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Projeto e Caracterização Experimental de Antena
Monopolo Assistida por Metamateriais

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AUTORIZO A REPRODUÇÃO TOTAL OU PARCIAL DESTE TRABALHO,
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*To my beloved parents, Cristina and William,
with love, admiration and my eternal gratitude.*

*And to all those who seek knowledge
as a form of progress for humanity.*

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Everything in space obeys the laws of physics. If you know these laws, and obey them, space will treat you kindly. And don't tell me man doesn't belong out there. Man belongs wherever he wants to go – and he'll do plenty well when he gets there.

Wernher Magnus Maximilian von Braun

ABSTRACT

LIMA, L. C. P. S. **Design and Experimental Characterization of a Metamaterial-assisted Monopole Antenna.** 2014. Dissertation (Master of Science) – Escola de Engenharia de São Carlos, Universidade de São Paulo, São Carlos, 2014.

In recent years a new class of materials, the metamaterials, has emerged in the scientific community. The use of these materials makes possible to achieve unique electromagnetic properties, such as the negative refractive index. Today there exist several applications that take advantage of these special properties, such as sensors, antennas and invisibility cloaks, aiming at improving their intrinsic characteristics. Based on these considerations, this project aims at developing metamaterials structures to control the radiation properties of antennas in the microwave range, such as gain and directivity. More specifically, it was also chosen chiral metamaterials, mainly due to the phenomenon of electromagnetic field rotation which opens the possibility to control efficiently the radiation properties of antennas. In addition, chiral metamaterials, which have proved to be a more attractive alternative to obtain negative or zero refractive index, enable a greater degree of freedom in the design of different structures. This work encompasses all different phases of the structure design, namely: project, computational modeling, fabrication, and characterization of the proposed structures. We show improvements for the gain that in some cases reaches more than the double of the conventional monopole antenna gain and for the return loss parameter, which reaches minimum values. We also could maintain good efficiency and improve the input impedance matching. Finally, it is worth mentioning that this new technology also has the great potential to be applied in the telecommunication devices, particularly to improve communications based on antennas.

Keywords: metamaterials, chiral metamaterials, antennas, microwaves, negative and near zero index of refraction, electromagnetic theory, telecommunications.

RESUMO

LIMA, L. C. P. S. **Projeto e Caracterização Experimental de Antena Monopolo Assistida por Metamateriais**. 2014. Dissertação (Mestrado) – Escola de Engenharia de São Carlos, Universidade de São Paulo, São Carlos, 2014.

Nos últimos anos uma nova classe de materiais, os metamateriais, emergiu na comunidade científica. O uso desses materiais torna possível alcançar propriedades eletromagnéticas singulares, como o índice de refração negativo. Hoje existem vastas aplicações que usufruem destas propriedades especiais, como os sensores, mantas de invisibilidade e antenas, onde se procura o aperfeiçoamento de suas características intrínsecas. Com base nestas considerações, este projeto buscou desenvolver estruturas metamateriais para controle das propriedades de radiação de antenas na faixa de micro-ondas, tais como diretividade e ganho. Mais especificamente, foram utilizados os metamateriais quirais, principalmente devido ao fenômeno de rotação do campo eletromagnético que abre a possibilidade de controle mais eficiente das propriedades de radiação de antenas. Além disso, os metamateriais quirais, por se mostrarem uma alternativa mais atraente para se obter meios com índice de refração zero ou negativo, possibilitam um maior grau de liberdade no projeto de diferentes estruturas. Este trabalho contempla, ainda, todas as etapas de projeto de tais estruturas, quais sejam: projeto, modelagem computacional, fabricação, e caracterização das estruturas. Mostramos melhorias para o ganho que, em alguns casos, chega a mais do que o dobro do ganho da antena monopolo convencional e para o parâmetro de perda de retorno, que atinge valores mínimos. Nós também mantivemos uma boa eficiência e melhoramos o casamento de impedância de entrada. Finalmente, vale salientar que essa nova tecnologia também apresenta grande potencial de ser aplicada em dispositivos de telecomunicações, com o intuito de aprimorar a comunicação baseada em antenas.

Palavras-chave: metamaterias, quirais, antenas, micro-ondas, índice de refração zero e negativo, teoria eletromagnética, telecomunicações.

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LIST OF ACRONYMS

2D	Two Dimensional
3D	Three Dimensional
AUT	Antenna Under Test
BSE	Boresight Error
DNG	Double Negative
DPS	Double Positive
ENG	Epsilon Negative
FEM	Finite Element Method
FR4	Flame Retardant #4
FZP	Fresnel Zone Plate
GPS	Global Positioning System
HFSS	High Frequency Structure Simulator
LCP	Left-Circularly Polarized
LHM	Left-Handed Material
MNG	Mu Negative
PBC	Periodic Boundary Conditions
PEC	Perfect Electric Conductor
PMC	Perfect Magnetic Conductor
PTFE	Polytetrafluoroethylene
RCP	Right-Circularly Polarized
RF	Radio Frequency
Tx	Transmitter Antenna
RHM	Right-Handed Material
Rx	Receiving Antenna
SMA	SubMiniature version A
SRR	Split Ring Resonator
VNA	Vector Network Analyzer

LIST OF SYMBOLS

n	Index of refraction
ϵ	Electric permittivity (F/m)
ϵ_r	Relative electric permittivity (F/m)
ϵ_0	Permittivity in vacuum (F/m)
μ	Magnetic permeability (H/m)
μ_r	Relative electric permeability (H/m)
μ_0	Permeability in vacuum (H/m)
c	Velocity of light in vacuum (m/s)
λ	Wavelength (mm)
λ_0	Wavelength in vacuum (mm)
f_0	Operating/design frequency (GHz)
f_r	Resonance frequency (GHz)
f_{r1}	First resonance frequency (GHz)
f_{r2}	Second resonance frequency (GHz)
f_{r3}	Third resonance frequency (GHz)
G_{fr}	Gain in the resonance frequency (dB)
G_{fr1}	Gain in the first resonance frequency (dB)
G_{fr2}	Gain in the second resonance frequency (dB)
G_{fr3}	Gain in the third resonance frequency (dB)
η_{fr}	Efficiency in the resonance frequency
η_{fr1}	Efficiency in the first resonance frequency
η_{fr2}	Efficiency in the second resonance frequency
η_{fr3}	Efficiency in the third resonance frequency
Z_{in}	Input impedance (Ω)
Z_{infr2}	Input impedance in the first resonance frequency (Ω)
Z_{infr2}	Input impedance in the second resonance frequency (Ω)
Z_{infr3}	Input impedance in the third resonance frequency (Ω)
z	Impedance (Ω)
Z_0	System impedance (Ω)
θ	Azimuth angle ($^\circ$)
η	Ellipticity angle ($^\circ$)
κ	Chirality parameter
ρ	Charge density (C/m ³)
χ	Dimensionless magneto-electric parameter
n_+	Refers to RCP
n_-	Refers to LCP
l	Antenna length (mm)
r	Radius of the antenna's conducting wire (mm)
w_{gp}	Ground plane width (mm)
d	Distance from the antenna to the metamaterials cover (mm)
w_u	Width of the metamaterial cells (mm)
w_c	Width of the metal rings (mm)
l_c	Wire length (mm)
d	Width of the metamaterial cell (mm)
φ	Offset angle of the chiral metamaterial cell ($^\circ$)
φ_0	Tilting angle of the chiral metamaterial cell ($^\circ$)

α	Rotation angle of the 2-D chiral cells (°)
$\tan\delta$	Loss tangent of the material
D	Maximum linear dimension that contains the antenna (mm)
R	Separation between the antenna and a region in space (mm)
E	Electric field intensity (V/m)
H	Magnetic field intensity (A/m)
D	Electric displacement (C/m ²)
B	Magnetic displacement (Wb/m ²)
J	Current density (A/m ²)
J_{ext}	Electric current density [A/m ²]
P	Polarization
M	Magnetization
S	Poynting vector
k	Wave vector
ρ_{ext}	Volume charge density [C/m ³]
P_{rad}	Power radiated by the antenna
P_{in}	Power supplied to the input of the antenna
U_{max}	Maximum radiation intensity in the main beam
U_{avg}	Average radiation intensity over all space
$D(\text{dB})$	Directivity expressed in dB
G	Gain of the antenna
S_{11}	Reflection coefficient or (Input) Return Loss (dB)
S_{22}	Reflection coefficient or (Output) Return Loss (dB)
S_{12}	Transmission coefficient (dB)
S_{21}	Transmission coefficient (dB)
k_0	Wave number
T_{xx}	Linear transmission coefficient
T_{yy}	Linear transmission coefficient
T_{xy}	Linear transmission coefficient
T_{yx}	Linear transmission coefficient
T_{++}	Transmission coefficients for RCP
T_{--}	Transmission coefficients for LCP
$P1$	Port 1 of a two-port network
$P2$	Port 2 of a two-port network
a	Incident signal on each port
a_1	Input signal at P1 of the two-port network
a_2	Input signal at P2 of the two-port network
b	Output signal of each port
b_1	Output signal at P1 of the two-port network
b_2	Output signal at P2 of the two-port network
r_n	Radius of the patch antenna (mm)
r_a	Minor radius of the patch antenna (mm)
r_b	Major radius of the patch antenna (mm)
L	Substrate length of the patch antenna (mm)
L_f	Feed line length of the patch antenna (mm)
W	Substrate width of the patch antenna (mm)
W_f	Feed line width of the patch antenna (mm)

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CHAPTER 1

1 INTRODUCTION

1.1 MOTIVATION

The interaction of electric and magnetic fields with different materials plays an essential role in the study of electromagnetism. Aiming constructive interactions, a new class of material, namely the metamaterials, is causing a tremendous impact in the scientific community in the last years, mainly due to their extraordinary electromagnetic properties, such as simultaneously negative electrical and magnetic responses and, consequently, the negative refractive index.

Metamaterials can be defined as artificial and structured materials with constitutive parameters designed to produce desired electromagnetic responses [1]. Nowadays, the progress in this technology has imposed challenging demands on material properties. For this reason, metamaterials have become ubiquitous in sensors [2],[3], in cloaking devices to guide electromagnetic waves around hidden objects [4]-[6], in antennas to improve their performance [7]-[27], in superlenses to achieve resolution beyond the diffraction limit [28], in absorbers to absorb large amounts of electromagnetic radiation [29],[30], and so on.

Antennas, in particular, have become one of the most exciting applications of metamaterials due to the possibility of significantly improving their performance [31]. It is worthwhile pointing out that the performance of these antennas (in terms of gain, return loss, directivity, and so on) is greatly influenced by a large number of parameters, such as substrate and geometry. Therefore, engineered materials are seen as the perfect choice for optimizing performance since they allow one an infinite degree of freedom to control (even on an individual basis) each one of these parameters. An additional challenge is the utilization of chiral metamaterials. Provided by these structures, the ability of the field to rotate can be greatly helpful for tailoring the radiation pattern and gain of these antennas, for instance. The effectiveness of (chiral) metamaterials is strongly dependent on the cell geometry adopted (as well as the substrate). Therefore, this work also investigates the performance of different metamaterial's cell geometries in the microwave region.

