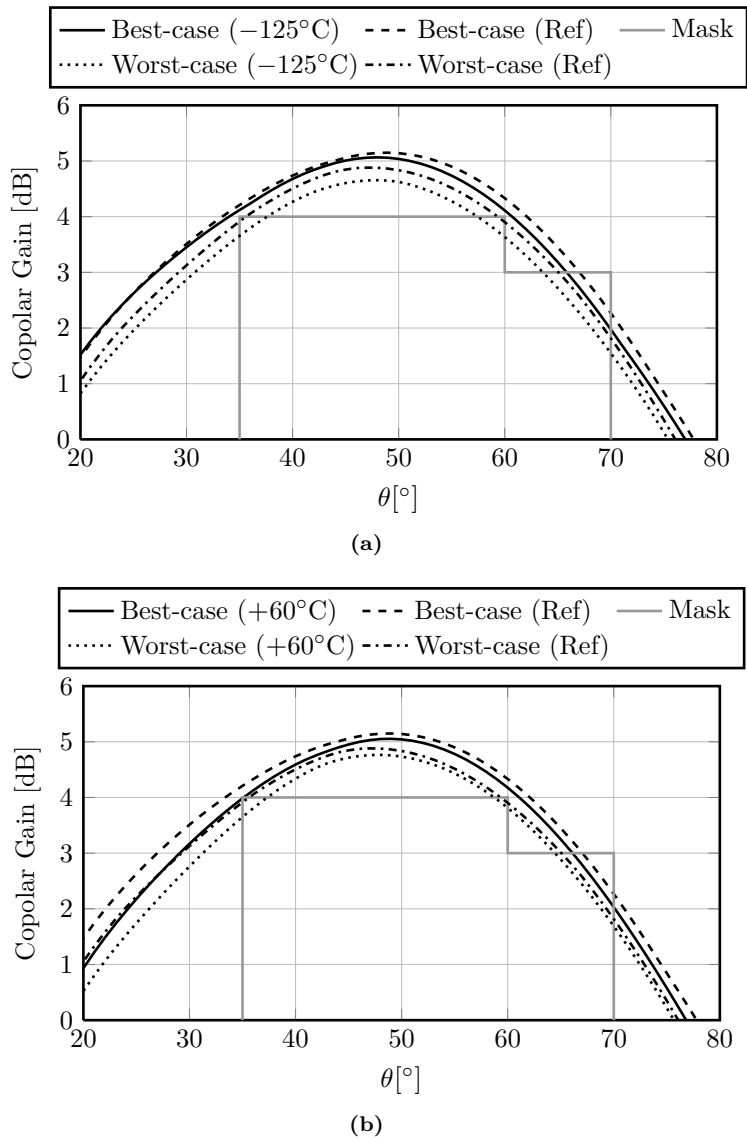


Figure 5.16: Best- and worst-case simulated copolar gains at (a) -125°C and (b) $+60^{\circ}\text{C}$ ambient temperatures.

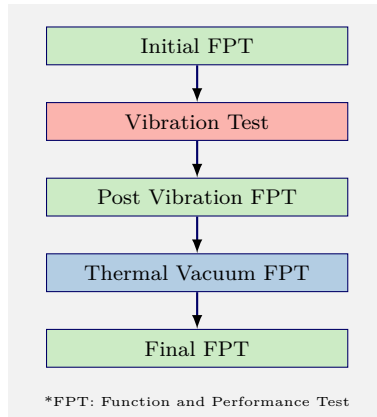


Figure 5.17: Antenna qualification test flow (From Antwerp Space).

5.4 Qualification Workflow

The LaRa antennas had to undergo through a space qualification procedure by subjecting the antennas to flight and environmental conditions. An Engineering Qualification Model (EQM) was built and qualified. The qualification test flow for the EQM models is shown in Fig. 5.17. The model approach for LaRa antenna qualification is EQM + PFM approach [94, Chapter 6]. PFM stands for Proto Flight Model.

The initial and final function and performance tests consisted of electromagnetic tests (impedance and radiation patterns). During the vibration and thermal tests, the reflection coefficient was monitored. The post vibration test simply consisted of reflection coefficient measurement. The results in Section 5.5 are the outcomes of the final function and performance tests.

5.5 Measurement Results

The LaRa antenna housing is manufactured out of a single block of aluminum using precision machining. The radome, shown in Fig. 5.18(a), is made of a fiber-glass composite material (HexPly 8552 resin system with woven glass fiber) having a thickness of 1 mm. The screws are all staked with Scotch-Weld epoxy adhesive. The pattern measurement was carried out in an anechoic chamber using SATIMO quad ridge reference horn QH800 at

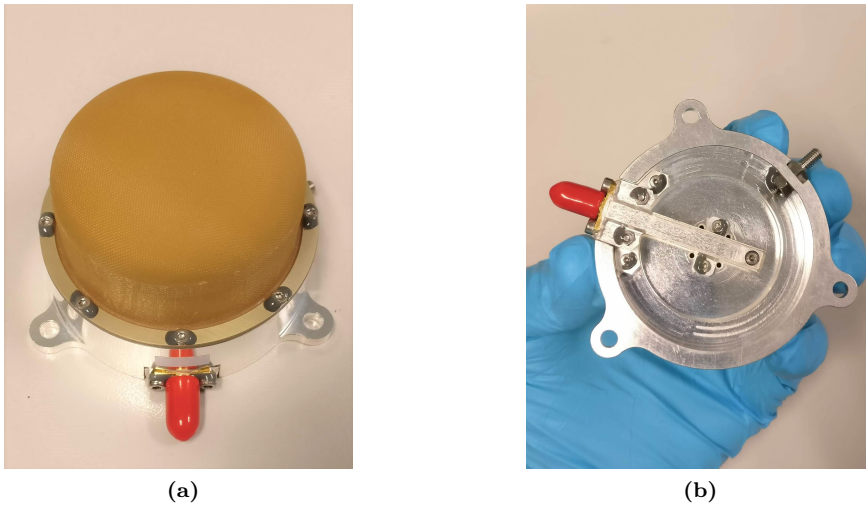


Figure 5.18: Manufactured TX LaRa antenna: (a) top view with the radome, and (b) bottom view with the square coaxial feeding line.

a distance of 3.94 m from the test antenna. The test antenna was rotated with a turntable while the reference horn was fixed.

Reflection Coefficient

The measured reflection coefficient seems to vary compared to the simulated one (see Fig. 5.19). This can be explained by the tight manufacturing tolerance on the square coaxial feeding line. Moreover, the radome has a rough surface, which could play a minor role in the change in performance.

Radiation Pattern

The RHCP copolar gain is shown in Fig. 5.20. The measured gain shows good correspondence with the simulated result, although the worst-case gain is slightly deteriorated. This could be because of using an inaccurate model in the simulation for the radome composite material, whose dielectric properties were not accurately known beforehand. A similar trend is seen in the worst-case AR plot of Fig. 5.21, where the antenna performs better in higher elevation (lower θ).

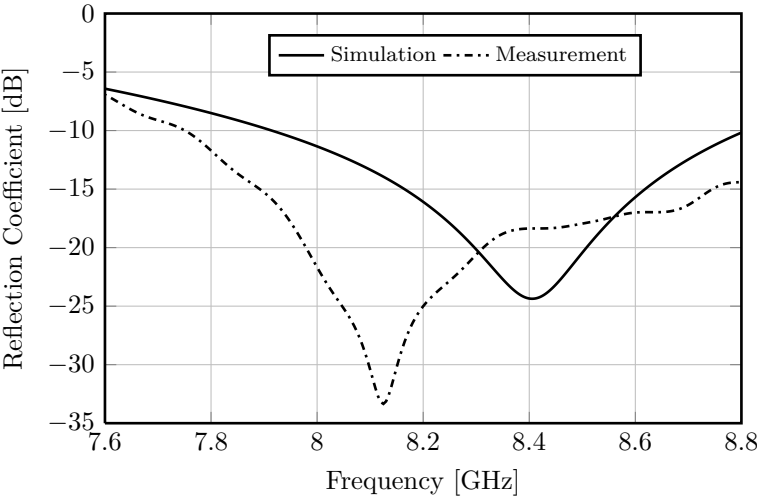


Figure 5.19: Reflection coefficient of the measured LaRa antenna.

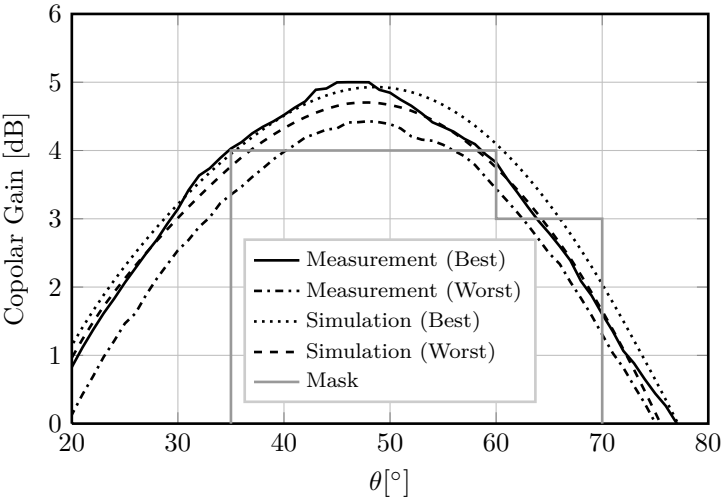


Figure 5.20: Best and worst case copolar gains of LaRa antenna at 8.428 GHz.

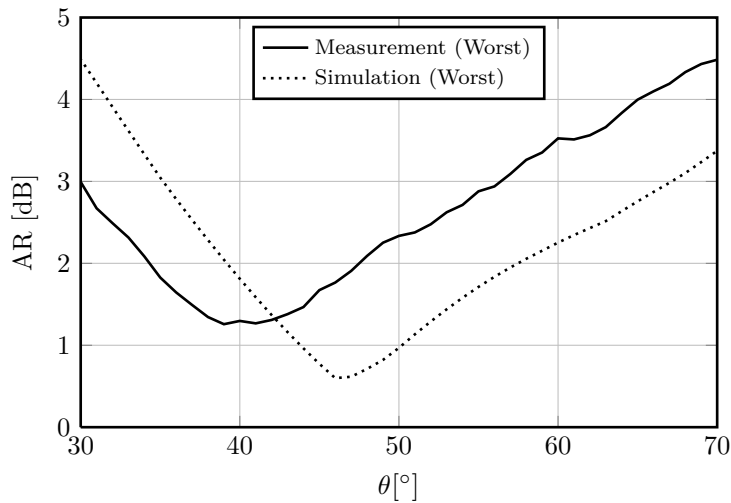


Figure 5.21: Worst case AR of LaRa antenna at 8.428 GHz.

5.6 Conclusion

Chapter 5 provided the details on the adaptation of the 3D monopole parasitic antenna for the LaRa instrument. The Martian conditions imposed additional design constraints on the antennas, and a multiphysics approach was required. Because of the all-metallic body, the thermoelastic effect could be minimized. This proves that the novel design proposed for the LaRa instrument is a robust alternative to the patch antenna originally foreseen with even superior electromagnetic performance.

CHAPTER 6

Array Analysis using Inhomogeneous Plane Waves

This chapter introduces the concepts of the hybrid Method of Moments (MoM) techniques using Inhomogeneous Plane Waves (IPW) and Macro Basis Functions (MBF). Examples from slot antennas to the frustum array antenna described in Chapter 2 validate the numerical techniques, and reinforce the idea that this hybrid technique can become competitive when compared to other fast techniques, such as the Fast Multipole Method (FMM).

Section 6.1 introduces the MoM numerical method applied to solving the Maxwell's equations for antenna radiation conditions. Furthermore, it describes the MBF techniques employed to reduce the solution time of the MoM matrix.

Section 6.2 expands on the hybrid MBF-IPW technique, and its validation with some numerical examples. The acceleration is achieved as a result of the separability of the Green's function.

6.1 Introduction

For antenna radiation and scattering problems in electromagnetism, the time-varying fields and their inter-relationships are of utmost importance. Michael Faraday postulated that the electric field is produced by a changing magnetic field, and James Clerk Maxwell provided a theoretical basis that the magnetic field is produced by a changing electric field. Maxwell unified a set of laws already in existence (Gauss's Law, Faraday's Law, Ampere's Law), removing their asymmetric inconsistencies and formulating what became known as the Maxwell's equations [95, Section 7-3]. The four Maxwell's equations in time-harmonic form (fields written as phasors) are given below:

$$\nabla \times \mathbf{E} = -j\omega\mu\mathbf{H} \quad (6.1a)$$

$$\nabla \times \mathbf{H} = \mathbf{J} + j\omega\epsilon\mathbf{E} \quad (6.1b)$$

$$\nabla \cdot \mathbf{E} = \frac{1}{\epsilon}\rho \quad (6.1c)$$

$$\nabla \cdot \mathbf{H} = 0 \quad (6.1d)$$

The first law, Faraday's law (6.1a), stipulates that a changing magnetic field gives rise to a circulating electric field. In nature, however, the change in magnetic field is not linear, but is rather oscillating. This oscillating magnetic field gives rise to the electric field, which itself is constantly changing over time. Now, from Ampere's law (6.1b), a changing electric field again gives rise to a circulating magnetic field, such that this process continues and sustains the propagation of electromagnetic waves.

The electric field at the observation point $\mathbf{r}$ due to electric charge at source point $\mathbf{r}'$ in the integral form is written as [96, Section 2.1],

$$\mathbf{E}(\mathbf{r}) = -j\omega\mu \iint_{S'} G(R)\mathbf{J}(\mathbf{r}') + \frac{1}{k^2} \nabla \nabla \cdot \iint_{S'} G(R)\mathbf{J}(\mathbf{r}') d\mathbf{r}' \quad (6.2)$$

where $G(R)$ or the scalar Green's function for free space in 3D is,

$$G(R) = G(|\mathbf{r} - \mathbf{r}'|) = \frac{e^{-jk|\mathbf{r} - \mathbf{r}'|}}{4\pi|\mathbf{r} - \mathbf{r}'|} \quad (6.3)$$

6.1.1 The Method of Moments

The classical book on the Method of Moments (MoM) by Harrington [97] remains to this day the best reference. The MoM is a numerical technique where the integral equations of the form (6.2) can be solved in the form of linear equations using high-speed digital computers [96].

The MoM solves linear equations of the form [97, Section 1.3]

$$L(f) = g \quad (6.4)$$

where L is a linear operator, g is the source or excitation, and f is the field or response to be determined.

The unknown function f is expanded into known functions defined by $f_1, f_2, f_3 \dots$ in the domain of L ,

$$f = \sum_n \alpha_n f_n \quad (6.5)$$

where α_n are constants, and f_n are called the *basis functions*.