1.2 HISTORICAL OVERVIEW

The development of artificial materials for manipulating electromagnetic waves dates from the late 19th century. Some of the earliest structures that may be considered metamaterials date back to 1898, when Bose conducted the first microwave experiment on twisted structures [32] with chiral properties. Later, in 1920, Karl Lindman studied the wave interactions with metallic helices of subwavelength size [33], considered an isotropic artificial chiral media.

It is known that in 1948 W. E. Kock created microwave lenses organizing periodically some structures to adapt the effective refractive index of the material [34]. The need of radar technology for higher permittivity low loss materials resulted in the development of artificial dielectrics [35].

Artificial dielectrics were also studied in the 1950s and 1960s. Brown explored in 1953 structures based on arrays of thin wires [36] that, later, were used by Smith [37] and Shelby [38] to produce simultaneously negative permittivity and negative permeability (and therefore, negative refractive index). In 1955, Thompson demonstrated the existence of negative permeability with experiments performed on ferrite [35].

Viktor Veselago, in turn, theoretically investigated plane wave propagation in a material whose permittivity and permeability constants were considered negative [39]. In his theoretical study conducted in 1967, Veselago proved that substances with a negative index can indeed transmit light. He showed that for a monochromatic uniform plane wave propagating through such a medium, the direction of the Poynting vector is anti-parallel to the direction of the phase velocity, contrary to what is expected in naturally-occurring materials [1].

These ideas remained quietly in the background until the new millennium, where in 2000, Sir John Pendry ushered in the remarkable ideas that support the development of these new materials, and published a theoretical study [28] based on the negative refractive index to produce the so called perfect lens. This perfect lens was not diffraction limited, and was able to focus all Fourier components of a 2D image. This paper gave an enormous visibility to the new field of metamaterials, causing it to emerge as a new promising and revolutionary technology.

1.3 THE CONCEPT OF METAMATERIAL

A metamaterial can be created from other conventional materials, where one can change some geometric parameters such as shape, size and composition in order to obtain the desired properties. One can also integrate these conventional materials by modifying their arrangement and/or alignment. By working with these characteristics, it is possible to design a metamaterial with specific electromagnetic responses, allowing a variety of new possibilities in the creation process of metamaterials [34].

The electromagnetic parameters to be analyzed here are: the refractive index (n) and the impedance (z), that characterize the electromagnetic properties of a material and are used when discussing wave propagation; and the electric permittivity (ϵ) and the magnetic permeability (μ), assumed here as analytical variables that impart a material interpretation [40]. Due to the constructive characteristics of metamaterials, these properties are considered in effective terms hereinafter.

As was already mentioned, Smith et al. reported their experiments of electromagnetic metamaterials by periodically arranging split-ring resonators (SRR) with thin wire structures [37]. For this reason, one of the ways to obtain metamaterials is by using these types of resonators arrays, consisting of metal rings that do not close completely (they present a gap), and that emulate and outweigh the electromagnetic properties of natural solids [41]. Each resonator (or cell) in a metamaterial can be seen as a macroscopic atom so, as an illustration, a comparison between atoms of a natural solid and those of a metamaterial manufactured from resonators is shown in Figure 1-1.

Figure 1-1: Arrangement of metamaterials resonators compared to atoms (adapted from [41]).

To better understand the materials' classification, Veselago [39] first introduced an $\varepsilon - \mu$ diagram where, according to the characteristics of the permittivity and the permeability, it was possible to define the material type.

In Figure 1-2, it is possible to see that the first quadrant of the diagram contains the majority of isotropic dielectrics, for which ε and μ are positive. In the second quadrant, there are plasmas, where ε is negative and μ is positive. By the time Veselago published the English version of his famous paper, in 1968, the third and fourth quadrants of this diagram were unoccupied. But later, when metamaterials were already being consolidated in the scientific community, Engheta and Ziolkowski [34] presented a new $\varepsilon - \mu$ diagram, with all four categories fulfilled and new denominations assigned.

Figure 1-2: Classification of materials (adapted from [42]).

In the first quadrant we have DPS (Double Positive) materials, such as the dielectrics, in the second quadrant ENG (Epsilon Negative - negative ϵ) materials, such as plasmas, metals at optical frequencies, just as suggested by Veselago. Now, in the third quadrant we have DNG (Double Negative) materials physically realizable by artificially engineered metamaterials, and in the fourth quadrant we have MNG (Mu negative - negative μ) materials which are the magnetic materials, such as pure ferromagnetic metals and semiconductors [39]. It is important to remember that metamaterials can be designed to work not only in the third quadrant, where ϵ and μ are simultaneously negative, but in any of the four quadrants presented in Figure 1-2 (therefore resulting in DPS, ENG, DNG or MNG materials). Still in this figure, at the intersection of the axes where a red circle is displayed, we have the region known as "near-zero refractive index" or region of "zero refractive index". Metamaterials with zero refractive index, for example, are utilized to increase the directivity of antennas [23].

When a material exhibits negative permittivity and permeability (DNG material), its unique characteristics can reveal strategic concepts with potential applications, such as lenses [28], invisibility cloaks [4],[5], sensors [2],[3], and antennas, focus of the present work.

In the case of antennas, their design is considered an art of Engineering, and an important factor in the successful use of these devices is related to the great advances in computing architectures and numerical calculations [43].

Studies of metamaterials applications on these devices have shown distinct improvement in performance characteristics, such as increased radiated power or gain [7]-[11],[44], increased efficiency [9],[11]-[13], improved impedance matching [9],[14], minimized losses [15],[16] and increased bandwidth [17]. It is also possible to manipulate and control the directivity of an antenna and, in the case of a receiving antenna system, it is possible to increase the electric field reception [18]. When metamaterials with refractive indices near zero are used, increased directivity [19], increased gain and decreased side lobes are observed [20]. Near zero refractive index metamaterials also enable substantial improvement in efficiency [21] and gain [22],[23], reduction of side lobes and good impedance matching [22], and improved directivity [23]. As can be seen, there is a wide variety of antenna parameters that are possible to be improved with the use of metamaterials.

In this work, emphasis is given to chiral metamaterials, mainly due to its electromagnetic field rotation properties, and also because they are an attractive way of obtaining the negative refractive index at any frequency, offering a simpler and more efficient geometry [45]. Moreover, for the sake of comparison, it is also investigated the characteristics of conventional metamaterials to ultimately define the optimized structure.

Chiral metamaterials are metamaterials made of unit cells without symmetry planes [45]. By ordering these unit cells in a periodic arrangement, it is possible to obtain special properties, such as negative refraction and rotation of the magnetic field. Studies of chiral media in the microwave region have suggested applications in several areas, such as antennas, polarizers and waveguides [46]. In antennas, chiral media have been used to obtain circular polarization in all directions of radiation [24]-[27]. Numerical and experimental studies involving structures composed of chiral metamaterials were conducted to prove the increase of optical activity and polarization effects [47]-[51], circular dichroism [52], and the negative index of refraction [52],[53].

As can be seen, the use of metamaterials media has been consistently investigated by different authors in different frequency regimes (from optical to microwave), giving rise to different propagation effects and structures. More examples can be found for phase compensation with small resonators as suggested by Engheta [54]. It was also observed, among many other results, the inverted effect of Cherenkov radiation (derived in [55]), the inversion of Doppler shift [56], and the displacement of Bragg regime [57]. Metamaterials can also be applied to the design of compact microwave filters and diplexers [3], since they are frequency-selective structures by nature. SRRs and CSRRs (Complementary SRRs) proved to be useful cells for the design of compact planar filters in microstrip technology [58]. Reconfigurable metamaterials have been also investigated [59]-[61].

In the case of metamaterial transmission lines, it is possible to design devices with a better performance compared to conventional ones (such as enhanced bandwidth devices), to design components based on new functionalities (such as dual-band components), or microwave devices with smaller dimensions [3].

Dual-band components (devices that present two different operating frequencies), for instance, are of great interest for modern microwave and wireless communication systems, once they make possible operation at two different bands without the need to design different mono-band circuits [31]. Contrary to conventional (right-handed) transmission lines, which

are intrinsically mono-band structures, composite right-/left-handed (CRLH) transmission lines exhibit a dual-band behavior [62]. This kind of dual-band behavior will also be explored in this work.

With all information provided, the potential of metamaterials and metamaterial-based structures for the design of microwave devices is in fact notable, mainly due to the novel functionalities and improved performance of their unique and controllable electromagnetic properties. The demand for metamaterial-based applications will continue to grow in future years, and this includes civil and military applications. Therefore, we expect with the present work to provide a small contribution to this ever-growing and dynamic area.

1.4 CONTRIBUTIONS OF THIS WORK

Succinctly, the purpose of this work is to develop structures assisted by (chiral) metamaterials aiming at improving the performance characteristics of antennas.

Fundamentally, we review the concept of metamaterials, their properties and their main applications, and how electromagnetic wave propagation is affected when propagating through these materials. In practical terms, we investigate a conventional antenna that could be implemented with metamaterials and the effects of using metamaterial structures coupled to antennas.

To do so, we design, numerically analyze, and experimentally characterize a monopole antenna and investigate the effectiveness of a (chiral) metamaterial cover as a way to optimize/improve the main figure of merits on this antenna, such as gain, return loss, efficiency, and so on.

For this case, we successfully show improvements for the gain that in some cases reaches more than the double of the conventional monopole antenna gain and for the return loss parameter, which reaches minimum values. We also could maintain good efficiency and improve the input impedance matching.

Moreover, we also investigate the effect of structured ground planes (concentric rings as opposed to flat surfaces) on the performance of a elliptical patch antenna, and how they affect the gain and directivity.

1.5 ORGANIZATION OF DISSERTATION

This dissertation is organized into seven chapters. The first chapter contains the introduction of this work and our main objectives.

The second chapter addresses relevant topics which are required for the understanding of this work, such as the definition of metamaterials and chiral metamaterials, and the phenomenology of electromagnetic wave propagation in metamaterials.

The third chapter deals with fundamental concepts about antennas, their operation principle, with examples of some of the most common antenna types used in telecommunications. Moreover, it emphasizes the use of antennas with metamaterials, the focus of this dissertation.

Chapter 4 presents the numerical analysis and the software used to simulate the problem. The analyses of the metamaterials main parameters, using retrieval methods, are also presented.

Chapter 5 presents the simulation results together with detailed discussions, highlighting the improvements achieved with the proposed approach. In this chapter we also present the experimental characterization of the antenna with a cover for a given distance.

Chapter 6 is a special chapter and introduces the concept of a new patch antenna with a ground plane based on concentric rings.