Substituting (6.5) into (6.4), the linear operator L yields

$$\sum_n \alpha_n L(f_n) = g \quad (6.6)$$

Consider an inner product or *moment* between a basis function $f_n(\mathbf{r}')$ and a *testing or weighting function* $w_m(\mathbf{r})$ as

$$\langle w_m, f_n \rangle = \int_{w_m} w_m(\mathbf{r}) \cdot \int_{f_n} f_n(\mathbf{r}') d\mathbf{r}' d\mathbf{r} \quad (6.7)$$

Taking the inner product of (6.6) with the testing functions w_m , gives

$$\sum_n \alpha_n \langle w_m, L(f_n) \rangle = \langle w_m, g \rangle \quad (6.8)$$

This can be written in $N \times N$ matrix form as

$$[Z_{mn}][\alpha_n] = [g_m] \quad (6.9)$$

where,

$$[Z_{mn}] = \begin{bmatrix} \langle w_1, L(f_1) \rangle & \langle w_1, L(f_2) \rangle & \cdots \\ \langle w_2, L(f_1) \rangle & \langle w_2, L(f_2) \rangle & \cdots \\ \vdots & \vdots & \ddots \end{bmatrix} \quad (6.10)$$

$$[\alpha_n] = \begin{bmatrix} \alpha_1 \\ \alpha_1 \\ \vdots \end{bmatrix} \quad [g_m] = \begin{bmatrix} \langle w_1, g \rangle \\ \langle w_2, g \rangle \\ \vdots \end{bmatrix} \quad (6.11)$$

The unknown constants α_n can simply be calculated by matrix inversion of the non-singular Z_{mn} , such that

$$[\alpha_n] = [Z_{mn}]^{-1} [g_m] \quad (6.12)$$

Subsequently, the unknown function f can be computed from (6.5). The choice of $\mathbf{f}_n$ and $\mathbf{w}_m$ plays an important role in the convergence of the MoM matrix. In this work, the set of testing functions is chosen to be the same as the set of basis functions, which is known as the Galerkin's method [98].

Discretization and Basis Functions

To solve the antenna radiation problems using the MoM technique, the antenna structure needs to be discretized and represented with appropriate basis functions. Two basis functions are of particular interest given the ease with which they can be defined on a mesh produced by a CAD software: rooftop basis functions and Rao-Wilton-Glisson (RWG) [99] basis functions. Their geometries are given in Figs. 6.1(a) and 6.1(b).

Normally, the rooftop basis function is used to represent the feeding node, since the current can be approximated with better precision [100, Section 2.2]. Once the geometry is created, the discretization and the mesh generation is done using Gmsh [101] mesh generator.

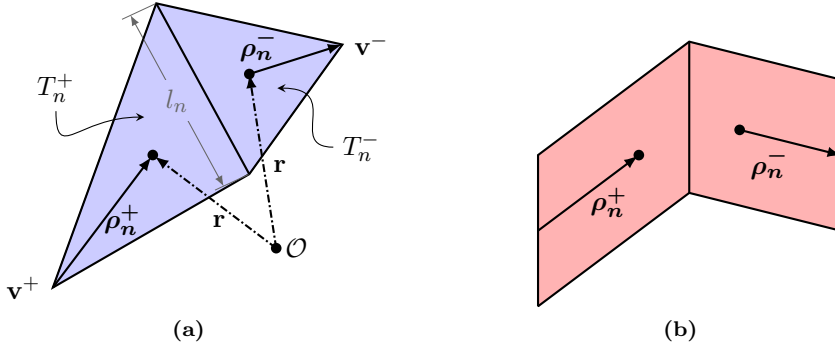


Figure 6.1: Geometry of the (a) RWG basis function, and the (b) rooftop basis function.

The RWG basis functions are defined as

$$\mathbf{f}_n(\mathbf{r}) = \begin{cases} \frac{l_n}{2A_n^+} \rho_n^+, & \mathbf{r} \text{ in } T_n^+ \\ \frac{l_n}{2A_n^-} \rho_n^-, & \mathbf{r} \text{ in } T_n^- \\ 0, & \text{otherwise} \end{cases} \quad (6.13)$$

where, l_n is the length of the common edge, $A_n^\pm$ is the area of the triangular facet.

The vector ρ_n^+ in T_n^+ points away from the vertex $\mathbf{v}^+$, and is given as

$$\rho_n^+ = \mathbf{v}^+ - \mathbf{r}, \text{ when } \mathbf{r} \text{ in } T_n^+ \quad (6.14)$$

Similarly, ρ_n^- points toward the vertex $\mathbf{v}^-$, and is given as

$$\rho_n^- = \mathbf{r} - \mathbf{v}^-, \text{ when } \mathbf{r} \text{ in } T_n^- \quad (6.15)$$

Only the internal edges of the RWG basis function have a normal component, such that the boundary edges do not have any charges. The surface

divergence of the RWG basis function is, thus, given as

$$\nabla_s \cdot \mathbf{f}_n(\mathbf{r}) = \begin{cases} \frac{l_n}{A_n^+}, & \mathbf{r} \text{ in } T_n^+ \\ -\frac{l_n}{A_n^-}, & \mathbf{r} \text{ in } T_n^- \\ 0, & \text{otherwise} \end{cases} \quad (6.16)$$

Electric Field Integral Equation

As an extension to (6.5), the currents $\mathbf{J}(\mathbf{r})$ on the RWG basis functions are written as

$$\mathbf{J}(\mathbf{r}) = \sum_n \alpha_n \mathbf{f}_n(\mathbf{r}) \quad (6.17)$$

where $\mathbf{f}_n(\mathbf{r})$ is given by (6.13).

Applying the boundary condition $\hat{n} \times \mathbf{E}_s(\mathbf{r}) = -\hat{n} \times \mathbf{E}_i(\mathbf{r})$, the scattered or the radiated electric field of (6.2) can be written as

$$\frac{1}{j\omega\epsilon} \hat{n} \times \left[k^2 \bar{I} + \nabla \nabla \cdot \right] \iint_{S'} G(R) \mathbf{J}(\mathbf{r}') d\mathbf{r}' = -\hat{n} \times \mathbf{E}_i(\mathbf{r}) \quad (6.18)$$

Equation (6.18) is also known as the *Electric Field Integral Equation* (EFIE).

Substituting (6.17) in (6.18) for *all* basis functions ($n = 1, 2, \dots, N$), and using *Galerkin*¹ testing [100, Section 2.2]

$$\begin{aligned} \frac{1}{j\omega\epsilon} \sum_n \alpha_n \int_S \mathbf{J}_m(\mathbf{r}) \cdot \left[k^2 \bar{I} + \nabla \nabla \cdot \right] \int_{S'} \mathbf{J}_n(\mathbf{r}') G(R) d\mathbf{r}' d\mathbf{r} = \\ - \int_S \mathbf{J}_m(\mathbf{r}) \cdot \mathbf{E}_i(\mathbf{r}) d\mathbf{r} \end{aligned} \quad (6.19)$$

Therefore, the MoM impedance matrix is given as

$$\mathbf{Z}_{mn} = \frac{1}{j\omega\epsilon} \int_S \mathbf{J}_m(\mathbf{r}) \cdot \left[k^2 \bar{I} + \nabla \nabla \cdot \right] \int_{S'} \mathbf{J}_n(\mathbf{r}') G(R) d\mathbf{r}' d\mathbf{r} \quad (6.20)$$

¹If $\mathbf{J}_n$ and $\mathbf{J}_m$ represent the basis and testing functions, respectively, *Galerkin* testing gives $\mathbf{J}_n = \mathbf{J}_m$.

Further expanding the $\nabla \nabla \cdot$ operator [100, Appendix B], and suppressing $\mathbf{r}$ and $\mathbf{r}'$ for brevity

$$\mathbf{Z}_{mn} = \frac{1}{j\omega\epsilon} \int_S \int_{S'} [k^2 \mathbf{J}_m \cdot \mathbf{J}_n - (\nabla \cdot \mathbf{J}_m) (\nabla \cdot \mathbf{J}_n)] G(R) d\mathbf{r}' d\mathbf{r} \quad (6.21)$$

6.1.2 Macro Basis Functions

The Macro Basis Function (MBF) technique belongs to a class of domain-decomposition methods [15], [16], [17] used to solve large radiation and scattering electromagnetic problems. With the MBF technique, it is possible to significantly reduce the size of the MoM matrix, such that the solution of the linear equation (6.12) becomes memory efficient and less computationally intensive.

In domain-decomposition methods, the elementary basis functions are grouped into higher-level subdomains. The MBF technique then postulates that a solution for each subdomain can be found as a linear combination of pre-computed subdomain solutions [102].

In its very basic form, the MBF of an isolated subdomain can be represented by a column vector $\mathbf{q}_{ik}$, where i represents the index of the subdomain and k represents the index of the MBF in that subdomain. From (6.9), the MoM system of equations in matrix form can be written as

$$\mathbf{Z} \boldsymbol{\alpha} = \mathbf{g} \quad (6.22)$$

If $\mathbf{Q}_i$ represents the matrix containing all the MBFs $\mathbf{q}_{ik}$, such that $\mathbf{Q}_i^H$ represents its Hermitian transpose, then (6.22) can be rewritten as

$$\begin{aligned} \mathbf{Q}_i^H \mathbf{Z} \boldsymbol{\alpha} &= \mathbf{Q}_i^H \mathbf{g} \triangleq \mathbf{g}_{red} \\ \mathbf{Q}_i^H \mathbf{Z} \mathbf{Q}_i \mathbf{y}_i &= \mathbf{Q}_i^H \mathbf{g} \end{aligned} \quad (6.23)$$

where $\mathbf{Q}_i \mathbf{y}_i = \boldsymbol{\alpha}$.

Now, the reduced MoM matrix $\mathbf{Z}_{red}$ can be written as

$$\mathbf{Z}_{red} = \mathbf{Q}_i^H \mathbf{Z} \mathbf{Q}_i \quad (6.24)$$

The reduction depends on the size of subdomains (or the size of $\mathbf{Q}_i$). The reduced MoM system of equation is solved as follows

$$\mathbf{Z}_{red} \mathbf{y}_i = \mathbf{g}_{red} \quad (6.25)$$

$$\mathbf{y}_i = [\mathbf{Z}_{red}]^{-1} \mathbf{g}_{red} \quad (6.26)$$

$$\boldsymbol{\alpha} = \mathbf{Q}_i \mathbf{y}_i \quad (6.27)$$

Now, the selection or the generation of the MBFs is an important aspect of this technique. In this work, multiple-scattering MBFs [103], [104] are used to generate primary and secondary MBFs (tertiary MBFs are also possible). The primary MBF is generated from excitation in a particular subdomain, whereas the secondary MBFs are the currents excited on the neighboring subdomains as a result of exciting the subdomain containing the primary MBF [105, Section 3.3]. Section 6.2.2 gives the details on the generation of these MBFs.

6.2 Macro Basis Function based Inhomogeneous Plane Waves

Methods based on the multipole decompositions [14] have led to a revolution in terms of efficiency in the field of computational electromagnetics. Improvements on multipoles or alternatives are now being investigated, among others to try to avoid the near-field limitations. Many attempts concern domain-decomposition techniques. In that context, the complex patterns of the different subdomains can be used for a direct computation of the reduced interaction matrix using a set of plane waves propagating along their centers. The acceleration is achieved by grouping and representing the basis functions in the subdomains as Macro Basis Functions (MBFs) and Macro Testing Functions (MTFs) and computing their Fourier Transforms.

The Fast Inhomogeneous Plane Wave Algorithm (FIPWA) [20] provides a plane-wave decomposition of the Green's function, where the latter is integrated in spectral domain, following a special contour, such as the Steepest Descent Path (SDP). An important distinction of this method is that the number of plane waves needed decreases with distance between the subdomains. However, the FIPWA also requires the decomposition of specific patterns for each subdomain pair, which has been remedied in the 2D case, starting from a harmonic decomposition of the real patterns [106]. Martini [107] and Alkhalifeh [108] provided a new sampling approach, which exploits a systematic compensation between aliasing and truncation errors.

6.2.1 Separability of the Green's Function

The complex plane-wave decomposition of the 3D Green's function in terms of wavenumbers (k_x, k_y, k_z) is given as [20] [106]

$$G(R) = \frac{1}{(2\pi)^2} \iint \frac{e^{-jk_x(x-x')} e^{-jk_y(y-y')} e^{-jk_z|z-z'|}}{2jk_z} dk_x dk_y \quad (6.28)$$

for observation and source points $\mathbf{r} = (x, y, z)$ and $\mathbf{r}' = (x', y', z')$, respectively such that $R = |\mathbf{r} - \mathbf{r}'|$.

The free-space wavenumber k is defined as

$$k^2 = k_x^2 + k_y^2 + k_z^2 = \omega^2 \epsilon \mu \quad (6.29)$$

In the spectral domain, the Fourier transform (FT) of the basis function can be written as

$$\tilde{\mathbf{F}}_{\mathbf{b}}(k_x, k_y) = \int \mathbf{J}_{\mathbf{b}}(x', y', z') e^{j(k_x x' + k_y y' + k_z z')} d\mathbf{r}' \quad (6.30)$$

The groups of basis and testing functions (or the basis and testing subdomains) are rotated in such a way that the line connecting their centers is aligned along the z axis, and their centers are shifted to a global center $(0,0,0)$ as can be seen in Fig. 6.2.

From the classic Method of Moments relation [97], and using (6.28)

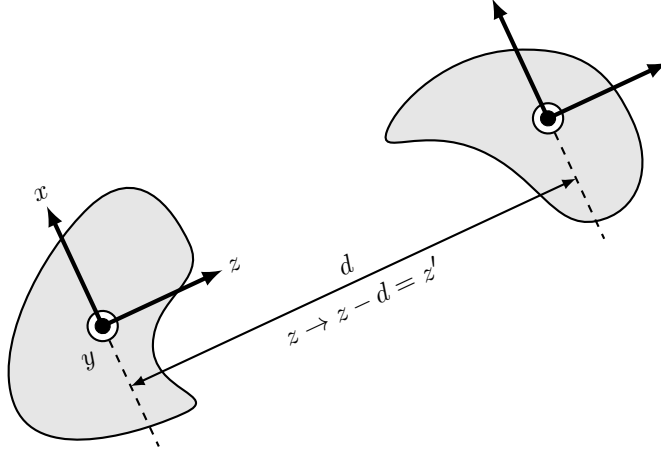


Figure 6.2: Change of coordinates for the treatment of the Green's function.

and (6.30), the interaction between the basis and testing function is

$$\mathbf{Z} = -\frac{jk\eta}{(2\pi)^2} \iint \frac{\tilde{\mathbf{F}}_{\mathbf{b}} \tilde{\mathbf{F}}_{\mathbf{t}}^{**} e^{-jk_z d}}{2jk_z} dk_x dk_y \quad (6.31)$$

where, η is the free-space impedance, and $\tilde{\mathbf{F}}_{\mathbf{t}}^{**}$ is the Fourier Transform of the testing function defined as

$$\tilde{\mathbf{F}}_{\mathbf{t}}^{**}(k_x, k_y) = \int \mathbf{J}_{\mathbf{t}}^*(x', y', z') e^{-j(k_x x' + k_y y' + k_z z')} d\mathbf{r}' \quad (6.32)$$

The spectral integration follows a special contour such as the SDP to avoid a singular integrand [109]. Equation (6.31) can be rewritten as

$$\mathbf{Z} = -\frac{jk\eta}{(2\pi)^2} \int_{\beta} \int_{\phi} \frac{\tilde{\mathbf{F}}_{\mathbf{b}} \tilde{\mathbf{F}}_{\mathbf{t}}^{**} e^{-jk_z d}}{2jk_z} d\phi d\beta \quad (6.33)$$

where, β is a complex quantity written as $\beta = \beta_r + j\beta_i$ (β_r, β_i describe the chosen integration contour) and satisfies $\beta^2 = k_x^2 + k_y^2$, and ϕ is the azimuthal angle over the (k_x, k_y) plane.

Applying the change of coordinates [109] in (6.33) gives

$$\mathbf{Z} = -\frac{jk\eta}{(2\pi)^2} \iint \tilde{\mathbf{F}}_{\mathbf{b}} \tilde{\mathbf{F}}_{\mathbf{t}}^{**} H(k_x, k_y) dk_x^r dk_y^r \quad (6.34)$$

where the exponents $(^r)$ denote the real part, and

$$H(k_x, k_y) = \frac{e^{-jk_z d}}{2jk_z} \left(1 + j\frac{\beta_i}{\beta_r}\right) \left(1 + j\frac{d\beta_i}{d\beta_r}\right) \quad (6.35)$$

6.2.2 Spectral MBF Approach

While the filling of the MoM impedance matrix $\mathbf{Z}$ is accelerated with the use of FIPWA, solving this matrix of the form $\mathbf{Z}\mathbf{x} = \mathbf{v}$ for current $\mathbf{x}$ and excitation $\mathbf{v}$ is done using the MBF approach [15] as it allows fast direct inversion of the MoM matrix to compute the current coefficients.

MBF Generation

The MBFs are generated using the multiple scattering MBF approach [104]. The primary MBF is obtained by exciting one of the subdomains and its neighboring subdomains, and the secondary MBF is obtained by considering the subdomain in the vicinity of the primary subdomain and its auxiliary subdomains, such that the secondary MBFs are also unique. Tertiary MBFs can also be considered for higher accuracy. The MBFs in each subdomain is then selected based on the singular value decomposition of the orthonormalized set of MBFs.

For electrically connected subdomains, preconditioners such as shielded block preconditioners [104] are used based on the strongest interactions. If $\tilde{\mathbf{Z}}_{i,i}$ and $\tilde{\mathbf{Z}}_{i,i^a}$ respectively represent the preconditioned block of self and mutual impedance in the MoM matrix for a subdomain S_i and its auxiliary subdomains S_{i^a} , the primary MBFs $\mathbf{q}_i$ and $(\mathbf{q}_i)^a$ when S_i is excited are computed as

$$\mathbf{q}_i = [\tilde{\mathbf{Z}}_{i,i}]^{-1} w_i, \quad \text{and} \quad (\mathbf{q}_i)^a = [\tilde{\mathbf{Z}}_{i,i^a}]^{-1} w_{i^a} \quad (6.36)$$

where w_i is the solutions obtained for subdomain i .

For $j \neq i$ and $j \neq i^a$, the secondary MBFs $\mathbf{q}_j$ and $(\mathbf{q}_j)^a$ are

$$\mathbf{q}_j = -[\tilde{\mathbf{Z}}_{j,j}]^{-1} \tilde{\mathbf{Z}}_{j,i} \mathbf{q}_i \quad (6.37)$$

$$(\mathbf{q}_j)^a = -[\tilde{\mathbf{Z}}_{j,j}]^{-1} \tilde{\mathbf{Z}}_{j,i^a} \mathbf{q}_{i^a} \quad (6.38)$$

The full set of MBFs, $\mathbf{Q}$, combines the vectors given by (6.36), (6.37) and (6.38). Therefore,

$$\mathbf{Q} = [\mathbf{q}_i (\mathbf{q}_i)^a \mathbf{q}_j (\mathbf{q}_j)^a] \quad (6.39)$$

The reduced interaction matrix $\mathbf{Z}'_{ij}$ between subdomains i and j can be obtained as

$$\mathbf{Z}'_{i,j} = \mathbf{Q}_i^H \tilde{\mathbf{Z}}_{i,j} \mathbf{Q}_j \quad (6.40)$$

The reduced excitation vector $\mathbf{v}'_i$ and current $\mathbf{x}'_i$ then becomes:

$$\mathbf{v}'_i = \mathbf{Q}_i^H \mathbf{v}_i, \quad \text{and} \quad \mathbf{x}' = [\tilde{\mathbf{Z}}']^{-1} \mathbf{v}'$$

Finally, the full current is calculated as

$$\mathbf{x}_i = \mathbf{Q}_i \mathbf{x}'_i$$

Spectral MBF Formulation

The spectral-domain MoM formulation given in (6.34) holds even for the MBFs and MTFs. For each elementary basis function represented by j in an MBF, its Fourier transform is computed from that of the elementary basis functions as

$$\tilde{\mathbf{F}}_{MBF} = \sum_j \mathbf{q}_j \tilde{\mathbf{F}}_j \quad (6.41)$$

Using (6.34), (6.40), and (6.41)

$$\mathbf{Z}' = -\frac{jk\eta}{(2\pi)^2} \iint \tilde{\mathbf{F}}_{MBF} \tilde{\mathbf{F}}_{MTF}^{**} H dk_x^r dk_y^r \quad (6.42)$$

which allows a dramatic acceleration of the interactions and renders (6.40) obsolete if the Fourier Transforms of the MBFs are known. This may be viewed as an extension to the spectral plane approach of Tiberi [110].

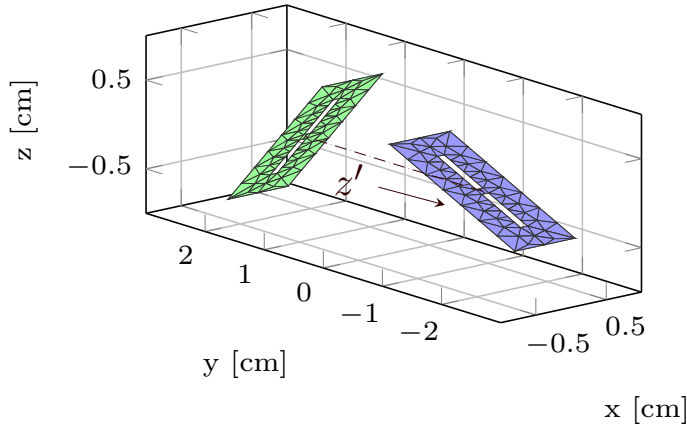


Figure 6.3: Slot antennas at 1λ separation.

6.3 Comparisons and Numerical Results

Two examples are shown to validate the proposed method: a pair of tilted slot antennas, and a 3D frustum array, similar to the design presented in Section 2.3.

6.3.1 Tilted Slot Antennas

Two tilted slot antennas (see Fig. 6.3) with centers varying at distances from 0.75λ to 10λ are considered at 10 GHz.

For separations of 0.75λ and 1λ , error levels of -24.59 dB and -34.63 dB were obtained for the excited port currents computed using the classical MBF and IPW-Spectral MBF methods (see Fig. 6.4). An interesting observation was made with the number of plane waves required (as shown in Fig. 6.5, taken from [111]), spanning from 15 for the closest groups to 7 for the distant ones.

6.3.2 Frustum Slot Array

Next, a 1×2 array of slot antennas placed on 12 facets of a slanted frustum designed at 10 GHz is considered for analysis. The frustum consisting of 14343 RWG basis functions is divided into 17 subdomains (see Fig. 6.6).

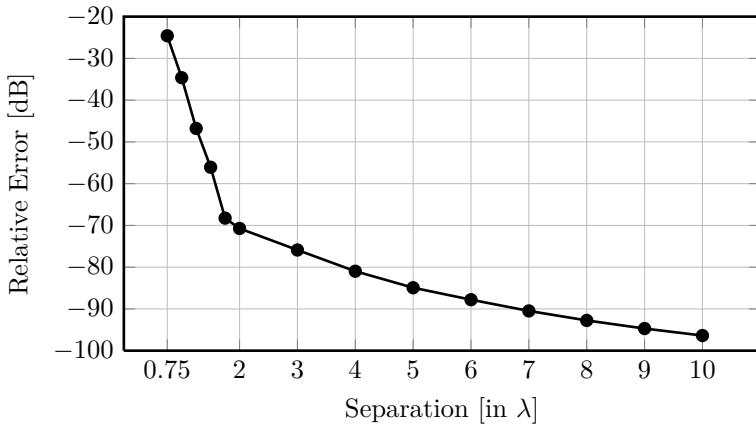


Figure 6.4: Relative errors in excited port currents between the Classical MBF and IPW-Spectral MBF methods.

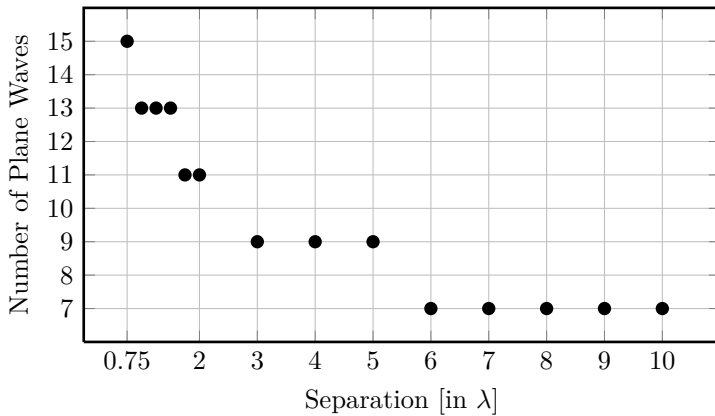


Figure 6.5: Number of plane waves required for varying separations.

Only one of the slot antennas in a subdomain is excited such that the excited subdomain and its auxiliary subdomains form the primary MBFs. The secondary MBFs are generated from the subdomains in the immediate vicinity of the subdomains used to generate the primary MBFs. The tertiary MBFs were also generated such that the total number of MBFs per subdomain was 9.

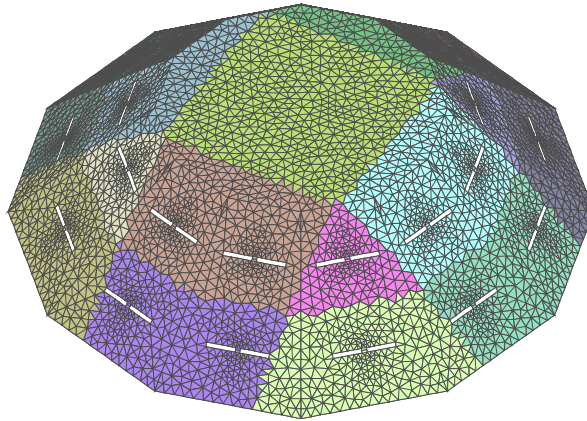


Figure 6.6: Subdomains of 3D Frustum Slot Array Antennas.

Port Currents

A slot antenna element is excited, and the current at the excited port is computed using the brute force method (BF) and the IPW-Spectral MBF method. At 10 GHz, the error level in excited port currents between these methods is -40.07 dB.

The currents at the remaining ports are also compared to see the effects of mutual coupling at 10 GHz. Fig. 6.7 shows the magnitude of port currents at all the ports. The error levels at the ports closer to the excited port are also very low. However, the error levels at the distant ports are higher, which means that a better choice of MBFs is needed to account for far interactions.

A frequency sweep from 6 to 12 GHz was also carried out and the excitation port currents compared as shown in Fig. 6.8. The error levels between these methods over the entire frequency range are consistently below -30 dB.

Computation time

The simulation was carried out in a PC with Intel(R) Xeon(R) CPU E5620 @2.40GHz and 12 GB of RAM. The computation times needed to fill up the full MoM impedance matrix using the brute force method was 30.67 mins. The self and near (closer than 0.75λ) interactions between the subdomains

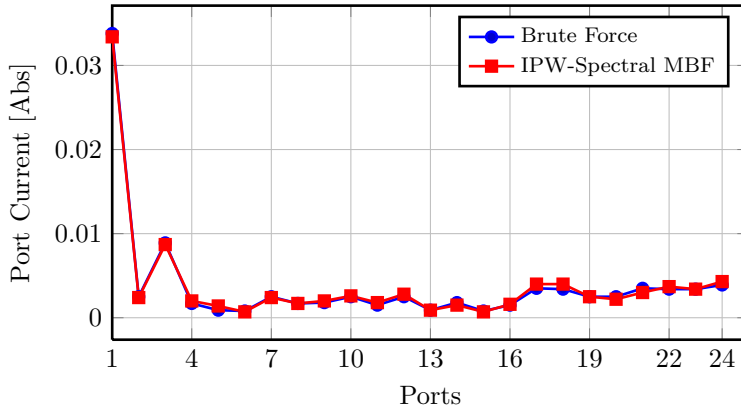


Figure 6.7: Port currents at 10 GHz computed using the brute force and the IPW Spectral-MBF methods.

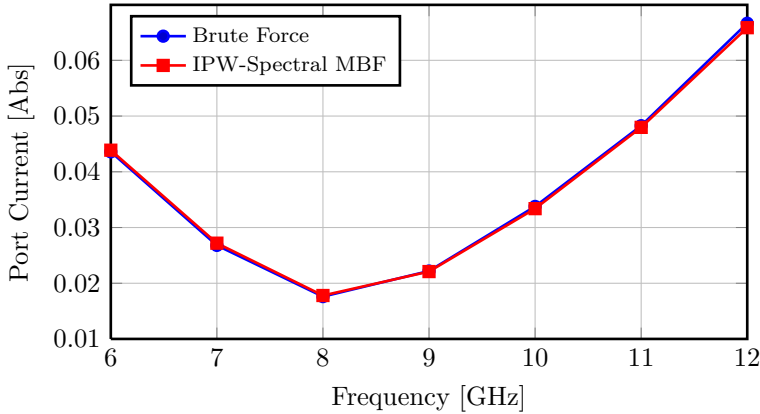


Figure 6.8: Excitation port currents from 6 to 12 GHz computed using the brute force and the IPW Spectral-MBF methods.

in the IPW method still use the brute force method. Moreover, the block of the reduced impedance matrix in such interactions is computed using the classical MBF approach as described by (6.40). The computation time required for computing the self and near interactions was 19.03 mins while the time required for computing the far interactions using the IPW method was only 4.45 mins. For 17 subdomains, there were 289 possible interactions, of which 153 interactions were self and near interactions (BF) while the

remaining 136 interactions were far interactions (IPW - MBF) (separation greater than 0.75λ). For the computation of the current coefficients, a direct inversion of the full MoM matrix required 2.9 mins while the direct inversion of the reduced MoM matrix required only 0.005 sec. With uniformly-sized and numerous subdomains, the number of far interactions can be greatly increased, and the computation time for the reduced impedance matrix can be significantly decreased.