Finally, chapter 7 contains the conclusions and future perspectives for this work.

The appendices are also essential, since they present how the scattering parameters are obtained and the method utilized for the parameter retrieval. In addition, they provide information regarding the software utilized and its configurations, and the development of the structures, with details of the fabrication and measurements.

CHAPTER 2

2 ELECTROMAGNETIC WAVE PROPAGATION IN METAMATERIALS

This chapter deals with the propagation of electromagnetic waves in (chiral) metamaterials. It also discusses some metamaterial's applications in the microwave regime.

2.1 ELECTROMAGNETIC WAVE PROPAGATION IN METAMATERIALS

The basic principle of various electromagnetic devices is grounded in electricity and magnetism phenomena, which have been studied and analyzed since the 19th century. In 1820, Hans Ørsted discovered that electric currents in a carrying wire create magnetic fields. This fact led André-Marie Ampère to develop in the same year a theory to understand the relationship between electricity and magnetism, that later came to be called Ampère's law, relating magnetic fields and electric currents. Afterwards, Michael Faraday discovered in 1831 the electromagnetic induction, showing the relationship between a changing magnetic field and an electric voltage.

In the 19th century, James Clerk Maxwell united the previous understanding of electrical and magnetic phenomena into his well known theory of electromagnetism, demonstrating that electricity and magnetism are manifestations of electromagnetic fields. He proved that electric and magnetic fields travel through space as waves moving at the speed of light. First, we recall the macroscopic Maxwell equations, which are the basic equations governing the electromagnetic response. These fundamental equations are defined as [63]

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \quad (1)$$

$$\nabla \times \mathbf{H} = \mathbf{J}_{\text{ext}} + \frac{\partial \mathbf{D}}{\partial t} \quad (2)$$

$$\nabla \cdot \mathbf{D} = \rho_{ext} \quad (3)$$

$$\nabla \cdot \mathbf{B} = 0, \quad (4)$$

where $\mathbf{E}$ is the electric field strength [Volt/m], $\mathbf{H}$ is the magnetic field strength [Ampere/m], $\mathbf{D}$ is the electric flux density (or the electric displacement) [Coulomb/m²], $\mathbf{B}$ is the magnetic flux density (or the magnetic induction) [Weber/m²] or [Tesla], ρ_{ext} is the volume charge density [Coulomb/m³], and $\mathbf{J}_{ext}$ is the electric current density (charge flux) [Ampere/m²] of any external charges (neglecting any induced polarization charges and currents). Equation (1) is Faraday's law of induction, equation (8) is Ampère's law modified by Maxwell to include the displacement current $\partial\mathbf{D}/\partial t$, equations (3) and (4) are Gauss' laws for the electric and magnetic fields.

Observe that equations (1) to (4) link the four macroscopic fields $\mathbf{E}$, $\mathbf{H}$, $\mathbf{D}$ and $\mathbf{B}$ to the external charge and current densities ρ_{ext} and $\mathbf{J}_{ext}$ [63]. These fields are also linked using the polarization $\mathbf{P}$ and magnetization $\mathbf{M}$, such that [63]

$$\mathbf{D} = \varepsilon_0 \mathbf{E} + \mathbf{P} \quad (5)$$

$$\mathbf{H} = \frac{1}{\mu_0} \mathbf{B} - \mathbf{M}, \quad (6)$$

where ε_0 is the electric permittivity and μ_0 is the magnetic permeability of vacuum. $\mathbf{P}$ is related to the internal charge density via $\nabla \cdot \mathbf{P} = -\rho$, and the charge conservation ($\nabla \cdot \mathbf{J} = -\partial\rho/\partial t$) requires the internal charge and current densities to be linked by $\mathbf{J} = \partial\mathbf{P}/\partial t$ [63]. In this approach, the macroscopic electric field accounts for all polarization effects (both external and induced fields) [63].

Note that the densities ρ and $\mathbf{J}$ may be thought of as the electromagnetic field sources. However, when away from the sources (in source-free regions of space), Maxwell's equations can be defined in a simpler form [50]:

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \quad (7)$$

$$\nabla \times \mathbf{H} = \frac{\partial \mathbf{D}}{\partial t} \quad (8)$$

$$\nabla \cdot \mathbf{D} = 0 \quad (9)$$

$$\nabla \cdot \mathbf{B} = 0. \quad (10)$$

To understand the phenomenon of wave propagation in left-handed media, first, it is necessary to reduce Maxwell's equations to the wave equation [31]

$$\left(\nabla^2 - \frac{n^2}{c^2} \frac{\partial^2}{\partial t^2} \right) \psi = 0, \quad (11)$$

where n is the index of refraction, c is the velocity of light in vacuum, and $n^2/c^2 = \epsilon_0 \mu_0$, ϵ_0 is the vacuum permittivity, and μ_0 is the vacuum permeability.

Now, consider Maxwell's equation in phasorial form (or frequency domain, assuming time dependence as $\exp(+j\omega t)$):

$$\nabla \times \mathbf{E} = -j\omega\mu\mathbf{H} \quad (12)$$

$$\nabla \times \mathbf{H} = j\omega\epsilon\mathbf{E}, \quad (13)$$

where ω is the angular frequency, μ is the relative permeability of the medium and ϵ is the relative permittivity of the medium.

Considering plane-wave fields of the kind $\mathbf{E} = \mathbf{E}_0 \exp(-j\mathbf{k} \cdot \mathbf{r} + j\omega t)$ and $\mathbf{H} = \mathbf{H}_0 \exp(-j\mathbf{k} \cdot \mathbf{r} + j\omega t)$, equations (12) and (13) reduce to:

$$\mathbf{k} \times \mathbf{E} = \omega\mu\mathbf{H} \quad (14)$$

$$\mathbf{k} \times \mathbf{H} = -\omega \varepsilon \mathbf{E}, \quad (15)$$

where $\mathbf{E}$ and $\mathbf{H}$ are both mutually perpendicular and perpendicular to the wave vector $\mathbf{k}$ [64].

It is known that the electromagnetic propagation through a material is directly related to the behavior of its ε and μ parameters. In conventional materials, when ε and μ are greater than zero, the propagation of electromagnetic waves occurs so that the right-hand rule is obeyed, and therefore, this type of material is known as right-handed materials (RHM). Thus, it can be inferred that for positive values of ε and μ in equations (14) and (15), the vectors $\mathbf{E}$, $\mathbf{H}$ and $\mathbf{k}$ form a right-handed orthogonal system, as can be observed in Figure 2-1 (a).

On the other hand, if ε and μ are both negative, equations (14) and (15) can be rewritten as

$$\mathbf{k} \times \mathbf{E} = -\omega |\mu| \mathbf{H} \quad (16)$$

$$\mathbf{k} \times \mathbf{H} = \omega |\varepsilon| \mathbf{E}. \quad (17)$$

Consequently, in a singular way, the right-hand rule may not be obeyed for metamaterials with $\varepsilon < 0$ and $\mu < 0$. Such a material, initially studied by Veselago [39], is conveniently called left-handed materials (LHM), where ε and μ are smaller than zero (DNG material).

In Figure 2-1, it can be verified the orientation of the electric field vector $\mathbf{E}$, the magnetic field $\mathbf{H}$, the Poynting vector $\mathbf{S}$, the wave vector $\mathbf{k}$, and the direction of these vectors for both RHM and LHM. For RHM, the refractive index is positive. The opposite occurs with LHM, where the refractive index is negative, with this phenomenon experimentally verified by Shelby et al. [38].

Besides Maxwell's equations, the constitutive relations are also important, once they give the relationship between the flux density of the vector $\mathbf{D}$ with the electric field $\mathbf{E}$, and the flux density of the vector $\mathbf{B}$ with the magnetic field $\mathbf{H}$ when a wave is propagating.

Figure 2-1: Orientation of the vectors $\mathbf{E}$, $\mathbf{H}$, $\mathbf{S}$ e $\mathbf{k}$ for (a) Right-Handed Material (RHM), and (b) Left-Handed Material (LHM).

In the case of an isotropic medium, the constitutive relations can be written as [64]

$$\mathbf{D} = \varepsilon \mathbf{E} \quad (18)$$

$$\mathbf{B} = \mu \mathbf{H}. \quad (19)$$

The relative permittivity (ε_r) and relative permeability (μ_r) are defined by [31]

$$\varepsilon_r = \frac{\varepsilon}{\varepsilon_0} \quad (20)$$

$$\mu_r = \frac{\mu}{\mu_0}, \quad (21)$$

Consequently, equations (18) and (19) become

$$\mathbf{D} = \varepsilon_0 \varepsilon_r \mathbf{E} \quad (22)$$

$$\mathbf{B} = \mu_0 \mu_r \mathbf{H}. \quad (23)$$

And the index of refraction of a material is defined by [38]

$$n = \pm \sqrt{\varepsilon_r \mu_r}. \quad (24)$$

It is important to acknowledge that Maxwell's equations do not preclude the possibility that both ϵ_r and μ_r are negative. Negative-index media, as already mentioned, have both ϵ_r and μ_r simultaneously negative. When $\epsilon_r < 0$ and $\mu_r < 0$, n must be defined by the negative square root

$$n = -\sqrt{\epsilon_r \mu_r}. \quad (25)$$

Thus, the condition $n < 0$ and $\mu_r < 0$ implies that the characteristic impedance (z) of the medium will be positive, meaning that the energy flux is in the same direction as the direction of propagation [50], as will be verified in the numerical results in Section 4.2.

In metamaterials literature, ϵ_r and μ_r are in fact defined as effective parameters obtained from extraction parameter procedures [45],[46],[65],[66] discussed and detailed in APPENDIX A.

2.1.1 NEGATIVE INDEX OF REFRACTION

To comprehend the negative index of refraction effect, it is considered an optical ray incident on the interface of two different materials: the ordinary media, presenting $\epsilon > 0$ e $\mu > 0$, and the LHM media. In this situation, the boundary conditions impose continuity of the tangential components of the wave vector at the interface. Unlike what happens with ordinary refraction, the incident and refraction angles must now present opposite signals, as presented graphically in Figure 2-2.

As can be seen in Figure 2-2, the Poynting vector $\mathbf{S}$ and the wave vector $\mathbf{k}$ of the ordinary media (subscript 1) follow Snell's Law, as expected. And it follows that, from the continuity of the tangential components of the wave vectors $\mathbf{k}$, the incident and refracted rays are

$$\frac{\sin \theta_1}{\sin \theta_2} = \frac{-|\mathbf{k}_2|}{|\mathbf{k}_1|} \equiv \frac{n_2}{n_1} < 0, \quad (26)$$

where n_1 is the index of refraction of the ordinary media and n_2 is the index of refraction of the left-handed media.

Figure 2-2: Graphical representation of an optical beam incident on an interface between an ordinary media (subscript 1) and a left-handed media (subscript 2).