6.4 Conclusion

A novel spectral MBF approach accelerated with the inhomogeneous plane wave method has been proposed. The number of plane waves decreases with the distance between the subdomains and only a few of them are required, even for small distances. The proposed method yields an excited port current accuracy of more than -30 dB when compared with the brute force method. However, there is a need for improvement in MBF generation and the choice of MBF to further increase the accuracy in mutual coupling analysis. A tilted slot antenna array and a 3D frustum array of slot antennas were used to validate the method.

For separations smaller than 0.7λ , the proposed method requires very large number of plane waves, and thus becomes less efficient. For such problems, a novel method is proposed that splits the Green's function into static and smooth parts. Both these parts are separable, and thus allow fast matrix-vector multiplication. This method is explored further in Chapter 7.

CHAPTER 7

Fast Interactions in Dense Mesh

This chapter deals with the near-field or the dense-mesh problems in the classical Method of Moments (MoM) approach. Fast and efficient methods exist for the case where the basis and test functions are separated by a wavelength or more, but in the near-field when the separation is less than a wavelength (or maybe half a wavelength), these fast methods become unstable. This chapter aims to provide a solution to this problem by separating the Green's function into a static part and a smooth part. This allows the integrands to be separable versus source and observation domains.

Section 7.1 introduces the concepts and builds the theoretical background to solving near-field problems. Section 7.2 emphasizes on the splitting of the Green's function. Section 7.3 expands on the static or the singular part of the Green's function and validates the concept with some examples, and Section 7.4 does so for the smooth part of the Green's function. The theoretical framework to implement the fast matrix-vector multiplication is also provided.

7.1 Introduction

Fast iterative techniques, such as the MLFMA [112], allowed the method-of-moment solution of very large problems, with dimensions of the order of tens or even hundreds of wavelengths. However, multipole decompositions are not stable in the very-near field, where traditional brute-force calculation of interactions is essential between basis functions separated by less than order one (or half) wavelength. This becomes problematic when the mesh density becomes very fine compared to the wavelength, for instance to be able to catch very fine geometrical features of the scatterer. In that case, when the scatterer is not truly large compared to the wavelength, acceleration techniques such as the MLFMA do not offer a substantial acceleration.

Dynamic multipoles [14] are known not to apply for interaction between basis and testing functions separated by less than order of half a wavelength. From a physical point of view, this is due to the fact that the far-field patterns used in such interactions do not contain sufficient near-field information. From a mathematical perspective, it is expressed through the evaluation of spherical Hankel functions for arguments that are significantly smaller than the order of the function. Cheng *et al.* [113] proposed a static multipole expansion at extremely low frequencies, but this is still an approximate solution. For higher frequencies, the authors in [113] used a combination of exponential (similar to inhomogeneous plane-wave) and dynamic multipole expansions. The directional dependence of the exponential expansion was eliminated by using a QR decomposition method in a nondirective plane wave algorithm proposed by Bogaert *et al.* [114]. Ergül and Karaosmanoğlu [115] also proposed a simple “fix” to the low-frequency problem by considering a scaling of spherical Hankel and Bessel functions. This, however, supposes two conditions to be satisfied, one of which is really difficult to satisfy in dense meshes. Hence that method requires a careful optimization of the scaling factor for difficult scenarios. Xia *et al.* [116] proposed a hybrid method which simultaneously uses the static multipole and dynamic multipole method in the different frequency regimes.

In this chapter, a novel method focusing in the low-frequency regime is proposed. The main point of this chapter consists of splitting the scalar

Green’s function into singular (static) and continuous (smooth) parts [23], as is already the case for the efficient near-field interaction between elementary basis functions [99]. The singular part or the so-called static part involves both currents and charges, and it is not totally frequency-independent (see Section 7.2). However, it can be re-written in the form of a linear combination of four interactions whose kernel is exactly $1/R$. The continuous part is very smooth and, as long as the object is not very large compared to the wavelength, its convolution with basis and testing functions can be calculated rapidly as well, without any particular difficulty in the near-field.

For distances of the order of the wavelength or smaller, the convolution with the smooth interactions are sped up using local Taylor expansions in Cartesian coordinates. Such expansions were already computed around nodes covering a regular grid for the total Green’s function in [22]. An alternative numerical technique employing the dyadic Green’s function for the smooth interactions is also explored; it ensures higher accuracy in the computation. Interactions over larger distances can be computed using standard dynamic multipoles [14].

7.1.1 The Problem Statement

The dense mesh problem can be subdivided into two main problems:

- **Low frequency breakdown:** Parasitic loop-type solution appears in the numerical result because the $\nabla\nabla\cdot$ operator in (6.21) dominates and has those loops in its null-space (so they “escape” from the verification by the integral equation). This has been partially solved by Vecchi [117] with loop-star basis functions, and is under further investigation by Andriulli [118]. This is an *accuracy* problem.
- **Large near-field zone:** When a large number of basis functions are defined within a region that is very small compared to the wavelength, traditional fast matrix-vector multiplication (e.g. dynamic multipoles) cannot be used. This calls for traditional matrix-vector product. For dense meshes the computation of the near-field part of the matrix really becomes prohibitive. This is basically a *speed* problem. This problem is addressed in this chapter.

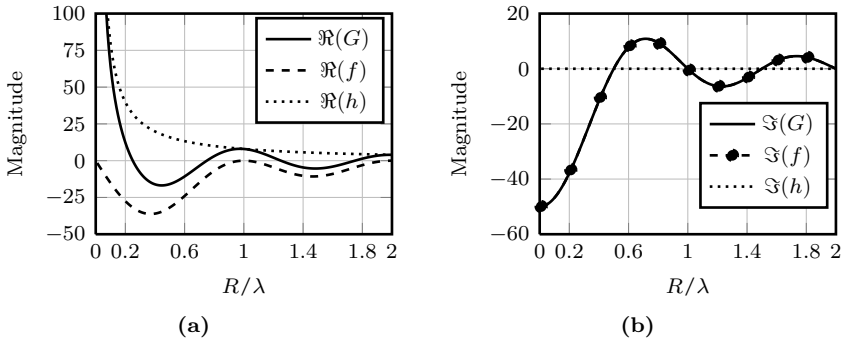


Figure 7.1: The scalar Green's function split into its static and smooth parts. (a) Real parts and (b) imaginary parts.

7.2 Splitting the Green's Function

The scalar Green's function can be split into smooth and static kernels, both of which can be written as separable functions versus source and observation points. That separability is key to most fast methods. The free space Green's function may be written as [119]

$$G(R) = \frac{e^{-jkR}}{4\pi R} = \frac{e^{-jkR} - 1}{4\pi R} + \frac{1}{4\pi R} = f(R) + h(R) \quad (7.1)$$

where $f(R)$ is smooth and continuous, while $h(R)$ is singular.

The singular part is also known as the electrostatic kernel, which can be treated with static multipoles [24].

Fig. 7.1 plots the real and imaginary parts of the Green's function split into its smooth and singular kernels. The oscillatory nature of the smooth Green's function can be clearly seen from the figure. However, it is possible to model the smooth Green's function using polynomial expansion in the near-field region where the smooth Green's function is relatively smooth. Hence, for separations less than 0.6λ , a second order Taylor series expansion is accurate enough for the model.

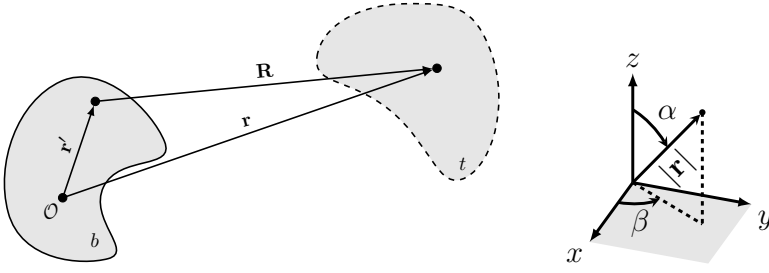


Figure 7.2: Geometrical representation in the formulation of the static Green's function.

7.3 Static Green's Function

7.3.1 Formulation

The so-called static part of the Green's function makes use of the static multipole formulation to speed up the matrix-vector product involved in the GMRES iteration. The multipole approximation to the electrostatic potential can be written as [24]

$$\frac{1}{R} = \frac{1}{|\mathbf{r} - \mathbf{r}'|} = \sum_{n=0}^{\infty} \sum_{m=-n}^n |\mathbf{r}'|^n Y_n^{-m}(\mathbf{r}') \frac{Y_n^m(\mathbf{r})}{|\mathbf{r}|^{n+1}} \quad (7.2)$$

where $Y_n^m(\mathbf{r})|\mathbf{r}|^{-n-1}$ is the spherical harmonic of degree $-n-1$.

The spherical harmonic Y is often expressed in terms of the associated Legendre polynomials, $P_n^{|m|}$. If β represents the azimuthal angle, and α represents the polar angle, then [120]

$$Y_n^m(\mathbf{r}) = (-1)^m \sqrt{\frac{(n-m)!}{(n+m)!}} P_n^{|m|}(\cos\alpha) e^{jm\beta} \quad (7.3)$$

The geometrical representation of the source and observation domains is shown in Fig. 7.2.

The static interactions are computed as a linear combination of four

interactions [see also (6.21)]:

$$\mathbf{Z}_{st} = -\frac{jk\eta}{4\pi} \int_S \int_{S'} \left[\mathbf{f}_t(\mathbf{r}) \cdot \mathbf{f}_b(\mathbf{r}') - \frac{1}{k^2} (\nabla \cdot \mathbf{f}_t) (\nabla \cdot \mathbf{f}_b) \right] \frac{1}{R} dS' dS \quad (7.4)$$

$$= \mathbf{Z}_{st}^x + \mathbf{Z}_{st}^y + \mathbf{Z}_{st}^z + \mathbf{Z}_{st}^\nabla \quad (7.5)$$

where,

$$\begin{aligned} \mathbf{Z}_{st}^x &= -\frac{jk\eta}{4\pi} \int_S \int_{S'} f_{tx}(\mathbf{r}) f_{bx}(\mathbf{r}') \frac{1}{R} dS' dS \\ &\vdots \\ \mathbf{Z}_{st}^\nabla &= +\frac{j\eta}{4\pi k} \int_S \int_{S'} (\nabla \cdot \mathbf{f}_t) (\nabla \cdot \mathbf{f}_b) \frac{1}{R} dS' dS \end{aligned}$$

Each of those terms can be treated efficiently using the static multipole formulation, described in [24] and [120].

Let us denote as $\mathbf{H}_b$ the current distribution over a given basis subdomain, itself described using elementary basis functions. Besides, $\mathbf{h}_t$ denotes the elementary testing function. Then, the corresponding matrix-vector product is given as

$$\mathbf{I} = \iint \mathbf{h}_t \mathbf{H}_b \frac{1}{R} dS' dS \quad (7.6)$$

$$= \sum_{n=0}^{\infty} \sum_{m=-n}^n \left[\int \mathbf{H}_b |\mathbf{r}'|^n Y_n^{-m}(\mathbf{r}') dS' \right] \left[\int \mathbf{h}_t \frac{Y_n^m(\mathbf{r})}{|\mathbf{r}|^{n+1}} dS \right] \quad (7.7)$$

$$\approx \sum_{n=0}^{p-1} \sum_{m=-n}^n \left[\int \mathbf{H}_b |\mathbf{r}'|^n Y_n^{-m}(\mathbf{r}') dS' \right] \left[\int \mathbf{h}_t \frac{Y_n^m(\mathbf{r})}{|\mathbf{r}|^{n+1}} dS \right] \quad (7.8)$$

The finite order p normally decreases with the ratio of the size of the basis group to the separation between the basis and test groups.

7.3.2 Complexity for Single Level

Three factors contribute to the computational complexity of this method.

- $\mathbf{M}^2$ with M number of groups
- $\mathbf{N}/\mathbf{M}$ for computation of moments of elementary basis functions on

subdomains

- **p** order, which decreases with D/R where D is the size of the subdomain and R is the distance between the subdomains. It does not depend on the wavelength.

Therefore, the total complexity of the method becomes **pMN**. In this way, $\mathcal{O}(N \log N)$ or even $\mathcal{O}(N)$ [24] can be achieved.

7.3.3 Validation

On Pairs of RWG Basis Functions in an Octree Subdivision

The static multipole is first validated on pairs of RWG basis functions. An octree division is considered, where each of the closest basis and testing cubic blocks contains one RWG basis function as shown in Fig. 7.3. The central basis block is tested by the 16 surrounding testing blocks, and the basis and each of the test blocks is separated by at least one block of same size. In a first instance, a block edge length of $\lambda/10$ and a basis function common edge length of $\lambda/64$ are considered for the analysis. A finer mesh will be considered in a subsequent example. The basis functions are randomly placed in their respective blocks.

Fig. 7.4 shows the absolute impedance value computed in brute force ($|\mathbf{Z}_{brute}|$) and the absolute error levels ($|\mathbf{Z}_{st} - \mathbf{Z}_{brute}|$) using the static multipole orders 5 and 6. Their average values are also shown in the figure. For a sparse mesh of block edge length $\lambda/10$, the average error level achieved for multipole order 5 is around -61 dB, whereas that for multipole order 6 is around -77 dB as seen in Fig. 7.4(a). Similarly, for a dense mesh of block edge length $\lambda/64$, the average error level achieved for multipole order 5 is around -57 dB, whereas that for multipole order 6 is around -73 dB as seen in Fig. 7.4(b). Notice that the absolute impedance value $|\mathbf{Z}_{brute}|$ in the case of a fine mesh has lowered compared to that in the case of coarse mesh because the RWG basis functions have also been scaled accordingly to the size of the blocks.

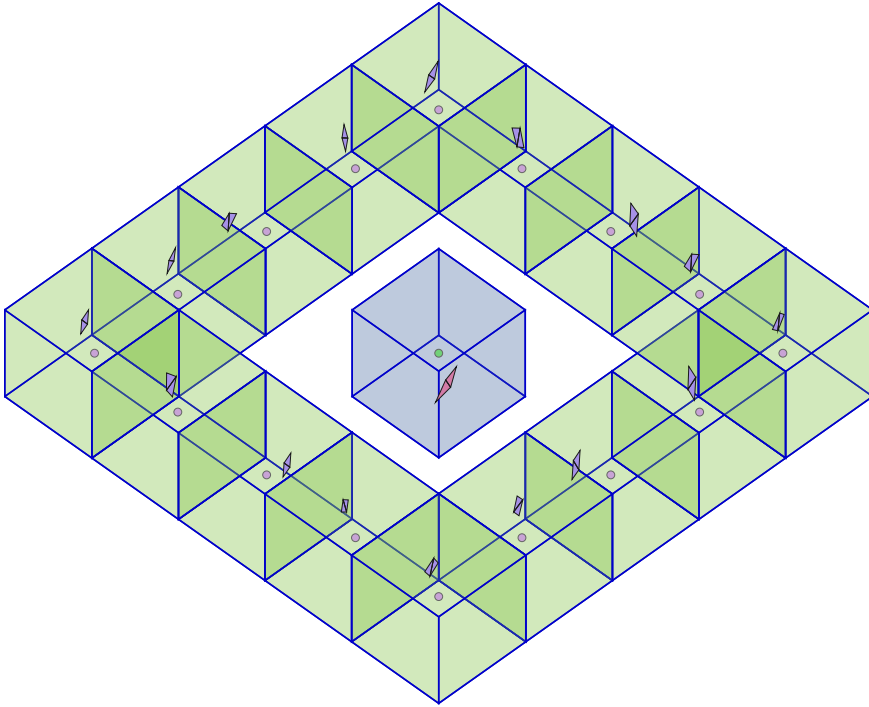


Figure 7.3: Octree subdivision for validation of the static multipole formulation.

On a Metallic Sphere

The static multipole is then validated on a metallic sphere of radius 1λ , with a mesh size of the order $\lambda/20$. The discretization is done using the RWG basis functions. The total number of RWG basis functions is 18960. For an octree level 3 subdivision, there are 272 subdomains with an average of 70 RWGs per subdomain as shown in Fig. 7.5. The order p is kept at 5 for all separations between test and basis blocks. Fig. 7.6 then shows the absolute average error levels and the difference with the brute force computation for the basis functions in all the subdomains, except the ones contiguous to the basis block (259 block interactions in total).

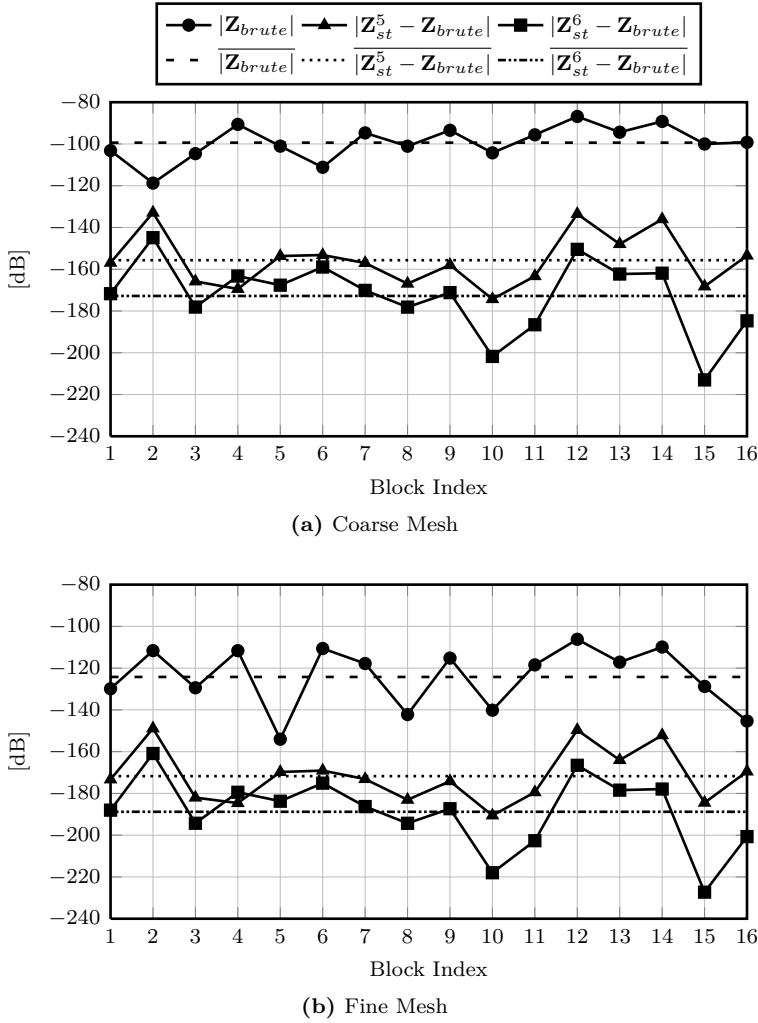


Figure 7.4: Absolute impedance values and error levels for different pairs of RWG basis functions for static multipole order 5 (Z_{st}^5) and order 6 (Z_{st}^6) (a) when using a coarse mesh (block size: $\lambda/10$, RWG size: $\lambda/64$) and (b) when using a dense mesh (block size: $\lambda/64$, RWG size: $\lambda/410$).

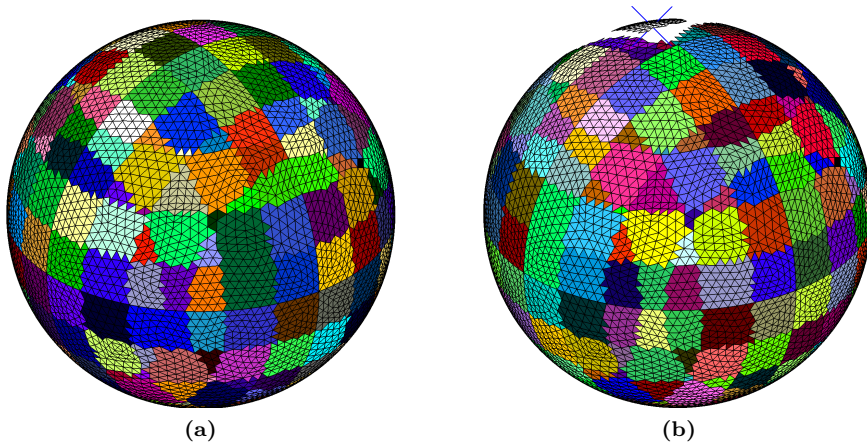


Figure 7.5: Metallic sphere (a) discretized and divided into subdomains, and (b) with remaining subdomains when tested on a subdomain.

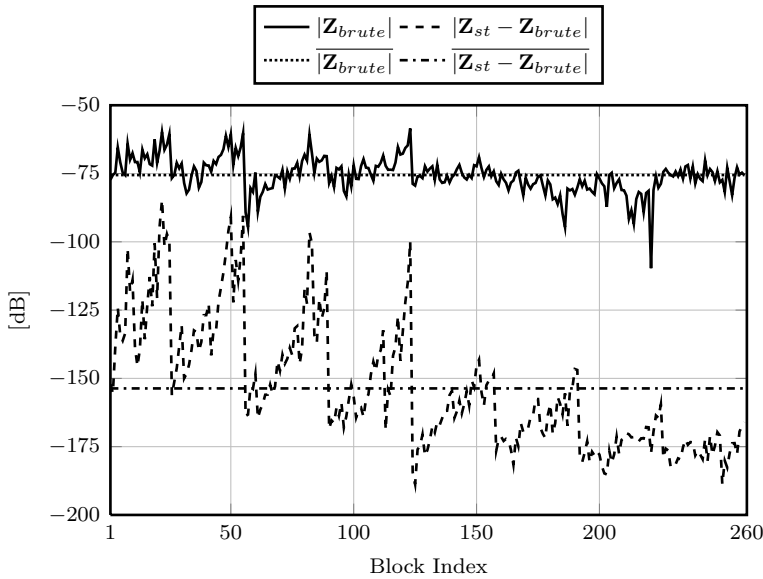


Figure 7.6: Absolute average impedance values and average error levels in different blocks of RWG basis functions in a sphere for static multipole order 5 ($\mathbf{Z}_{st}$) as compared with the brute-force solution ($\mathbf{Z}_{brute}$).

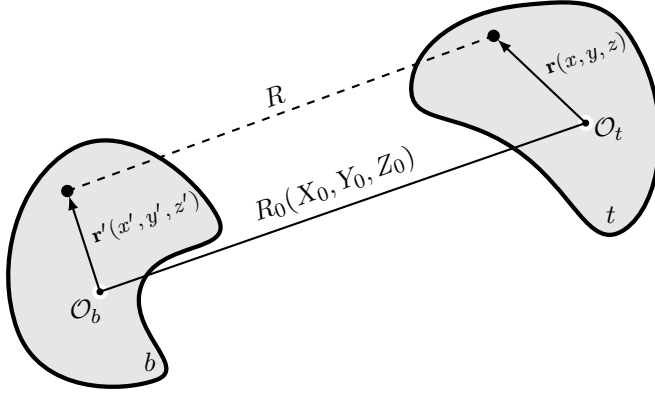


Figure 7.7: Geometrical representation in the formulation of the smooth Green's function.

7.4 Smooth Green's Function

7.4.1 Formulation with the Taylor Series Expansion

The MoM impedance matrix for the smooth Green's function is written as

$$\mathbf{Z}_{sm} = -\frac{jk\eta}{4\pi} \int_S \int_{S'} \left[\mathbf{f}_t(\mathbf{r}) \cdot \mathbf{f}_b(\mathbf{r}') - \frac{1}{k^2} (\nabla \cdot \mathbf{f}_t) (\nabla \cdot \mathbf{f}_b) \right] f(R) dS' dS \quad (7.9)$$

where $f(R) = (e^{-jkR} - 1)/(4\pi R)$, from (7.1) with the geometry shown in Fig. 7.7.

Suppose $R_0(X_0, Y_0, Z_0)$ separates the centers of the basis (b, i) and testing (t, j) groups. Separation of the basis (x', y', z') and test points (x, y, z) is given by

$$R = \sqrt{(X_0 + x - x')^2 + (Y_0 + y - y')^2 + (Z_0 + z - z')^2} \quad (7.10)$$

The Taylor series expansion of $f(R)$ yields

$$f(R) = \frac{1}{4\pi} \sum_m \alpha_m x^{rxm} y^{rym} z^{rzm} (x')^{sxm} (y')^{symb} (z')^{szm} \quad (7.11)$$

where m depends on the order of the Taylor series. This formulation is

similar to the one found in [22], where no singular-part extraction was considered.

Table 7.1 gives the analytical expansion and the coefficients of the Taylor series up to order 2.

The intermediate and partial derivatives of $f(R)$ are given as:

$$\begin{aligned}
 f'(R) &= \frac{df}{dR} = \frac{1 - e^{-jkR} - jkRe^{-jkR}}{R^2} \\
 f''(R) &= \frac{d^2f}{dR^2} = \frac{-2 + 2e^{-jkR} + 2jkRe^{-jkR} - k^2R^2e^{-jkR}}{R^3} \\
 \frac{\partial f}{\partial x} &= f'(R) \left(\frac{\partial R}{\partial x} \right); \quad \frac{\partial R}{\partial x} = \frac{X_0 + x - x'}{R} = -\frac{\partial R}{\partial x'} \\
 \frac{\partial^2 f}{\partial x^2} &= \left[f''(R) - \frac{f'(R)}{R} \right] \left(\frac{\partial R}{\partial x} \right) + \frac{f'(R)}{R} = \frac{\partial^2 f}{\partial x'^2} \\
 \frac{\partial^2 f}{\partial x \partial y} &= \left[f''(R) - \frac{f'(R)}{R} \right] \left(\frac{\partial R}{\partial x} \right) \left(\frac{\partial R}{\partial y} \right) = \frac{\partial^2 f}{\partial y \partial x} = \frac{\partial^2 f}{\partial y' \partial x'} = \frac{\partial^2 f}{\partial x' \partial y'}
 \end{aligned}$$

From (7.9) and (7.11), the MoM impedance matrix for the smooth Green's function becomes

$$\begin{aligned}
 \mathbf{Z}_{sm}^{ij} &= -\frac{jk\eta}{4\pi} \sum_m \alpha_{ij}^m \left[\iint_S \mathbf{f}_t^j x^{rxm} y^{rym} z^{rzm} dS \right. \\
 &\quad \cdot \iint_{S'} \mathbf{f}_b^i (x')^{sxm} (y')^{sym} (z')^{szm} dS' \\
 &\quad \left. - \frac{1}{k^2} \left\{ \iint_S (\nabla \cdot \mathbf{f}_t^j) x^{rxm} y^{rym} z^{rzm} dS \right. \right. \\
 &\quad \left. \left. \iint_{S'} (\nabla \cdot \mathbf{f}_b^i) (x')^{sxm} (y')^{sym} (z')^{szm} dS' \right\} \right] \quad (7.12)
 \end{aligned}$$

Equation (7.12) shows that the integrals concerning source and observation domain can be separately computed and summed, which is the key to acceleration using this method. Indeed, the integrals, which may be named *moments*, can be computed for all basis functions, and then all subdomains, in advance. The next section explains how such *moments* can be exploited to carry out fast matrix-vector products.

Table 7.1: Taylor series expansion of the smooth Green's function.

m	r_{xm}	r_{ym}	r_{zm}	s_{xm}	s_{ym}	s_{zm}	$f_m(R)$	α_m
1	0	0	0	0	0	0	α_1	$\alpha_1 = f(R_0)$
2	1	0	0	0	0	0	$\alpha_2 x$	$\alpha_2 = \frac{\partial f}{\partial x} \Big _{R_0}$
3	0	1	0	0	0	0	$\alpha_3 y$	$\alpha_3 = \frac{\partial f}{\partial y} \Big _{R_0}$
4	0	0	1	0	0	0	$\alpha_4 z$	$\alpha_4 = \frac{\partial f}{\partial z} \Big _{R_0}$
5	0	0	0	1	0	0	$\alpha_5 x'$	$\alpha_5 = -\alpha_2$
6	0	0	0	0	1	0	$\alpha_6 y'$	$\alpha_6 = -\alpha_3$
7	0	0	0	0	0	1	$\alpha_7 z'$	$\alpha_7 = -\alpha_4$
8	2	0	0	0	0	0	$\alpha_8 x^2$	$\alpha_8 = \frac{\partial^2 f}{\partial x^2} \Big _{R_0}$
9	0	2	0	0	0	0	$\alpha_9 y^2$	$\alpha_9 = \frac{\partial^2 f}{\partial y^2} \Big _{R_0}$
10	0	0	2	0	0	0	$\alpha_{10} z^2$	$\alpha_{10} = \frac{\partial^2 f}{\partial z^2} \Big _{R_0}$
11	0	0	0	2	0	0	$\alpha_{11} x'^2$	$\alpha_{11} = \alpha_8$
12	0	0	0	0	2	0	$\alpha_{12} y'^2$	$\alpha_{12} = \alpha_9$
13	0	0	0	0	0	2	$\alpha_{13} z'^2$	$\alpha_{13} = \alpha_{10}$
14	1	1	0	0	0	0	$\alpha_{14} x y$	$\alpha_{14} = \frac{\partial^2 f}{\partial x \partial y} \Big _{R_0}$
15	0	1	1	0	0	0	$\alpha_{15} y z$	$\alpha_{15} = \frac{\partial^2 f}{\partial y \partial z} \Big _{R_0}$
16	1	0	1	0	0	0	$\alpha_{16} x z$	$\alpha_{16} = \frac{\partial^2 f}{\partial x \partial z} \Big _{R_0}$
17	0	0	0	1	1	0	$\alpha_{17} x' y'$	$\alpha_{17} = \alpha_{14}$
18	0	0	0	0	1	1	$\alpha_{18} y' z'$	$\alpha_{18} = \alpha_{15}$
19	0	0	0	1	0	1	$\alpha_{19} x' z'$	$\alpha_{19} = \alpha_{16}$
20	1	0	0	1	0	0	$\alpha_{20} x x'$	$\alpha_{20} = -2\alpha_8$
21	0	1	0	0	1	0	$\alpha_{21} y y'$	$\alpha_{21} = -2\alpha_9$
22	0	0	1	0	0	1	$\alpha_{22} z z'$	$\alpha_{22} = -2\alpha_{10}$
23	1	0	0	0	1	0	$\alpha_{23} x y'$	$\alpha_{23} = -\alpha_{14}$
24	0	1	0	0	0	1	$\alpha_{24} y z'$	$\alpha_{24} = -\alpha_{15}$
25	1	0	0	0	0	1	$\alpha_{25} x z'$	$\alpha_{25} = -\alpha_{16}$
26	0	1	0	1	0	0	$\alpha_{26} x' y$	$\alpha_{26} = -\alpha_{14}$
27	0	0	1	0	1	0	$\alpha_{27} y' z$	$\alpha_{27} = -\alpha_{15}$
28	0	0	1	1	0	0	$\alpha_{28} x' z$	$\alpha_{28} = -\alpha_{16}$

7.4.2 Fast Matrix-Vector Multiplication

Equation (7.12) may be trivially decomposed into four terms. If Cartesian coordinates are used, three terms will be considered to deal with the scalar product in (7.12), while the fourth term will deal with the divergences. Consider $\mathbf{J}_j(\mathbf{r})$ as a given component of the current distribution in subdomain j .