According to equation (26), Snell's law is satisfied with a negative index of refraction that does not depend on the angle of incidence, making the negative refraction a unique property of isotropic left-handed media [3].

Despite many positive effects, negative refraction at the interface between ordinary and left-handed medium has been criticized due to its high dispersive nature [3]. However, nowadays negative refraction in LHM is well established, once both theoretical calculations [67] and experiments [38],[68] have confirmed Veselago's predictions.

2.2 METAMATERIALS DESIGN

In the design of engineered materials, there is the so called bulk artificial media or bulk metamaterials. They are discrete media made of a combination of unit cells of small electrical size at the frequency of interest [3].

A standard procedure was established after the works of Rotman [69], Pendry [70],[71], and Smith [37] to design bulk metamaterials with negative parameters at microwave frequencies. This procedure is based on the use of metallic wires and/or plates to obtain negative dielectric permittivity, and an additional system with SRRs (Split-Ring Resonators) to obtain negative magnetic permeability. Thus, the SRR, with the addition of metallic wires, was one of the first metamaterials known in the literature.

The steps for the construction of a SRR is shown in Figure 2-3, where in *a*) there is a

circular metal that has no magnetic properties. When a metal ring is created with this metal (in *b*), a current is induced by the magnetic field H , but in this case there is only a weak magnetic response. To introduce a resonance, a cut is made in the ring, as shown in *c*). To improve the resonance response, it is added a concentric ring within the first ring (in *d*) separated by a distance x .

Consequently, it became possible to develop a new material with the desired electromagnetic responses. Depending on the configuration and dimensions of these rings, different resonances and responses can be obtained, such as negative refractive index.

Figure 2-3: Steps for the construction of a SRR.

Usually, the metamaterial structures are arranged periodically, as presented in Figure 2-4.

Figure 2-4: Periodic arrangement of SRR structure. The unitary cell of this SRR is delimited by a dotted line.

In Figure 2-5, it can be seen four different categories of materials in the $\epsilon - \mu$ diagram, illustrating some examples of the types of materials and/or structures that can be used to obtain different electromagnetic properties. For example, to obtain a material with

$\epsilon < 0$ and $\mu > 0$, a property shown by most of the metals, one can use a wired medium.

The artificial materials (metamaterials) with $\epsilon < 0$ and $\mu < 0$, can be obtained using a medium composed of SRR's and metal wires. To obtain a material with $\epsilon > 0$ and $\mu < 0$ (as shown by some natural magnetic materials), one can use a medium with SRR's. Usually, natural (or ordinary) materials exhibit both ϵ and $\mu > 0$.

Figure 2-5: Example of metamaterials in accordance with the values of permittivity (ϵ) and permeability (μ).

It is emphasized that the model structures shown in Figure 2-5 are just simple examples of how to obtain different values of permittivity ϵ and permeability μ . For other more complex cell geometries both in the microwave and optical regime, please refer to [45]-[48],[51],[72]-[75].

The diagram shown in Figure 2-6 helps to better understand the behavior of the four types of materials (DPS, ENG, DNG, MNG) when an electromagnetic wave is incident. The orientation of the electric field vector $\mathbf{E}$, the magnetic field $\mathbf{H}$ and the wave vector $\mathbf{k}$ can be checked when the signs (positive or negative) of the permittivity (ϵ) and magnetic permeability (μ) are varied.

As can be observed, metamaterials have tremendously interesting properties, which is why they are being extensively used by the scientific community. However, besides the conventional metamaterials, there is a class known as chiral metamaterials that stands out, and it will be explained in the next subsection.

Figure 2-6: Behavior of an incident wave on the various types of existing materials (adapted from [76]).

2.3 CHIRAL METAMATERIALS

As discussed previously, chiral metamaterials have the ability of rotating the polarization of the field, which is a consequence of the chirality parameter κ (materials with $\kappa \neq 0$ are chiral). Interesting enough, this parameter alleviates the necessity of having the permittivity and permeability simultaneously negative to obtain negative refraction, as required by ordinary metamaterials. Thus, a chiral design can offer a more efficient way to obtain the negative refraction index with simpler metamaterial cells [45].

Chiral media belong to a wider range of bi-isotropic media and they are characterized by the following constitutive relations [46]

$$\mathbf{D} = \varepsilon_0 \varepsilon_r \mathbf{E} + (\chi + j\kappa) \sqrt{\varepsilon_0 \mu_0} \mathbf{H} \quad (27)$$

$$\mathbf{B} = \mu_0 \mu_r \mathbf{H} + (\chi - j\kappa) \sqrt{\varepsilon_0 \mu_0} \mathbf{E}, \quad (28)$$

where χ is the dimensionless magneto-electric parameter and describes the reciprocity of the material, c is the speed of light, and j is the imaginary number.

Considering a reciprocal medium, where $\chi = 0$, $\kappa \neq 0$, and remembering that $c = 1/\sqrt{\varepsilon_0\mu_0}$, and $j = \sqrt{-1}$, equations (27) and (28) become,

$$\mathbf{D} = \varepsilon_0\varepsilon_r\mathbf{E} + j\frac{\kappa}{c}\mathbf{H} \quad (29)$$

$$\mathbf{B} = \mu_0\mu_r\mathbf{H} - j\frac{\kappa}{c}\mathbf{E}. \quad (30)$$

Several authors have investigated the effects of chirality and discussed the improvements it can produce on different electromagnetic devices [45],[46]-[51]. The chirality parameter (κ) characterizes the strength of the cross-coupling between the electric and magnetic fields [51], i.e., characterizes the magnetoelectric coupling [77].

Considering the propagation of a plane wave in a reciprocal chiral medium, and combining equations (29) and (30) with the frequency-domain source-free Maxwell's equations, one can obtain the wave equation for the electric field $\mathbf{E}$ [46]:

$$\mathbf{k} \times (\mathbf{k} \times \mathbf{E}) = -k_0^2(\varepsilon\mu - \kappa^2)\mathbf{E} + 2j\kappa k_0(\mathbf{k} \times \mathbf{E}), \quad (31)$$

where $k_0 = \omega/c$ is the free-space wave number. Assuming that $\mathbf{k} = k\hat{z}$ is the wave vector propagating on the chiral metamaterial in z direction, equation (31) is simplified to [46]

$$k_{\pm} = k_0(\sqrt{\varepsilon\mu} \pm \kappa). \quad (32)$$

In left-circularly polarized (LCP) waves and in right-circularly polarized (RCP) waves, the refractive index of chiral metamaterials can be calculated by the following equation [46]

$$n_{\pm} = \sqrt{\varepsilon\mu} \pm \kappa, \quad (33)$$

where n_+ refers to RCP and n_- to LCP.

Looking at equation (33), it is remarkable that depending on the value of κ the

refractive index can be negative, even if the permittivity and permeability are both positive.

The effect of polarization azimuth rotation angle of elliptically polarized light occurs when the polarization plane of a linearly polarized wave is rotated after passing through a chiral medium. This effect is called optical activity and is characterized by [46],

$$\theta = \frac{1}{2} [\arg(T_{++}) - \arg(T_{--})], \quad (34)$$

where T_{++} is the transmission coefficients for RCP and T_{--} is the transmission coefficients for LCP.

Furthermore, there is the circular dichroism, an effect that causes a difference in the absorption and distortion of the RCP and LCP polarized waves. This effect is characterized by the ellipticity angle (η), which is defined as the difference in transmitted power of the two polarizations [46],[75]:

$$\eta = \tan^{-1} \left(\frac{|T_{++}| - |T_{--}|}{|T_{++}| + |T_{--}|} \right). \quad (35)$$

The angle θ is proportional to the chirality κ , and using the default settings of θ and η , it is possible to calculate changes in the polarization of a linearly polarized wave in chiral structures [45].

Once explained about metamaterials, we can introduce some important concepts about antennas.

CHAPTER 3

3 FUNDAMENTAL CONCEPTS OF ANTENNAS

In this chapter, fundamental concepts related to antennas, their basic operating characteristics and their main figure-of-merits will be addressed.

3.1 ANTENNAS

The research field of antennas is vigorous and dynamic, and over the years the technology of this ubiquitous device has been an indispensable partner of the communications revolution [43]. In a communications system, the antenna is a linear reciprocal passive device [78] that, at the transmission, has the function of converting electrical current into electromagnetic waves (radiant energy) and, at the reception, of collecting the wave and converting it into electrical current [79]. The antenna can also be defined as the structure associated with the transition region between a waveguide and the free space, or vice versa [80]. Figure 3-1 shows the antenna as a transition device, where the transmission line is used to carry electromagnetic energy from the transmitter to the antenna, or from the antenna to the receiver.

Ideally, the energy generated by the source should be fully transferred to the transition device. However, in practice there are losses due to the nature of the transmission line and the antenna itself. In addition, reflections from these devices occur, a phenomenon known as “impedance mismatch” [43].

One of the important factors when designing an antenna is finding ways to realize the impedance matching, where the input impedance of the antenna is equaled to the output impedance of the generator. It is often necessary to use impedance matching circuits, which are devices that maximize energy transfer between the transmission lines and the antenna.

Figure 3-1: Antenna (II) as a device of transition between transmission line (I) and free space (III) (adapted from [43]).

Some of the main characteristics desired to improve an antenna are:

- a) Radiation efficiency: the ratio of the total power radiated by an antenna to the net power accepted by the antenna from the connected transmitter [78]. In other words, the efficiency of an antenna is the ratio of the desired output power to the supplied input power [81]. The efficiency is mathematically defined as [43],[81]

$$\eta_{fr} = \frac{P_{rad}}{P_{in}}, \quad (36)$$

where P_{rad} is the power radiated by the antenna and P_{in} is the power supplied to the input of the antenna.

- b) Directivity: the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions [78]. It is the measure of the focusing ability of an antenna, defined as the ratio of the maximum radiation intensity in the main beam to the average radiation intensity over all space [81]. The directivity is mathematically defined as [43],[81]

$$D = \frac{U_{max}}{U_{avg}} = \frac{4\pi U_{max}}{P_{rad}}, \quad (37)$$

where U_{max} is the maximum radiation intensity in the main beam and U_{avg} is the average radiation intensity over all space.

This dimensionless power ratio parameter is usually expressed in dB as [81]

$$D(dB) = 10 \log(D). \quad (38)$$

- c) Gain: the ratio of radiation intensity in a given direction to the radiation intensity that would be obtained if the power accepted by the antenna would be radiated isotropically [78]. In other words, antenna gain is the product of directivity and efficiency and is written as [43],[81]:

$$G = \eta_{fr} D. \quad (39)$$

- d) Bandwidth: the range of frequencies within which the performance of the antenna with respect to some characteristic conforms to a specified standard [78].

Improvements are also found in other characteristics, such as the input impedance, the radiation pattern, polarization, radiation intensity, beam width, among others. Thus, the great potential in the use of metamaterials in antennas is noteworthy due to the wide range of possibilities for improvement.