A fast method is proposed for computing the fields radiated by that current distribution as tested by any (component of) basis function $f_t^{i,p}$ in subdomain i . Based on (7.12), such an interaction can be computed as

$$\mathbf{I}_{sm}^{(i,p)_j} = -\frac{jk\eta}{4\pi} \sum_m \alpha_{ij}^m \mathbf{M}_{i,p}^m \mathbf{M}_j^m \quad (7.13)$$

where, still for the chosen component, $\mathbf{M}_{i,p}^m$ describes the m -th moment of basis function p found in subdomain i :

$$\mathbf{M}_{i,p}^m \triangleq \iint f_b^{p,i} x^{rxm} y^{rym} z^{rzm} dS \quad (7.14)$$

while the moment for a full current distribution on subdomain j ($\mathbf{J}_j = \sum_q \mathbf{x}_{j,q} f_b^{j,q}$) is obtained by superposition:

$$\mathbf{M}_j^m \triangleq \sum_q \mathbf{x}_{j,q} \mathbf{M}_{j,q}^m \quad (7.15)$$

The separability of interaction (7.13) versus source and observation domains and the availability of all moments $\mathbf{M}_{i,p}^m$ allow the following fast matrix-vector multiplication procedure:

1. Given the vector $\mathbf{x}$ describing everywhere the current distribution, the moments $\mathbf{M}_j^m$ on all subdomains j are computed. The complexity of this step is $\mathcal{O}(MN)$ where M is the number of moments considered and N is the total number of unknowns.
2. For all groups i , compute the following aggregation of moments: $\mathbf{L}_i^m = \sum_j \alpha_{ij}^m \mathbf{M}_j^m$, which entails a sum over all source domains j as well. The complexity of this step (which parallels translation in multipole

decomposition) is of order $\mathcal{O}(MB^2)$ where B is the number of subdomain (blocks).

3. Following (7.13), the tested fields on any basis function p in subdomain i can be written as

$$\mathbf{v}^{i,p} = -\frac{jk\eta}{4\pi} \sum_m \mathbf{L}_i^m \mathbf{M}_{i,p}^m \quad (7.16)$$

The complexity of this step is also $\mathcal{O}(MN)$, as step 1.

Assuming that, as is done with multipole decompositions [14], the numbers B of subdomains and the number M of moments are both $\mathcal{O}(\sqrt{N})$, the complexity of the whole operation is $\mathcal{O}(N^{1.5})$. Actually, for extremely dense mesh, it is expected to have more than $\mathcal{O}(\sqrt{N})$ basis functions per subdomain, in which case $B \sim M \ll \sqrt{N}$ such that the global complexity would scale closer to N than to $N^{1.5}$. This better scaling essentially would arise from the fact that the wavelength no longer appears (or very marginally) in the number of needed modes, which is only commanded by the distance between subdomains, relative to the domain size.

A multi-level approach is expected to fill the gap left by the separation between source and observation domains. This is done by locally considering interactions between sub-blocks. There is effectively no limitation to the size of blocks and related minimum distance (typically 1 block size) between them. This is again possible because of the fact that the minimum distance between blocks does not need to scale with the wavelength.

7.4.3 Formulation with the Dyadic Green's Function

The term in (7.4) involving divergences may involve differences between quite large and quite close terms, which may make the formulation ill-posed. Therefore, a dyadic approach has been considered, in which derivatives apply to the Green's function rather than on the basis functions.

The compact form of the smooth part of the dyadic Green's function

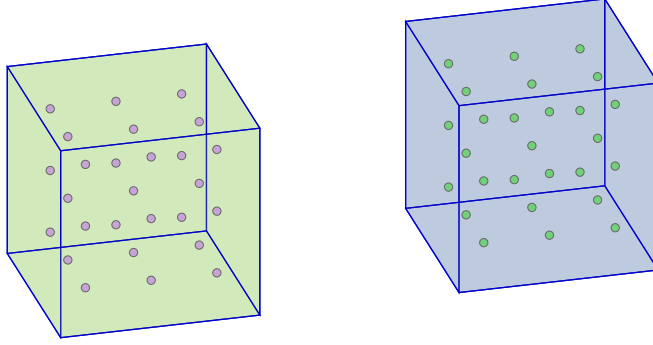


Figure 7.8: Dyadic Green's function modeling with near field blocks.

can be written as

$$\begin{aligned}\mathbf{G}(R) &= \left[\bar{\bar{I}} + \frac{1}{k^2} \nabla \nabla \cdot \right] \frac{e^{-jkR} - 1}{4\pi R} \\ &= \left[\bar{\bar{I}} + \frac{1}{k^2} \nabla \nabla \cdot \right] f(R)\end{aligned}\quad (7.17)$$

In dyadic matrix form,

$$\mathbf{G}(R) = \frac{1}{k^2} \begin{pmatrix} k^2 + \frac{\partial^2}{\partial x^2} & \frac{\partial^2}{\partial x \partial y} & \frac{\partial^2}{\partial x \partial z} \\ \frac{\partial^2}{\partial y \partial x} & k^2 + \frac{\partial^2}{\partial y^2} & \frac{\partial^2}{\partial y \partial z} \\ \frac{\partial^2}{\partial z \partial x} & \frac{\partial^2}{\partial z \partial y} & k^2 + \frac{\partial^2}{\partial z^2} \end{pmatrix} f(R) \quad (7.18)$$

The MoM impedance matrix with the dyadic Green's function simply becomes

$$\mathbf{Z}_{sm} = -\frac{jk\eta}{4\pi} \int_S \int_{S'} \mathbf{f}_t \mathbf{G} \mathbf{f}_b dS' dS \quad (7.19)$$

The Taylor series expansion of (7.11) can be directly used for each entry in (7.18). However, this would require the analytical computation of the 4th order derivatives. Instead of that, each entry of the tensor in (7.18) has been computed analytically, while its Taylor expansion versus source and observation coordinates are obtained through a numerical approximation.

Consider two near-field blocks p and q separated at most by $\lambda/2$ con-

taining numerous points inside each block (see Fig. 7.8). Prior to further computation with $\mathbf{Z}_{sm}$, the Taylor series coefficients α_m of (7.11) are obtained in a *least squares* sense.

The model to fit for Taylor order 2 is

$$f(R) = \sum_{m=1}^{28} \alpha_m \underbrace{x^{r_{xm}} y^{r_{ym}} \dots (z')^{s_{zm}}}_{f^{(m)}} \quad (7.20)$$

If $\mathbf{a} = [\alpha_1 \alpha_2 \dots \alpha_{28}]^T$ between blocks p and q , then

$$\underbrace{\begin{bmatrix} f^{(1)} & f^{(2)} & \dots & f^{(28)} \\ \vdots & \vdots & \vdots & \vdots \end{bmatrix}}_{\mathbf{F}} \mathbf{a} = \mathbf{G}_{pq} \quad (7.21)$$

where, $\mathbf{G}_{pq}$ is calculated by substituting $f(R) = (e^{-jkR} - 1)/(4\pi R)$ in (7.18) for *all* points in the p and q near-field blocks, i.e.

$$\mathbf{G}_{pq} = \frac{1}{4\pi k^2} \begin{pmatrix} k^2 + \frac{\partial^2}{\partial x^2} & \frac{\partial^2}{\partial x \partial y} & \frac{\partial^2}{\partial x \partial z} \\ \frac{\partial^2}{\partial y \partial x} & k^2 + \frac{\partial^2}{\partial y^2} & \frac{\partial^2}{\partial y \partial z} \\ \frac{\partial^2}{\partial z \partial x} & \frac{\partial^2}{\partial z \partial y} & k^2 + \frac{\partial^2}{\partial z^2} \end{pmatrix} \frac{e^{-jkR_{pq}} - 1}{R_{pq}} \quad (7.22)$$

Solving (7.21) for $\mathbf{a}$ in a least squares sense yields

$$\mathbf{a} \approx (\mathbf{F}^H \mathbf{F})^{-1} \mathbf{F}^H \mathbf{G}_{pq} \quad (7.23)$$

$$= \mathbf{F} \setminus \mathbf{G}_{pq} \quad (7.24)$$

These α coefficients are then stored and used with the actual RWG basis functions belonging to a particular (p, q) block. The fast matrix-vector multiplication described in Section 7.4.2 remains valid to the dyadic Green's function approach as well. It is important to note that, as soon as a given box size has been chosen, the α coefficients can be computed once and for all. A *lookup table* could be established for α coefficients for a few box sizes.

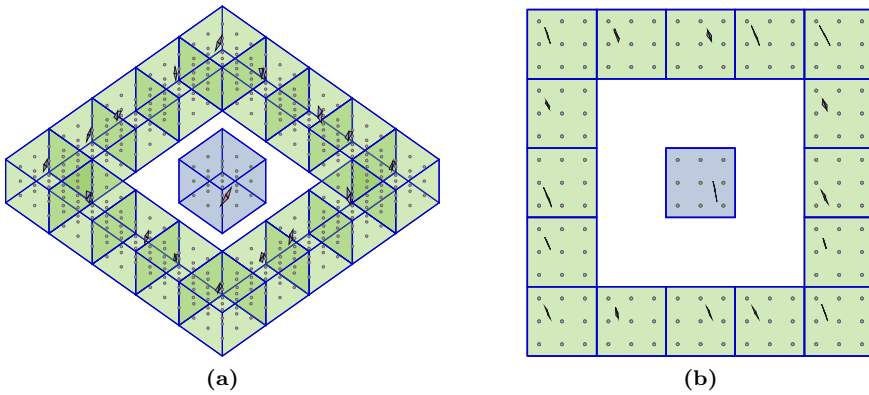


Figure 7.9: Dyadic smooth Green's function validation using test points and near field blocks in octree subdivision showing (a) a perspective view, and (b) the location of test points and basis functions from the top view.

7.4.4 Validation

The fast interaction method for the smooth part of the Green's function is validated on a pair of RWG basis functions using both the analytical expansion of the Taylor series (see Section 7.4.1) and the smooth dyadic Green's function modeling (see Section 7.4.3).

On Pairs of RWG Basis Functions in an Octree Subdivision

An example similar to Fig. 7.3, which validated the singular part, is considered for validation on the smooth dyadic Green's function model and the analytical Taylor expansion. The model example is shown in Fig. 7.9. The number of points considered (3 points each in x, y, z directions) for modeling the dyadic Green's function is also shown. The RWG basis functions are again arbitrarily placed inside each of their respective blocks.

Fig. 7.10 shows the absolute impedance value computed in brute force ($|\mathbf{Z}_{brute}|$) and the absolute error levels ($|\mathbf{Z}_{sm} - \mathbf{Z}_{brute}|$) using the dyadic Green's approach and the analytical Taylor series expansion. For a sparse mesh of block edge length $\lambda/10$, the average error level achieved for dyadic Green's approach is around -45 dB, whereas that for Taylor series expansion

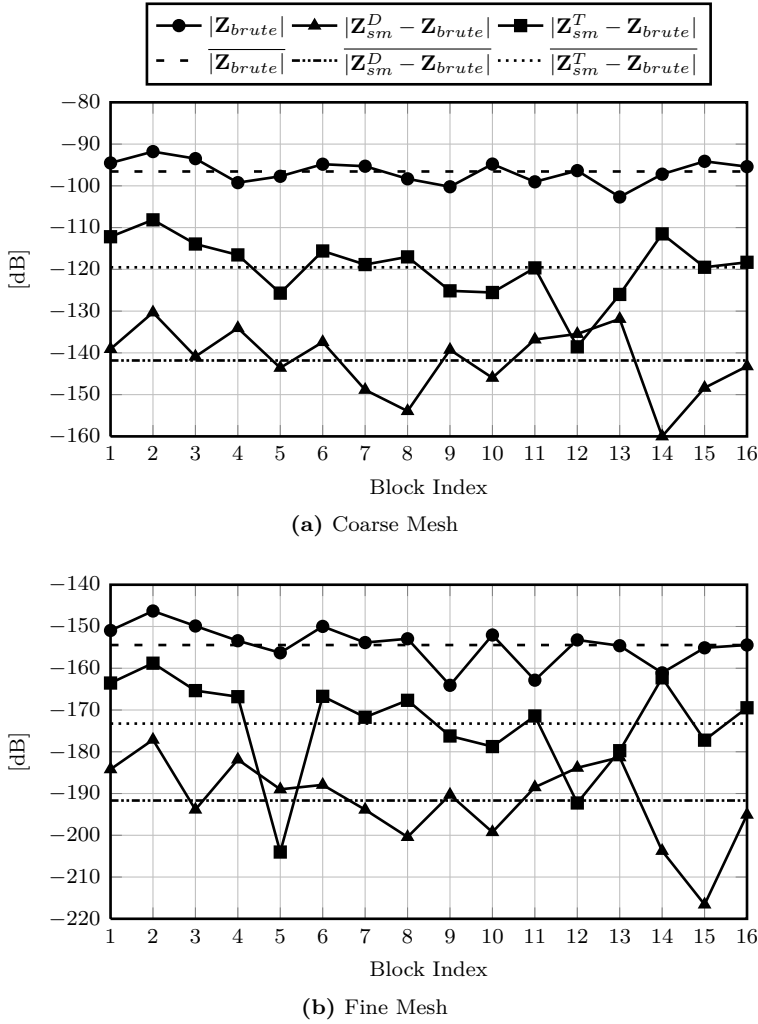


Figure 7.10: Absolute impedance values and error levels for different pairs of RWG basis functions in an octree for dyadic Green's model ($\mathbf{Z}_{sm}^D$) and analytic Taylor series ($\mathbf{Z}_{sm}^T$) implementations (a) when using a coarse mesh (block size: $\lambda/10$, RWG size: $\lambda/64$) and (b) when using a dense mesh (block size: $\lambda/64$, RWG size: $\lambda/410$).

approach is around -23 dB as seen in Fig. 7.10(a). Similarly, for a dense mesh of block edge length $\lambda/64$, the average error level achieved for the dyadic Green's approach is around -19 dB, whereas that for Taylor series expansion approach is around -37 dB, as seen in Fig. 7.10(b). This pleads in favor of the dyadic Green's approach, which offers superior error handling compared to the analytical Taylor series expansion.

The error level achieved for the smooth part with dyadic formulation is of the same order (or even a bit better) for the fine mesh; it is 20 dB worse for the coarse mesh. This is mainly due to the fact that for the fine mesh, the contribution of the smooth part becomes very small. Hence, one may expect that, for extremely fine meshes (basis functions smaller than $\lambda/1000$), lower-order Taylor expansion may be considered. Nevertheless, even for the coarse mesh used here, the -40 dB error level obtained here may be considered sufficient.

On a Metallic Sphere

The smooth part of the Green's function is then validated on a metallic sphere of radius 0.3λ using the matrix-vector multiplication described in Section 7.4.2. The discretization is done using the RWG basis functions. The total number of RWG basis functions is 9300. For an octree level 3 subdivision, there are 269 subdomains as shown in Fig. 7.5.

Fig. 7.12 shows the average error levels when implementing the fast matrix-vector multiplication on the smooth part of the Green's function using the dyadic Taylor series expansion on the entire sphere. The absolute value computed using a brute-force solution (only on the smooth part of the Green's function) is also given. The average error is about -40 dB. This can be further seen in the histogram plot of Fig. 7.13, generated from the relative error levels on all the elementary RWG basis functions.

7.5 Conclusion

This chapter explored the dense-mesh problem in the numerical analysis of the electromagnetic scattering by a metallic body. The problem to be solved corresponds to fast field evaluation on the whole computational domain, as

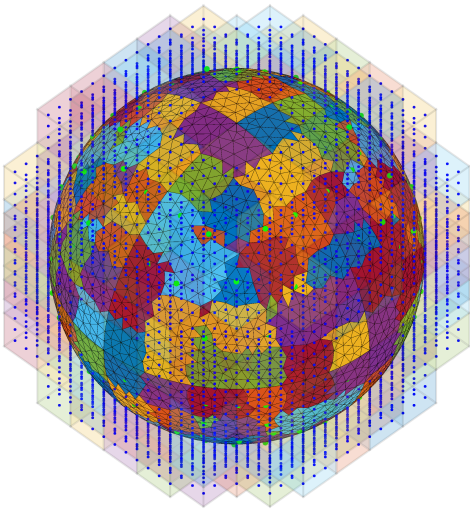


Figure 7.11: Metallic sphere of radius 0.3λ discretized and divided into subdomains.

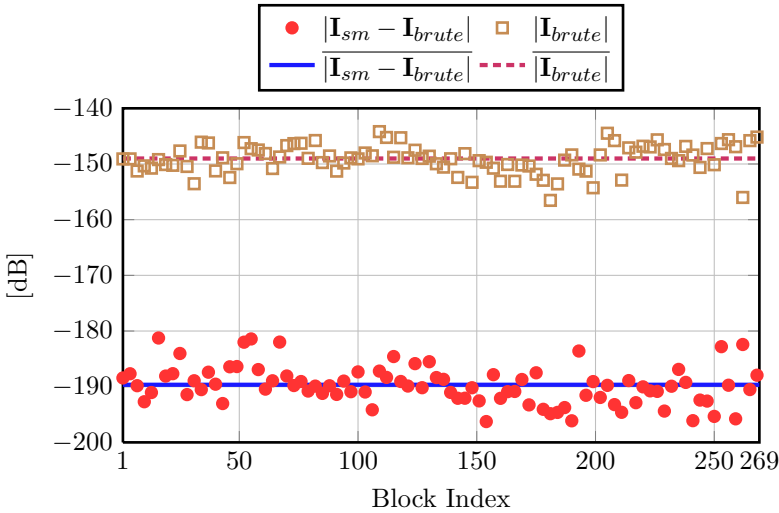


Figure 7.12: Tested fields using the fast matrix vector multiplication on the smooth part of the Green's function (I_{sm}) as compared with that of the brute-force solution (I_{brute}).

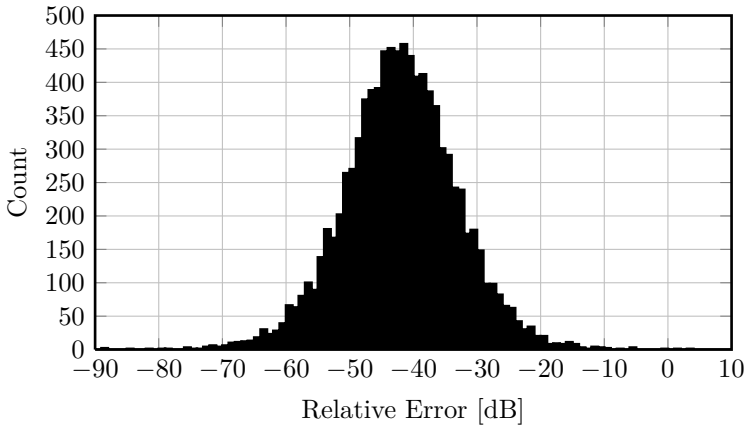


Figure 7.13: Histogram plot of the relative error in the tested fields using the fast matrix-vector multiplication on the smooth part of the Green's function and the brute-force solution.

radiated by currents on the same domain. In a basic MoM implementation, this operation is expressed as a fast matrix-vector product. Dynamic multipoles allow computation of this operation very fast without explicitly evaluating the matrix, but it is limited to a range larger than about a wavelength, which becomes an issue for meshes that are very fine compared to the wavelength as long as subwavelength-distance interactions are concerned. The improvement in speed comes from the ability to separate the Green's function into the singular and the smooth parts, and making the integrals in the EFIE separable. The technique scales very well with the wavelength. Therefore, the formulations proposed *do not* in principle depend on the wavelength, making this technique attractive to extremely dense meshes. For the case where the separation gets larger than half a wavelength, the classical multipoles can be applied.

Chapter 7 focused on the fundamental concepts, and the validations are mostly done on a few basis functions divided in an octree configuration. The fast matrix-vector multiplication has been briefly described, together with its general complexity. The problem with the dense mesh remains critical in antenna engineering and can be exploited in a wide range of applications.

Conclusions and Future Works

8.1 Conclusions

This thesis focused on the design, optimization, manufacturing and test of robust antennas intended for space applications. More specifically, the design activity focused on the circularly polarized conical beam antennas for the LaRa instrument aboard the ExoMars 2020 mission. The design activity for space applications introduced additional constraints in terms of the robustness of the design and the multi-physics nature of the analysis.

Furthermore, the thesis also dealt with the fundamental problem of improving the speed of electromagnetic modeling of the fine details inherent in antennas such as those for space applications. The numerical analysis aims to fill the gap between analysis of standalone antennas and the determination of their performance when mounted on large platforms, which is precisely the case for antennas devoted to deep-space probes or satellites. Because of the many finer details in the platform, more often than not, the antenna designers are compelled to resort to computationally intensive methods. For separations between the source and observation domains in the order of a wavelength, a hybrid technique employing Inhomogeneous Plane Wave method and Macro Basis Functions, may be used, which turns out to be efficient. But even more intense calculations happen at orders less

than half a wavelength, where classical multipole method cannot be applied. Therefore, a new formulation intended to speed up near-field evaluations on dense meshes has been conceptualized and presented in this thesis. It exploits the separability of the Green's function, which is itself split into singular and smooth parts.

The thesis work started off as a proposal to build a beamforming antenna aboard a Mars transponder for direct communication with Earth. The interactions of the antenna with the rest of the platform would be accelerated with a fast numerical technique. The beamforming network needed to be robust with low cost and low complexity, which was not possible without appropriate thermal control and exotic materials. Nevertheless, for terrestrial applications, the beamforming architecture proposed in this thesis provides a simple and low cost architecture.

The quest for an alternative to a beamforming antenna for the Mars transponder paved the way for simplifying the design with the added robustness and lower complexity. However, the pattern mask requirements and the envelope requirements did not make the design any easier. To completely satisfy the pattern mask requirements would have almost violated the energy conservation law given the size of the antenna and the sharp pattern requirements. In this respect, two different technologies were explored, and the innovations have been well detailed in this thesis work.

In particular, with respect to the planar antennas, low-profile slot array antennas and circular patch antennas have been explored. A complete design solution with the integrated feed network in a multi-layer layout has been detailed and analyzed. Because of a lack of space heritage of such antennas in a Martian-like environmental conditions, the planar designs had to be discarded. Again, the proposed solution is attractive to a terrestrial mobile terminal to satellite communications, especially at low elevation angles. With some modifications in the layout, the antenna can be easily mounted atop a moving vehicle for communications with satellites.

To meet the stringent requirements of the Martian transponder, a robust and high-performance solution was found in the form of a novel 3D metallic monopole parasitic antenna. The parasitic elements play an important role

in beamshaping the conical patterns over a wide range of elevation angles; they also ensure a high circular polarization purity. A low-loss feeding line for lateral feeding of the antenna, initially suggested by Marco Sabbadini from ESA, was also built for this design.

A summary of the original contributions is given below:

1. A simple and compact beamforming architecture for a conformal conical array [Chapter 2]
2. A simple feed and compact multi-layer design for planar antennas capable of generating circularly polarized conical patterns [Chapter 3]
3. A robust 3D monopole parasitic antenna capable of beamshaping the conical patterns with excellent circular polarization purity [Chapters 4 and 5]
4. A robust low-loss square coaxial feeding line [Chapter 5]
5. An analysis technique for complex antenna arrays using a hybrid IPW-MBF technique [Chapter 6]
6. A fast method for dense mesh problems [Chapter 7]

Contributions 3 and 6 are probably those that may have the most lasting impact, with applications in many areas within high-frequency electromagnetics.

8.2 Future Works

This thesis work has opened new avenues to further explore and exploit the technological advancements put forth. Specifically, the future works refer to improvements and extensions in the numerical analysis of the dense mesh, and to extensions in the beamshaping capability of the monopole parasitic antenna.

8.2.1 Numerical Techniques for Near Interactions

The full scale implementation of the analysis in dense mesh remains useful to reach a comprehensive parameterization and validation of the method.

It has been shown in Chapter 4 that the interactions can be sped up at very small separations without facing any stability issue. For too large separation (order of half a wavelength or more), the proposed method would not perform as accurately as the classical multipole method because of the oscillatory nature of the smooth Green's function. But, at this separation and larger, a *handover* from the Taylor series approach to the classical multipole method might occur. The efficiency of this handover still requires some attention. Then, an implementation combining static and dynamic multipole as well as Taylor expansion could lead to a software from which design of complex antenna could benefit.

8.2.2 Beamshaping with Metallic Parasitic Elements

The literature review on the circularly polarized conical beam antennas showed that the conical beam antennas for low-elevation angles were found lacking. This led to the design of a novel 3D metallic monopole parasitic antenna. Since a vertical monopole cannot radiate in the broadside, placing the gamma-shaped parasitic elements vertically around the monopole helped to control the angle of maximum gain of the conical beams. As such, the same design technology *cannot* be used to radiate at broadside. Many high-gain applications for space antennas require the ability to radiate in the broadside or have the ability to steer the beam off the broadside. A conical beam antenna would not be able to serve such purpose.

If it were possible to have the metallic elements radiate at broadside, an array of those elements would in principle make it possible to have high gain beamsteering patterns. The metallic elements would prove attractive to many space applications such as for Synthetic Aperture Radar (SAR) and telecommunication satellites. The elements themselves can be geometrically quite complex, requiring very dense mesh for analysis. With the method developed for dense mesh applied, the analysis can be done rapidly.

The parasitic elements can be made of multiple rings to allow for higher gain conical patterns. The challenging part would be to ensure high circular polarization purity. Scaled further down in their electrical size, those metallic elements also offer the possibility to homogenize the current distributions

in such a manner that an arbitrary radiation pattern may be synthesized. This is already the case for *metasurface* antennas. An extension of these types of antennas would be to 3D metallic elements or quite rightly termed *metavolume* antennas. The metallic elements are stressed here again simply because they prove to be an excellent radiator in hash environmental conditions, as is the case for space antennas.

APPENDIX A

Design of CP Conical Beam Circular Patch Antenna

This appendix serves as a complement to the work discussed in Chapter 3. More specifically, the works done in the context of the patch antenna design for the LaRa project warranted some more details with regards to the antenna assembly. Furthermore, the design methodology and the design parameters are also detailed in this appendix.

A.1 Antenna Assembly

The radiating elements correspond to a circular microstrip patch antenna, excited at four ports, to generate a TM_{31} field mode under the patch. To this end, the phase shift between excitation at consecutive ports is made equal to 90° . The phase shift between consecutive feed points, as well as the circular symmetry of the structure prevent radiation at broadside and creates a conical radiation pattern with radiation at an elevation angle significantly lower than requested in the LaRa specifications.