3.1.1 ANTENNA TYPES

An antenna may have different geometries, from a simple wire to complex structures, such as fractal antennas. Because of these different geometries, the antennas can be built with distinct specifications to suit certain applications. The optimum type of antenna for a given situation may depend, for example, on the distance to be covered and the frequency to be used.

The most common types of antennas known are the wire antennas, which may be, for example, monopole, dipole, helical, and loop. There are also aperture antennas, which are

widely used in aerospace applications, such as horn antennas and certain waveguides. Other type of antenna that became popularized since the 70s is the microstrip antenna (or patch antenna) [9]. Because of its easy construction (using Printed Circuit Boards - PCB), the microstrip antenna is widely used today in both government and commercial applications. Besides these, there is an extensive range of antennas, such as reconfigurable, fractals, intelligent/adaptive, spiral, log-periodic, among others that arise as technology advances and as the demand is required.

When one antenna does not achieve the desired characteristics, it is possible to construct an array of radiating elements that are repetitions of this antenna. Thus, the desired properties can be obtained using these arrays. This type of structure is known as “antenna array” where elements can be arranged in a variety of geometric configurations. It is important to note that the fields of the individual elements can be added constructively in some directions and destructively in others, so the project to be developed must be analyzed carefully.

For this project, it was chosen the monopole antenna. Besides being a simple antenna from a manufacturing point of view, it is one of the most used in communications systems, broadcasting and on measurements of electric and magnetic fields [82]. The monopole has several applications, such as car radios, mobile communications (like walk-talk radio), medical imaging, including in certain cancer treatments [82].

3.2 ANTENNAS AND METAMATERIALS

As previously discussed, the possibility of improving the performance of antennas (or any other electromagnetic device) is virtually endless, particularly since the advent of metamaterials. Antennas have become one of the most exciting applications of metamaterials due to the possibility of significantly improving their performance and, for this reason, many papers [7]-[23], just to mention a few, have been published since the boom of metamaterials in 2000 [3]. In these papers, authors have demonstrated both numerically and experimentally how it is possible to obtain improvements with the use of metamaterials.

For example, in 2002, Enoch et al. [19] published a work showing that it is possible to control the directivity of an antenna by adding a metamaterial medium. The periodic structures formed by metal grids are considered as the metamaterial medium and a monopole

antenna is inserted between these structures. With the proper excitation conditions, the energy radiated by the antenna can be concentrated in such a way that it produces a narrow directivity.

Using this same type of metal grid (also known as fishnet), in 2006 a group [20] demonstrated that it is possible to improve the gain of a circular waveguide antenna.

Metamaterial structures can be useful to introduce changes in antenna impedance, therefore obtaining an improvement in antenna response [3]. In [7], the authors showed that it is possible to match an electrically small dipole to the free-space impedance using a metamaterial spherical cover with negative index of refraction. In the structure proposed in [7], an increase in the radiated power and a reduction of the reactance could be obtained simultaneously.

In 2006, it was presented by Ziolkowski and Erentok [8] the model of electrically small antennas (ESA), such as dipole and monopole, covered by a shell of homogeneous and isotropic materials with negative permittivity ϵ (ENG material). The homogeneous materials are an artifice to facilitate the simulations. In Figure 3-2, it can be seen both types of antennas.

The inductive nature of the ENG material was used to compensate the capacitive nature of the ESA to form a resonant system. It has been shown that such systems can actually be resonant with a great improvement in radiation power, in comparison with the antenna itself in free space.

Figure 3-2: Model of the antennas covered by an ENG material – a) Dipole antenna; b) Monopole antenna (adapted from [8]).

Using analytical results, it was also demonstrated that the inclusion of dispersive materials in the properties of the shell with ENG material posed a significant impact on the bandwidth of the dipole antenna, but not in other characteristics, such as the total peak power or the radiation diagrams. To Ziolkowski and Erentok [8], the challenge is to try to minimize the effects of dispersion and losses of the ENG material in the overall system, a subject that continues to be investigated in the present days.

In 2007, Alù et al. [83] analyzed the radiation properties of microstrip antennas, modeled with dimensions smaller than the wavelength (*subwavelength*), with blocks of homogeneous metamaterial, such as DNG, DPS and SNG (Single Negative). According to the authors [83], the rectangular patch antenna did not provide efficient radiation, therefore, this is not a valid option for the purpose of designing an efficient subwavelength radiator. For this reason, the same principle was tested with a circular microstrip antenna with blocks of homogeneous metamaterial. Alù and his group [83] demonstrated that the circular geometry provides additional degrees of freedom for selecting the proper mode of operation in order to get an efficient radiation from a subwavelength radiator loaded with metamaterials.

Using the same principle presented in [83], Bilotti et al. [84] proposed a circular microstrip antenna with ideal homogeneous metamaterials MNG and DPS, as shown in Figure 3-3.

Afterwards, magnetic inclusions were added inside the system. It was an attempt to make the DPS and MNG metamaterials real because, previously, blocks of homogeneous metamaterial were often used to facilitate the numerical analysis.

Figure 3-3: Circular microstrip antenna embedded in blocks of MNG and DPS metamaterial (adapted from [84]).

Presented in [84], the numerical results of radiation efficiency and gain in the first resonant frequency (470 MHz) are 0.67 and 3.4 dBi, respectively. For the second resonance frequency (2.44 GHz), the efficiency is 0.92 and the gain is 6.5 dBi. The authors also

presented adequate comparisons with analytical results and numerical simulations assuming an ideal isotropic MNG metamaterial. This confirmed the capacity of practical implementation and effectiveness of the proposed approach for the realization of microstrip antennas with metamaterials showing a good performance.

The cases mentioned above are only examples showing that it is indeed possible to use metamaterials to improve the performance characteristics of different types of antennas. Dong and Itoh [85] presented a detailed review of the current research related to antennas based on metamaterials, mainly to electrically small antennas (ESA). Some difficulties and limitations for the development of this technology are presented and possible approaches to solve these problems are pointed out.

As can be noted, it is common to use blocks of homogeneous material to emulate the characteristics of metamaterials to facilitate the numerical analysis, but it is not a real system that can be readily used. To characterize a metamaterial, it is needed to find a geometry that has the same properties of the homogeneous material. It is noteworthy that the idea of this work is to use a medium composed of dielectric and metal, forming a metamaterial that is physically realizable and ready for use. Therefore, it will be possible to manufacture and experimentally test these structures. But first, it is essential to perform the numerical analysis of all structures to, thereby, make the manufacturing feasible.

CHAPTER 4

4 NUMERICAL ANALYSIS

In this chapter, it is explained how the numerical analysis was performed with a commercial software, the boundary conditions used, and all necessary parameters to simulate the structures.

4.1 NUMERICAL SIMULATION

The *High Frequency Structure Simulator* (HFSS) software from *Ansys* was used to perform all numerical simulations of the structures involved in this project. HFSS is a commercial software based on the finite element method (FEM) where the problem-modeling is performed in three dimensions (3D). This software generates a mesh to efficiently and accurately simulate and solve numerically a variety of structures, such as antennas, microwave and radio frequency (RF) components, and even high speed components, such as embedded on-chip passive devices. With HFSS it is possible to extract matrix parameters, such as the scattering parameters (S-parameters), to visualize electromagnetic fields, and different results for the analysis of each problem.

4.1.1 MONOPOLE ANTENNA

As mentioned before, the monopole antenna was chosen for this work. The monopole is a standard antenna widely used in telecommunications, mainly due to the simplicity in its design and construction.

Combined with HFSS, it is used an extension package from *Ansys* called HFSS Antenna Design Kit (ADK), which uses the antenna's equations available in Balanis [43]. With this kit, modeling the antenna becomes faster, because there is a database used to generate a structure based on the operating frequency chosen.

From the intrinsic equations of this software, one obtains some values, such as the antenna length, the radius, the width of the ground plane and also the correct size of the geometry used to simulate a coaxial port on the system.

In Figure 4-1 it is possible to visualize a model of the monopole antenna on a finite ground plane and the variables of the geometric dimensions that are calculated by ADK. For this kind of antenna, it was used an excitation called lumped port. In HFSS, the lumped port represents an inner surface in which a signal enters or leaves the modeled geometry.

Figure 4-1: Model of the monopole antenna on a finite ground plane using ADK, showing the antenna length (l) and the width of the ground plane (w_{gp}). The detail shows the monopole antenna radius (r) and the feed gap of the lumped port.

Lumped ports are similar to traditional wave ports, but can be located internally and have a complex impedance that can be defined by the user with the desirable values. The complex impedance of the antenna is defined by $Z = 25 \, \Omega$, i.e., the resistance value is $25 \, \Omega$, and the value of reactance is $0 \, \Omega$. This value was defined based on the assumption that the radiation resistance of the monopole antenna can be modeled using the method of images, such as a dipole antenna with half the input impedance [86]. It is from the lumped port that it is possible to compute S-parameters directly at the port. This type of port can be modeled as a rectangle, as shown in the detail of Figure 4-1 with the element called “feed gap”.

Figure 4-2 shows the monopole antenna at a side view, containing the reference plane used, the antenna length (l), the width of the ground plane (w_{gp}), and the radius of the conducting wire (r).

Figure 4-2: Monopole antenna (side view).

The operating frequency $f_0 = 8$ GHz was chosen because it is a frequency that can be used for various telecommunication services [87] and also because, with the real size of the antenna in this frequency, it is simple to fabricate and to perform experimental tests. The dimensions used are listed in Table 4.1.

Table 4.1: Geometric parameters of the monopole antenna at 8 GHz.

Monopole Antenna 8 GHz	Size (mm)
Antenna length (l)	8.15
Antenna radius (r)	0.251
Width of the ground plane (w_{gp})	25.11

The material selected for the antenna itself (monopole) and the ground plane was the perfect electric conductor (PEC). To numerically evaluate the electromagnetic fields in the far-field region of the antenna, it must be defined a region surrounding the radiating object considered to be "endless". To do so, two air boxes were created in the simulations. One of the air boxes has dimensions $(x, y, z) \lambda_0/0.56 \times \lambda_0/0.56 \times \lambda_0/1.2$, remembering that $\lambda_0 = c/f_0$, where c is the speed of light in vacuum and f_0 is the operating frequency. In this box are selected faces with the condition of "infinite radiation" to calculate the far-field parameters. Another air box, larger, has dimensions $(x, y, z) \lambda_0/0.44 \times \lambda_0/0.44 \times \lambda_0/0.77$ where it is possible to obtain the scattering parameters and the input impedance. It is noteworthy that the use of these two air boxes is recommended by the software manufacturer, once it is possible to analyze several information regarding the structure in all field regions [88].