The fully assembled patch antenna assembly is shown in Fig. A.1. It consists of an antenna PCB, mounted on a housing, and fed with a semi-

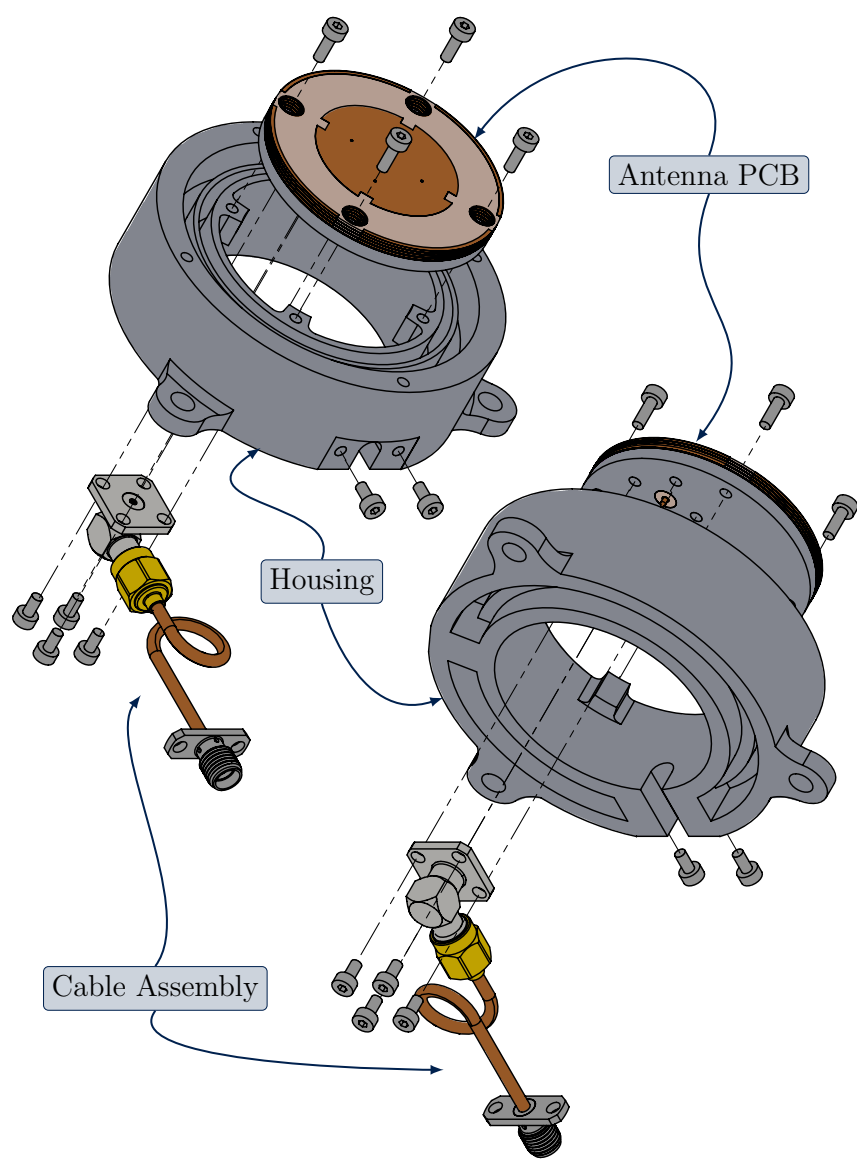


Figure A.1: Patch antenna assembly.

rigid cable assembly. The antenna PCB is first mounted with four screws. A pin protrudes from the bottom of the antenna PCB. One end of the pin is already soldered inside the PCB. This pin mates with the field-replaceable connector of the semi-rigid assembly. The semi-rigid assembly with the connectors and cable is built separately, and after the field-replaceable connector latches onto the pin from the PCB, the assembly is fastened with four screws to the bottom of the PCB, and also fastened to the antenna housing using two screws.

The antenna is fed with a coaxial connector on the bottom of the antenna. The inner pin of the connector penetrates the bottom ground plane and excites the feed chain. A hole is made in the *center* of the PCB to solder the stripline and connector pin from the top. For this, the bottom substrate layer (*Substrate 4*) with the feed network is first etched out. A hole is drilled at the feeding port, and a central contact (for instance, Radi-all R280 463 000) is inserted and soldered to the *Feed Network* layer. The rest of the layers is sequentially built. Once all the layers are stacked, the Plated Through Holes (PTH) are drilled. Four large PTHs are used for the screws to connect the PCB to the antenna housing. Four smaller PTHs are used as vias to connect the feed network to the patch layer. These PTHs are backdrilled to remove the excess stub from the *Ground Plane 2* to the *Feed Network* layer. While this was not entirely necessary, it was done to ensure that the feed PTHs were not shorted to the bottom aluminum carrier plate. The aluminum carrier plate is finally bonded to the PCB with a non-conductive adhesive (for instance, with Ablefilm 5025E from Henkel).

The feed chain is designed to equally split the amplitude of the power to four different pins with a consecutive phase difference of 90° . Those pins are directly connected to the patch antenna placed on top of the multilayer structure. An elbow connector is used to directly feed the antenna, while a second one connects the cable to the side of the holder. There are multiple grounding vias (vias from *Feed Network* to *Ground Plane 2*) to suppress any standing waves in the *Feed Network*. The stacked configuration is side-plated to ensure the ground planes are connected, which assures better performance of the stripline circuit and avoids lateral radiation.

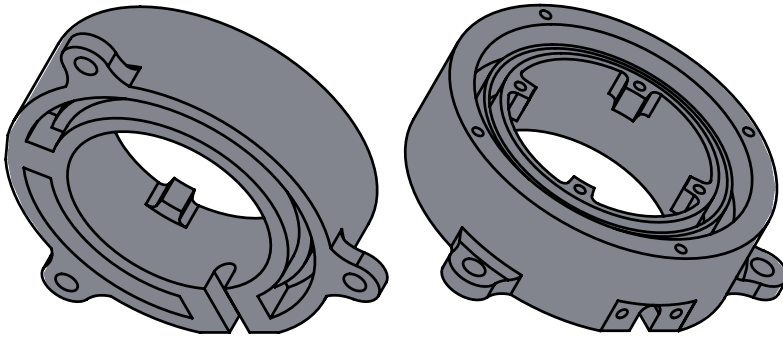


Figure A.2: Antenna choke ring housing of the CP patch antenna.

Choke Ring Housing

In order to support the antenna and to feed it from the bottom, the PCB antenna is held in an aluminum cylindrical structure, visible in Fig. A.2. This structure is fastened with three mounting holes to a platform. The holder also includes the choke rings, corresponding to several concentric rings around the antenna PCB.

Those rings stop the current from flowing further on the surface and rather create an effect of having a much larger ground plane [75]. The rings also limit the back-radiation, which in turn concentrates the energy in the upper hemisphere. The latter effect leads to an increased gain. The choke rings are characterized by their depths and the widths and are used to optimize the gain. Indeed, the choke rings elevate the radiation pattern, up to an angle consistent with the LaRa requirements.

Antenna Interface

The antenna interface consists of a semi-rigid cable assembly, and an elbow connector shown in Fig. A.3. The semi-rigid cable assembly is connected on one end to a 2-hold flange connector, while the other end is connected to a straight plug. This plug is then screwed to a field-replaceable elbow connector. The central contact soldered in the antenna PCB latches on to the elbow connector, which is then screwed in the aluminum carrier plate.

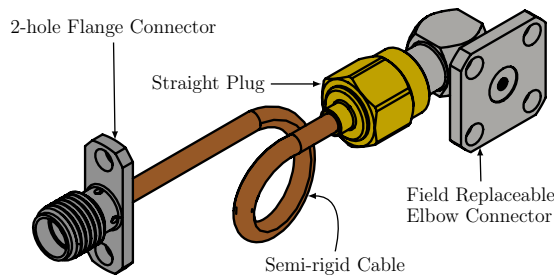


Figure A.3: Semi-rigid assembly for feeding the CP patch antenna.

A.2 Design Methodology and Parameters

The antennas have been designed using the CST Microwave Studio commercial software. The first step consists of separately designing the patch antenna and the feeding network. The former assumes four ideal feeds and the latter observes the voltages on four ideal loads. The common impedance of feeds and loads is a free design parameter; good performance has been observed for a common impedance of $80\ \Omega$ although bringing this impedance down to $50\ \Omega$ would increase the overall trace width of the stripline making it less susceptible to manufacturing tolerances.

The second step consists of the combined optimization of the microstrip antenna and feeding network. This supposes the modification of nearly all geometrical parameters of the structure. Here, the main quality criteria correspond to antenna matching and regularity of the conical radiation pattern versus azimuth. The latter parameter is highly sensitive to the phase imposed by the feeding network on the four feeding pins. This, in turn, means high sensitivity to the line width and lengths that make up the feeding network. This is why those dimensions should receive primary attention in the sensitivity analysis.

The third step includes the housing and the optimization of the choke rings to obtain the radiation patterns adhering to the gain and axial ratio requirements versus elevation. The housing also impacts the antenna impedance, and this step entails re-tuning of parameters determined in previous steps. This explains why the antenna optimization is a very lengthy process.

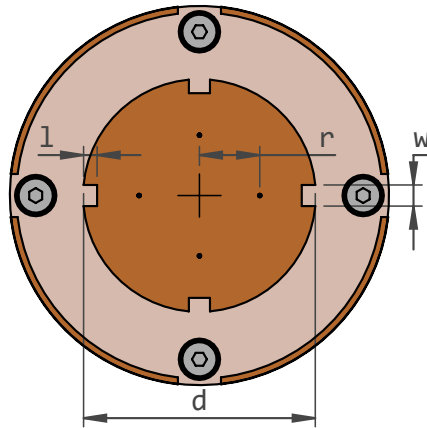


Figure A.4: Circular patch parameters of the CP patch antenna.

There exist well established methodologies to analytically design patch antennas, power splitters and transmission lines. These analytical results are used as the starting point of the antenna design. Then, each part of the antenna is rigorously analyzed with a full-wave method (CST Microwave Studio) and then optimized to obtain the desired characteristics. This optimization process is quite intensive. In a first instance, the feed chain, the antenna patch and the choke rings of the antenna housing are designed independently. Then, they are combined and re-optimized to assure a better performance for the complete antenna. The global design of the antenna comprehends several tens of variables and each iteration of the electromagnetic analysis of the complete antenna takes more than an hour in a workstation. Therefore, these characteristics make the optimization problem extremely complex and lengthy.

Microstrip Patch Antenna

The patch antenna is mainly defined by its diameter of the patch, and by the position of the excitation pins. To preserve symmetry, just the radial distance is considered, as shown in Fig. A.4. Four notches (width and length) are included to improve the azimuthal variation and to improve the matching of the patch antenna.

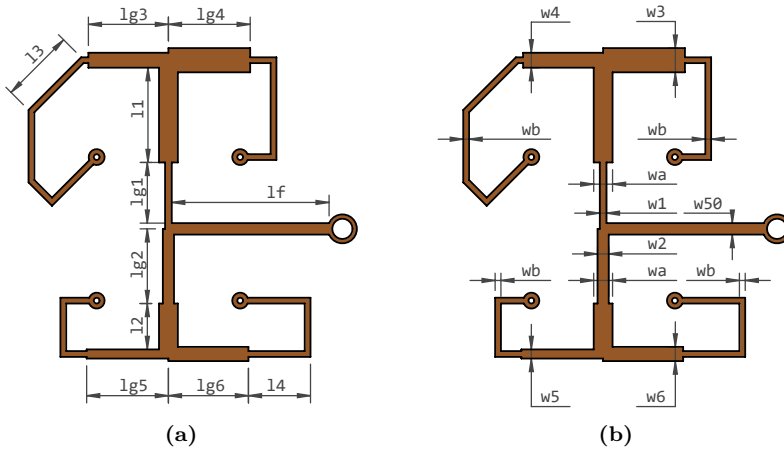


Figure A.5: CP patch antenna feed network dimensions: (a) Lengths and (b) Widths.

Feed Network

The feed chain is shown in Fig. A.5; the widths and lengths that play an important role in the optimization are represented in the figure. The antenna is excited with a $50\ \Omega$ SMA coaxial connector. The connector pin is connected in the center of the antenna PCB to maintain the overall symmetry of the design and to allow sufficient space to attach the connector beneath the PCB. The central contact is directly connected to the feed chain. Then, the impedance is decreased with quarter wave transmission lines to avoid very thin lines (consequence of high impedance lines). An extra transmission length is included at one side for obtaining the desired phase distribution. In both cases, the power is divided again and the widths are modified to account for the proper impedance in each excitation pin. The feed chain is very sensitive to small variations, and it has many parameters of interest, as seen in Fig. A.5.

Choke Ring Housing

A section view of the choke rings is shown in Fig. A.6. The choke rings help avoid the back radiation and elevate the patterns. The depth (d_1, d_2),

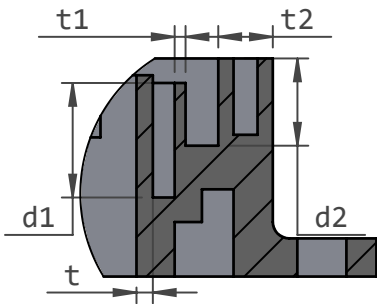


Figure A.6: Choke ring parameters of the CP patch antenna housing.

separation ($t1, t2$) and the number of choke rings are the design parameters used to enhance the antenna performance. The depth is usually a quarter wavelength to begin with, but depending on the optimization goals, the rings can have different depths. The separation and depth of the choke rings can be quite different for each choke ring and the depth can be a decreasing or increasing height.

Design Parameters

Table A.1 provides the values of the most important parameters used in the design of the patch antenna.

Table A.1: Design parameters of the patch antenna.

Parameter	Value	Parameter	Value
<i>lg1</i>	4.18	<i>w1</i>	0.46
<i>lg2</i>	5.15	<i>w2</i>	0.77
<i>lg3</i>	5.52	<i>w3</i>	1.67
<i>lg4</i>	5.64	<i>w4</i>	1.03
<i>lg5</i>	5.64	<i>w5</i>	0.63
<i>lg6</i>	5.52	<i>w6</i>	0.98
<i>l1</i>	6.53	<i>wa</i>	1.3
<i>l2</i>	3.16		
<i>l3</i>	5.13		
<i>l4</i>	4.28	<i>wb</i>	0.4
<i>lf</i>	10.85	<i>w50</i>	1
<i>r</i>	7	<i>t</i>	1.5
<i>d</i>	27	<i>t1</i>	1
<i>w</i>	2.45	<i>t2</i>	5
<i>l</i>	1.57	<i>d1</i>	10.5
		<i>d2</i>	8

APPENDIX B

List of Publications and Research Talk

Journal Paper

- J1. S. Karki**, M. Sabbadini, K. Alkhalifeh, and C. Craeye, “Metallic monopole parasitic antenna with circularly polarized conical patterns,” under review in *IEEE Transactions on Antennas and Propagation*, submitted on 30 September 2018.

Conference Papers

- C1. S. Karki**, J. Ignacio Echeveste, T. Pairon and C. Craeye, “Sequentially shifted beamforming from a conformed conical array,” *2017 European Radar Conference (EuRAD)*, Nuremberg, 2017, pp. 367-369.
- C2. S. Karki**, C. Craeye, M. Mitrovic and V. Dehant, “A conical frustum-type array devoted to a Mars-based transponder,” *2016 10th European Conference on Antennas and Propagation (EuCAP)*, Davos, 2016, pp. 1-4.
- C3. S. Karki** and C. Craeye, “Macro Basis Function interactions ac-

celerated with the Inhomogeneous Plane Waves,” *2015 International Conference on Electromagnetics in Advanced Applications (ICEAA)*, Turin, 2015, pp. 1314-1317.

- C4.** E. Martini, **S. Karki**, C. Craeye and S. Maci, “Error analysis of the harmonics-based plane wave method,” *2015 International Conference on Electromagnetics in Advanced Applications (ICEAA)*, Turin, 2015, pp. 1493-1495.
- C5.** K. Alkhalifeh, S. N. Jha, **S. Karki** and C. Craeye, “On the use of Contour-FFT for the MBF-based analysis of arrays of antennas placed vertically above a multi-layered substrate,” *2015 9th European Conference on Antennas and Propagation (EuCAP)*, Lisbon, 2015, pp. 1-5.

Research Talk

- T1.** **S. Karki**, C. Raucy, and C. Craeye, “Fast Near-Field Iteration Based on Static and Dynamic Green’s Functions.” *2017 XXXIIInd General Assembly and Scientific Symposium of the International Union of Radio Science (URSI GASS)*, Montréal, Aug. 2017.

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