The radiation boundary condition is used to simulate an open problem (i.e., that allows electromagnetic waves to radiate infinitely in space without reflection back to the

computational window). The radiation surface does not need to be spherical, but usually it has to be located at least a quarter of the wavelength of the radiation source, in this case, the antenna.

It is in the far-field region, also known as the Fraunhofer region, that the radiation pattern does not change with distance, but the fields are still evanescent with the ratio $1/R$ and the density of energy with $1/R^2$. The equations used to respect the Fraunhofer region are given by [89]:

$$R > \frac{2D^2}{\lambda} \quad (40)$$

$$R \gg D \quad (41)$$

$$R \gg \lambda \quad (42)$$

where D is the maximum linear dimension that contains the antenna and R is the separation between the antenna and a region in space. Equations (40) and (41) help ensure the fields in the region of far-field behave as plane waves. Equation (42) ensures there is only radiating fields (which decays to $1/R$).

For the sake of clarity, Figure 4-3 presents the diagram of the regions of an antenna.

Figure 4-3: Far-field, radiating near-field, and reactive near-field regions of an antenna (adapted from [43]).

Tests considering the Fraunhofer region were performed, but intensive computational resources and high computational costs were needed. For this reason, the air boxes have been minimized to achieve approximately the same responses that would be obtained considering a much larger area.

The simulations for the monopole antenna were performed to verify if it would operate on the designed frequency with acceptable values of the reflection parameter S_{11} (also known as Return Loss), gain and efficiency.

In Figure 4-4 a) the results of the reflection parameter S_{11} are shown, and in Figure 4-4 b), the gain is shown, both varying as a function of frequency in a range from 6 GHz to 12 GHz. The resonant frequency is $f_r = 7.99$ GHz, the gain in this frequency is $G_{f_r} = 1.65$ dB, and the efficiency is approximately $\eta_{f_r} \approx 1$. In the operational limit of -10 dB, ranging from 7.61 GHz to 8.48 GHz, the antenna has a bandwidth of 870 MHz. The gain at 8 GHz is 1.66 dB.

Figure 4-4: Simulation results for a monopole antenna– a) Return loss, S_{11} (dB); b) Antenna gain (dB).

After verifying that the standard antenna is operating correctly, it was chosen a different size for the ground plane to support and to be compatible with the new metamaterial-based structure to be coupled to the antenna. For this purpose, the width $w_{gp} = 8/5 \lambda_0 = 60$ mm (see Figure 4-2) is used, so it is possible to vary the distance of the new structure in relation to the radiating element (antenna) as a function of λ_0 , as will be seen later. The new dimensions of the antenna are listed in Table 4.2.

Table 4.2: Dimensions of the monopole antenna at 8 GHz with modified ground plane.

Monopole Antenna 8 GHz	Size (mm)
Antenna length (l)	8.15
Antenna radius (r)	0.251
Width of the ground plane (w_{gp})	60

Therefore, the antenna was simulated again with the modified ground plane. The results are close to those found with the first dimensions set, as shown in Figure 4-5.

Figure 4-5: Simulation results for a monopole antenna with modified ground plane – a) Return loss, S_{11} (dB); b) Antenna gain (dB).

As can be seen, Figure 4-5 *a*) shows the result of the return loss (reflection parameter) S_{11} and the gain in Figure 4-5 *b*). The resonant frequency is $f_r = 8.24$ GHz, the gain in this frequency is $G_{f_r} = 3.8$ dB and the efficiency is approximately $\eta_{f_r} \approx 1$. The antenna has a bandwidth of 920 MHz ranging from 7.8 GHz to 8.72 GHz, and the gain at 8 GHz is 4.2 dB.

Comparing the two structures, it is observed that with the modified ground plane the gain was increased by 2.54 dB at the operating frequency (8 GHz) and that there was also an increase of 50 MHz in the bandwidth. Consequently, this adapted structure can be used without modifying the main antenna responses.

Finally, with the monopole antenna designed, it is possible to design and to analyze the metamaterials cells.

4.1.2 METAMATERIAL CELLS

All metamaterials cells were modeled using HFSS. For conventional metamaterials, well-known structures in the literature were chosen for the sake of comparison. First, two types of conventional metamaterial cells were chosen, the conventional SRR [66] and the Omega (Ω) structure, as shown in Figure 4-6 *a*) and *b*), respectively.

Figure 4-6: Conventional metamaterials cells – a) Split Ring Resonator; b) Omega structure.

We also analyzed modified SRR cells, the single SRR and the double SRR, as shown in Figure 4-7 *a*) and *b*), respectively.

Figure 4-7: Conventional metamaterials cells – a) Split Ring Resonator; b) Omega structure.

For chiral metamaterials, the first cell chosen was based on the crossed wires structure introduced by Zhou et al. [45], shown in Figure 4-8 a). A cell developed by the Telecommunications Group at USP São Carlos [3] was also used, showing excellent results even for biosensors applications, as shown in Figure 4-8 b).

Figure 4-8: Chiral metamaterial cells – a) Cross-wired [45]; b) Curve-wired [3].

As can be observed, these metamaterials are designed to use a double metal layer, where the metallic inclusions are arranged on both sides of the dielectric substrate. However, there is a type of chiral metamaterial designed to use only one side of the metal layer of the substrate, facilitating their fabrication. These chiral metamaterials, that maintain the characteristics of unit cells without symmetry plane, are known as chiral metamaterials in two dimensions (2D). In 2009, Plum et al. [90] demonstrated a chiral metamaterial cell in 2D, as shown in Figure 4-9. This planar structure has asymmetry in both arches and gaps, which results in a 2D chiral metamaterial.

Figure 4-9: 2D chiral metamaterial cell.

Once selected the metamaterials cells, their analysis can be started. Before that, let us mention an important aspect when defining the dimensions of a metamaterial cell (either conventional or chiral metamaterial). To identify the optimal dimensions of each cell, it is possible to realize parametric simulations, and using HFSS with parametric analysis, one can choose one or more variables so their values are changed within a predefined range. Within this range, the starting and ending values have to be provided, as well as the step size. The software performs this analysis by modifying the values of the selected variable in each iteration. Subsequently, one can examine the results obtained with each modified value. This is a practical way to change and see the response pattern of a structure.

After defining the optimal size of each cell, each one must be analyzed numerically and, then, their scattering parameters can be extracted. The scattering parameters, commonly known as S-parameters, describe the response of a n -ports network to the input signal in each port analyzed. They can be obtained either by numerical or experimental analysis, this last one using a Network Analyzer. Additional information regarding the scattering parameters, their equations, and some definitions using the S-parameters can be found in APPENDIX B.

These parameters depend on the frequency of the electromagnetic field and are complex coefficients, i.e., they have real and imaginary parts. After obtaining these parameters, it is possible to extract other important data for the analysis of metamaterial cells, such as electrical permittivity (ϵ), magnetic permeability (μ), refractive index (n), and impedance (z). These may be extracted using specific methods, such as those proposed by Smith et al. [66].

The parameter retrieval or parameter extraction is an important technique to characterize the electromagnetic properties of the effective media. It is a procedure for obtaining the macroscopic parameters of a medium based on the transmission and reflection

coefficients of the S-parameters from a planar slab of this medium [46]. This technique is used with both numerical and experimental methods to guide the design of new metamaterials and identify the negative refractive behavior of metamaterials [46].

To extract the parameters from the metamaterials, first we obtain the S-parameters using HFSS commercial software. After that, we use a method described in [46] and [66] to extract (or retrieve) other essential parameters for the analysis of metamaterials. Additional information regarding the parameter retrieval, the equations used for conventional metamaterials and chiral metamaterials, can be found in APPENDIX A.

4.2 CHARACTERIZATION AND PARAMETER RETRIEVAL OF METAMATERIAL CELLS

After defining the cell geometries, materials, excitation ports, and boundary conditions, we now proceed to the simulation of the structures to obtain the S-parameters, and finally, extract the basic electromagnetic parameters of the metamaterials.

4.2.1 SPLIT RING RESONATOR (SRR)

The first metamaterial cell numerically analyzed in this work is the symmetric SRR, proposed by Smith et al. [66], as shown in Figure 4-10. This cell was chosen because it is well known in the literature, and it is a benchmark for the numerical analysis and the parameter retrieval used with the conventional metamaterials. The SRR is analyzed in a frequency range of 0.1 GHz to 20 GHz, with steps of 0.01 GHz.

The unit cell is cubic, with lateral dimensions $w_u = 2.5$ mm. The substrate is FR4 ($\epsilon_r = 4.4$ and loss tangent $\delta = 0.02$) with thickness $w = 0.25$ mm. The metal on both sides of the FR4 plate is copper, but on the numerical analysis it is used a PEC with zero thickness. This is a plausible approximation due to the low skin effect, since the wavelength in microwave frequencies is very large compared to the small thickness of the metal in the substrate.

The width of the metal rings is $w_c = 0.2$ mm, the length of the largest SRR ring is $w_{i1} = 2.2$ mm, and the length of the smaller ring is $w_{i2} = 1.5$ mm. The gap in each ring is $g = 0.3$ mm and the metal strip has width is $w_{ce} = 0.14$ mm.

Figure 4-10: Unit cell of the symmetric SRR. The vertical metal strip in the center has width w_{ce} and is on the opposite face of the FR4 substrate.

The S-parameters results both in magnitude and phase for the SRR cell can be seen in Figure 4-11.

Figure 4-11: S-parameters obtained for the SRR cell – a) Magnitude; b) Phase (in radians).

Observing the phase of S , it can be seen that there is a dip in S_{21} around 9 GHz, indicating the presence of a band with negative refractive index, as will be proved later on using the parameter retrieval method. The structure proposed by Smith et al. [66] is designed to be roughly impedance matched, where the refractive index $n = -1$, because this condition can be useful in some applications, such as the perfect lens proposed by Pendry [28].

The retrieved index in Figure 4-12 a) confirms the negative refractive index band that lies approximately between 9 and 12 GHz. These statements can be graphically visualized in Figure 4-12, which shows the retrieved index of refraction n , impedance z , relative permittivity ϵ , and relative permeability μ .

Comparing our numerical results with the results shown in Smith's work [66], we can assert there is a very good agreement between them, indicating that the parameter retrieval method can indeed be used for other metamaterial cells to be analyzed.

Figure 4-12: Electromagnetic response of the SRR – a) Index of refraction (n); b) Impedance (z); c) Relative permittivity (ϵ); d) Relative permeability (μ).

Nevertheless, it is important to note that both the refractive index and the impedance must satisfy the passivity requirements, where the imaginary part of n and the real part of z are greater or equal to zero. More details about the parameter retrieval of both n and z can be found in APPENDIX A.

However, the imaginary part of the permittivity, as presented in Figure 4-12 c), is negative, in contrast with the passivity condition. Also, the real part of ϵ presents an anti-resonant behavior, which is a nonphysical feature. In metamaterial retrieval procedures, frequency bands in which one of the two retrieved parameters (ϵ or μ) experiences an anti-resonant response with negative slope and negative imaginary part are common [91].

These anti-resonance artifacts are caused by a weak form of spatial dispersion effects associated with the finite phase velocity along the metamaterial array [91]. This effect is discussed in more details in [92], where the authors suggest this is a manifestation of a frequency dispersion of retrieved metamaterial parameters whose constitutive elements have resonances caused by the applied electromagnetic field. This kind of behavior has been found in many different metamaterial structures [44],[46],[66],[92]-[98], to mention a few, when the effective material parameters are extracted from simulated (or measured) plane-wave reflection and transmission coefficients using the Nicolson-Ross-Weir retrieval procedure [92]. Usually, either ϵ or μ exhibits the anti-resonance behavior, while the other one demonstrates the usual resonance of Lorentz's type [92], as can be observed in Figure 4-12 d)

for the real part of the permeability.

Generally, the frequency range for the resonance and anti-resonance coincides or overlaps, as can be observed in Figure 4-12 *c*) and *d*). In this case, the anti-resonance behavior of the permittivity (real part of ϵ , solid blue line in Figure 4-12 *c*) occurs close to the resonant frequency (around 9 GHz) of the resonance of the permeability (real part of μ , solid blue line in Figure 4-12 *d*). This kind of behavior will also be observed in the results discussed in Sections 4.2.2 and 4.2.3.

4.2.2 MODIFIED SRR

We simulate two different SRR structures, one without the wire on the back of the substrate and another one with the rings on both sides (back and front), named single SRR cell and double SRR cell, respectively. From the parametric simulation results, the cell size was defined with width $w_u = 5$ mm for both cell types. The copper width is $w_c = 0.5$ mm, gap $g = 0.5$ mm, major SRR height $w_{i1} = 4$ mm and smaller SRR $w_{i2} = 2$ mm. For these cells, we used an alumina substrate ($\epsilon_r = 9.4$ and $\tan\delta = 0.006$) with 0.7 mm of thickness, which is the one we have available in our Laboratory. More details about the geometry of this cell can be seen in Figure 4-13.

Figure 4-13: Unit cell of the modified SRR structure (single SRR).

The results are given in a frequency varying from 2 GHz to 8 GHz, where the resonant frequency of this metamaterial cell is located. The S-parameter results both in magnitude and phase for the single SRR can be seen in Figure 4-14.

Figure 4-14: S-parameter results for the single SRR cell – a) Magnitude; b) Phase (in radians).

Figure 4-15 presents the index of refraction n , impedance z , relative permittivity ϵ and relative permeability μ obtained using the adopted parameter retrieval method.

Figure 4-15: Electromagnetic response of single SRR cell – a) Index of refraction (n); b) Impedance (z), c) Relative permittivity (ϵ); d) Relative permeability (μ).

For the single SRR, we can observe an anti-resonance behavior of the permittivity (real part of ϵ , solid blue line in Figure 4-15 c) occurs close to the resonant frequency (around 4.5 GHz) of the resonance of the permeability (real part of μ , solid blue line in Figure 4-15 d). We can also observe that the imaginary part of n (Figure 4-15 a) and the real part of z (Figure 4-15 b) are in accordance with the passivity condition.

With the previous results, it can be recognized this is a MNG structure, since the real part of the permeability is negative from 4.5 GHz to 5 GHz, while its permittivity is greater than zero.

Now, the S-parameter results both in magnitude and phase for the double SRR is presented in Figure 4-16.

Figure 4-16: S-parameter results for the double SRR cell – a) Magnitude; b) Phase (in radians).

Figure 4-17 presents the index of refraction n , impedance z , relative permittivity ϵ and relative permeability μ obtained with the adopted parameter retrieval method.

Figure 4-17: Electromagnetic response of double SRR – a) Index of refraction (n); b) Impedance (z); c) Relative permittivity (ϵ); d) Relative permeability (μ).

For the double SRR, we can again observe that the anti-resonance behavior of the permittivity (real part of ϵ , solid blue line in Figure 4-17 c) occurs close to the resonant frequency (around 4.5 GHz) of the permeability resonance (real part of μ , solid blue line in Figure 4-17 d). Moreover, the imaginary part of n (Figure 4-17 a) and the real part of z (Figure 4-17 b) are also in accordance with the passivity condition.

The results obtained with the double SRR cells are very similar to those obtained with the single SRR. The double SRR is also a MNG structure, since the real part of the permeability is negative from 4.4 GHz to 5 GHz, while its permittivity is greater than zero.

4.2.3 OMEGA

For the Omega structure, the dimensions are as following: height $w_{u1} = 4.57$ mm, width $w_{u2} = 5.3$ mm, width of the copper $w_c = 0.53$ mm, gap $g = 0.53$ mm and internal diameter $d_i = 2.65$ mm. The substrate used was alumina ($\epsilon_r = 9.4$ and $\tan\delta = 0.006$) with 0.7 mm of thickness. Further details of the cell geometry are presented in Figure 4-18.

Figure 4-18: Unit cell of the Omega structure. This figure shows one omega shaped inclusion on each side of the substrate.

The S-parameter results, both in magnitude and phase, within the frequency range of 6 GHz to 9 GHz can be seen in Figure 4-19.

Figure 4-19: S-parameters – a) Magnitude; b) Phase (in radians).

The retrieved index of refraction, impedance, relative permittivity and relative permeability are presented in Figure 4-20.

Figure 4-20: Retrieved parameters of the omega cell – a) Index of refraction n ; b) Impedance z ; c) Relative permittivity ϵ ; d) Relative permeability μ .

As can be observed, the refractive index has a negative value from 7.6 GHz to 7.95 GHz, and it has a near zero value in almost all the frequency range. At approximately 7.8 GHz the refractive index presents a negative value $n = -1$. The retrieved impedance shows that the structure is roughly matched at approximately 7.8 GHz, where the imaginary part of z crosses zero. The real part of the permittivity assumes a negative value between approximately 6 GHz and 7.9 GHz, where it crosses zero. It is possible to see a resonant frequency (around 7.7 GHz) of the permeability (real part of μ , solid blue line in Figure 4-20 d) and the real part of the permeability is also negative from 7.8 GHz to approximately 8.5 GHz. The imaginary part of n (Figure 4-20 a) and the real part of z (Figure 4-20 b) are also in accordance with the passivity condition. Therefore, the Omega cell is considered a metamaterial of the DNG type, within the frequency range where $\epsilon < 0$ and $\mu < 0$.

4.2.4 CROSS-WIRED (CHIRAL METAMATERIAL)

For the cross-wired structure, adapted from [45], it was chosen two cells. The first has width $w_u = 3.75$ mm, copper width $w_c = 0.29$ mm, wire length $l_c = 3.175$ mm, offset angle $\varphi = 30^\circ$ and tilting angle $\varphi_0 = 45^\circ$. The second structure has width $w_u = 7.50$ mm, copper width $w_c = 0.29$ mm, wire length $l_c = 6.925$ mm, offset angle $\varphi = 30^\circ$ and tilting angle $\varphi_0 = 45^\circ$. The substrate used was alumina ($\epsilon_r = 9.4$ and $\tan\delta = 0.006$) with 0.7 mm of

thickness. More details regarding this cell geometry can be seen in Figure 4-21.

Figure 4-21: Unit cell of the cross-wired structure. Each cross-wired structure is defined on opposite sides of the substrate with an offset angle ϕ between them.

The results for the cross-wired structure with width $w_u = 3.75$ mm and frequency ranging from 15 GHz to 25 GHz (which is the range where the resonant frequency of this chiral metamaterial cell is located) is shown in Figure 4-22. This figure shows the linear transmission coefficients (T_{xx} , T_{yy} , T_{xy} and T_{yx}) for the cross-wired structure with width $w_u = 3.75$ mm.

Figure 4-22: Linear transmission coefficients for the cross-wired structure with width $w_u = 3.75$ mm – a) T_{xx} and T_{yy} coefficients (dB); b) T_{xy} and T_{yx} coefficients (dB).

Now, with the results obtained for the linear transmission, we can calculate the circular transmission coefficients (T_{++} for RCP, and T_{--} for LCP), both in dB and in phase, as presented in Figure 4-23.

Due to the asymmetric geometry of this cell along the propagating direction, the transmission responses for RCP (blue line) and LCP (red line) split into two curves, as can be seen in Figure 4-23 a).

Figure 4-23: Circular transmission coefficients for the cross-wired structure with width $w_u = 3.75$ mm – a) T_{++} (RCP) and T_{--} (LCP) coefficients (dB); b) Phase (degrees) of T_{++} (RCP) and T_{--} (LCP).

It can be noticed two resonance dips at frequencies 19.1 GHz and 23.1 GHz. For the first resonance (19.1 GHz), the transmission dip for RCP and LCP are almost the same, but for the RCP this dip is more pronounced, implying the resonance for RCP is stronger than LCP one. The opposite occurs with the second resonance (23.1 GHz), where it can be observed a stronger resonance for the LCP.

Using the standard definitions of polarization azimuth rotation and ellipticity, we can calculate the polarization changes in a linearly polarized wave incident on the cross-wired structures. Figure 4-24 presents the results for the azimuth angle θ and for the ellipticity η .

Figure 4-24: Results for the cross-wired structure with width $w_u = 3.75$ mm – a) Azimuth angle θ (degrees); b) Ellipticity angle η (degrees).

At the two resonance frequencies, the azimuth rotation and ellipticity reach their maximum values ($\theta = -157.7^\circ$, $\eta = -6.9^\circ$ in the first resonance, and $\theta = -78.1^\circ$, $\eta = 6.7^\circ$ in the second resonance) as observed in Figure 4-24 a) and b), respectively. In the region between the two resonance peaks (around 21 GHz), which is also the region with low loss and nearly zero dichroism, it is possible to observe a polarization rotation of -24° with $\eta \approx 0$. The sign change of η around 21 GHz reflects the different frequency dependence between the magnitude of the transmission $|T_{++}|$ and $|T_{--}|$. Consequently, linearly polarized incident waves

with frequencies below and above 21 GHz will exhibit different handedness (either right or left) after exiting this structure.

Figure 4-25 presents the chirality parameter κ , the refractive indices for RCP n_+ and for LCP n_- , and the conventional refraction index n . They were all extracted from simulation results using the adopted retrieval procedure.

Figure 4-25: Results for the cross-wired structure with width $w_u = 3.75$ mm – a) Real part of chirality parameter κ (dimensionless); b) Real part of the refractive index for n_+ (RCP), n_- (RCP) and n (dimensionless).

The chirality parameter κ in Figure 4-25 a) shows two resonances, one at 18.2 GHz and another at 22.4 GHz. Above the first resonance frequency, κ is negative between 18.2 GHz and 22.4 GHz, resulting in $n_+ = \sqrt{\epsilon\mu} + \kappa < 0$. Above the second resonance frequency, κ is positive between 22.4 to 25 GHz, resulting in $n_- = \sqrt{\epsilon\mu} - \kappa < 0$.

It can be noticed in Figure 4-25 b) that n (green line) is positive through almost the entire frequency range, except for the resonance frequencies, where it achieves values of -0.2 for the first resonance and -0.4 for the second resonance. However, n_+ (blue line) is negative from a range of 18.2 to 20.7 GHz, and n_- (red line) has a negative region from 22.5 GHz to approximately 25 GHz. With this in mind, we can affirm the negative refractive index for RCP and LCP originates from a strong chirality.

Now, we show the results for the cross-wired structure with width $w_u = 7.5$ mm at a frequency ranging from 5 GHz to 15 GHz, range within which the resonant frequency of this chiral metamaterial cell is located. The linear transmission coefficients (T_{xx} , T_{yy} , T_{xy} and T_{yx}) for the crossed wires structure with width $w_u = 7.5$ mm are shown in Figure 4-26.

Figure 4-26: Linear transmission coefficients for the cross-wired structure with width $w_u = 7.5$ mm – a) T_{xx} and T_{yy} coefficients (dB); b) T_{xy} and T_{yx} coefficients (dB).

With the results obtained for the linear transmission, we calculate the circular transmission coefficients (T_{++} for RCP, and T_{--} for LCP), both in dB and in phase, as shown in Figure 4-27. The transmission responses for RCP (blue line) and LCP (red line) split into two curves, as can be observed in Figure 4-27 a). This is a characteristic of this cell, as also observed in Figure 4-24 for width $w_u = 3.75$ mm. It can be noticed two resonances at frequencies 9.4 GHz and 11 GHz. At the first resonance (9.4 GHz), the values for RCP and LCP are almost the same, but it is more pronounced for RCP than for LCP, implying a stronger resonance for the RCP case. At the second resonance (11 GHz), on the other hand, the resonance for LCP becomes stronger.

Figure 4-27: Circular transmission coefficients for the cross-wired structure with width $w_u = 7.5$ mm – a) T_{++} (RCP) and T_{--} (LCP) coefficients (dB); b) Phase (degrees) of T_{++} (RCP) and T_{--} (LCP).

Figure 4-28 shows the results for the azimuth angle θ and ellipticity η . At the two resonance frequencies, the azimuth rotation and ellipticity reach their maximum values ($\theta = -120.1^\circ$, $\eta = -9^\circ$ at the first resonance, and $\theta = -85.5^\circ$, $\eta = 6.3^\circ$ at the second resonance) as observed in Figure 4-28 a) and b), respectively. In the region between the two resonance peaks (around 10.2 GHz), we observe a polarization rotation of -30.5° with $\eta \approx 0$. The sign change of η around 10.2 GHz reflects the different frequency dependence between the magnitude of the transmission $|T_{++}|$ and $|T_{--}|$. Thus, linearly polarized incident waves with

frequencies below and above 10.2 GHz will exhibit different handedness (right or left) when exiting this structure.

Figure 4-28: Results for the cross-wired structure with width $w_u = 7.5$ mm – a) Azimuth angle θ (degrees); b) Ellipticity angle η (degrees).

Figure 4-29 shows the chirality parameter κ , refractive index for RCP n_+ and for LCP n_- , and the conventional refraction index n . They were all extracted from simulation results using the adopted retrieval procedure.

Figure 4-29: Results for the cross-wired structure with width $w_u = 7.5$ mm – a) Real part of chirality parameter κ (dimensionless); b) Real part of the refractive index for n_+ (RCP), n_- (RCP) and n (dimensionless).

The chirality parameter κ in Figure 4-29 a) shows two resonances, one at 9.2 GHz and another at 10.8 GHz. Above the first resonance frequency, κ is negative between 9.2 GHz and 10.8 GHz, resulting in $n_+ = \sqrt{\epsilon\mu} + \kappa < 0$. Above the second resonance, κ is positive between 10.8 to 14.2 GHz, resulting in $n_- = \sqrt{\epsilon\mu} - \kappa < 0$.

In Figure 4-29 b) it can be noticed that n (green line) is positive through almost the entire frequency range, except at the resonance frequencies where it achieves values of -0.5 for the first resonance and -0.6 for the second resonance. However, n_+ (blue line) is negative for a frequency ranging from 9.2 to 10.2 GHz, and n_- (red line) has a negative region from 10.8 GHz to approximately 15 GHz.

4.2.5 CURVE-WIRED (CHIRAL METAMATERIAL)

The dimensions used in the simulations of the curve-wired chiral metamaterial cell are: $w_u = 7.5$ mm, copper width $w_c = 0.7$ mm, semicircle diameter $d_c = 1.6$ mm, rod height $l_c = 1.06$ mm and offset angle $\varphi = 30^\circ$. The substrate used was alumina ($\epsilon_r = 9.4$ and $\tan\delta = 0.006$) with 0.7 mm of thickness. More details can be seen in Figure 4-30.

Figure 4-30: Unit cell of the curve-wired structure. Each curve-wired inclusion is defined on opposite sides of the substrate with an offset angle between them.

The results for the curve-wired structure obtained for the frequency range of 5 GHz to 9 GHz (range within which the resonant frequency of this chiral metamaterial cell is located) is shown in Figure 4-31. This figure shows the linear transmission coefficients (T_{xx}, T_{yy}, T_{xy} and T_{yx}) for the curve-wired structure.

Figure 4-31: Linear transmission coefficients for the curve-wired structure – a) T_{xx} and T_{yy} coefficients (dB); b) T_{xy} and T_{yx} coefficients (dB).

From the results obtained for the linear transmission it can now be calculated the circular transmission coefficients (T_{++} for RCP, and T_{--} for LCP), both in dB and in phase, as presented in Figure 4-32.

Figure 4-32: Circular transmission coefficients for the curve-wired structure – a) T_{++} (RCP) and T_{--} (LCP) coefficients (dB); b) Phase (degrees) of T_{++} (RCP) and T_{--} (LCP).

Again, the transmission responses for RCP (blue line) and LCP (red line) split into two curves, as can be seen in Figure 4-32 a). It can also be noticed two distinct resonances at frequencies 6 GHz and 8 GHz. For the first resonance (6 GHz), the transmission dip for RCP is more pronounced than that for LCP, indicating a stronger resonance for the RCP case. On the other hand, for the second resonance (8 GHz), it can be observed a stronger resonance for the LCP case.

Figure 4-33 shows the results for the azimuth angle θ and for the ellipticity η .

Figure 4-33: Results for the curve-wired structure – a) Azimuth angle θ (degrees); b) Ellipticity angle η (degrees).

At the two resonance frequencies, the azimuth rotation and ellipticity reach their maximum values ($\theta = -112.6^\circ$, $\eta = -38.8^\circ$ in the first resonance, and $\theta = -84.3^\circ$, $\eta = 20^\circ$ in the second resonance) as can be observed in Figure 4-33 a) and b), respectively. In the region between the two resonance peaks (around 6.8 GHz), we observe a polarization rotation of -11.9° with $\eta \approx 0$. The sign change of η around 6.8 GHz reflects the different frequency dependence between the magnitude of the transmission $|T_{++}|$ and $|T_{--}|$. Again, an incident linear polarized wave with frequency either below or above 6.8 GHz will exhibit different handedness (right or left) when exiting this structure.

Figure 4-34 shows the chirality parameter κ , the refractive index for RCP n_+ and for LCP n_- , and the conventional refraction index n . They were all extracted from simulation results using the adopted retrieval procedure.

Figure 4-34: Results for the curve-wired structure – a) Real part of chirality parameter κ (dimensionless); b) Real part of the refractive index for n_+ (RCP), n_- (RCP) and n (dimensionless).

The chirality parameter κ in Figure 4-34 a) shows two resonances, one at 5.9 GHz and another at 7.8 GHz. Above the first resonance frequency, κ is negative between 5.9 GHz and 7.8 GHz, resulting in $n_+ = \sqrt{\epsilon\mu} + \kappa < 0$. Above the second resonance, κ is positive between 7.8 to 9 GHz, resulting in $n_- = \sqrt{\epsilon\mu} - \kappa < 0$.

In Figure 4-34 b) it can be noticed that n (green line) is positive throughout almost the entire frequency range, except at the resonance frequencies, where it achieves values of -1.3 for the first resonance and -0.7 for the second resonance. However, n_+ (blue line) is negative within the range of 5.9 to 6.2 GHz, and n_- (red line) has a negative region from 7.8 GHz to approximately 9 GHz.

The chirality of these cells can be explained as follows. Each individual inclusion can in fact be seen as an antenna, and both of them are strongly coupled. The offset angle between these inclusions is the responsible for the chirality effect (zero offset means no chirality). Therefore, the combined effect of these inclusions causes the field to be reradiated with an orientation different from that of the original excitation. With an appropriate offset angle, strong chirality can be obtained, with the refractive index presenting negative value for RCP at one resonance and for LCP at another resonance.

4.2.6 2D CHIRAL METAMATERIAL

In a two-dimensional (2D) chiral metamaterial the inclusions are defined on one side of the substrate only, considerably simplifying the fabrication process. The dimensions used in the simulations of the 2D chiral metamaterial cell are: $w_u = 5$ mm, copper width $w_c = 0.27$ mm, and radius $r_c = 1.6$ mm. The substrate used was alumina ($\epsilon_r = 9.4$ and $\tan\delta = 0.006$) with 0.7 mm of thickness. More details can be seen in Figure 4-35.

Figure 4-35: Unit cell of the 2D chiral metamaterial. The structure is defined on only one side of the substrate.

The results for the 2D chiral structure are obtained for frequencies ranging from 10 GHz to 16 GHz (range within which the resonant frequency of this chiral metamaterial cell is located). The linear transmission coefficients (T_{xx} , T_{yy} , T_{xy} and T_{yx}) are shown in Figure 4-36.

Figure 4-36: Linear transmission coefficients for the 2D chiral structure – a) T_{xx} and T_{yy} coefficients (dB); b) T_{xy} and T_{yx} coefficients (dB).

The circular transmission coefficients (T_{++} for RCP, and T_{--} for LCP), both in dB and in phase, are then calculated from the previous results and are shown in Figure 4-37.

Figure 4-37: Circular transmission coefficients for the 2D chiral structure – a) T_{++} (RCP) and T_{--} (LCP) coefficients (dB); b) Phase (degrees) of T_{++} (RCP) and T_{--} (LCP).

The transmission responses for RCP (blue line) and LCP (red line) presents just a single resonance, as can be seen in Figure 4-37 a). This resonance occurs at 13 GHz. The results for RCP and LCP are similar (both are single resonant) because the inclusions are defined on only one side of the substrate. This suggests that the extra resonance occurring in opposite-sided inclusions can be associated to the strong coupling between them.

The results obtained for the azimuth angle θ and for the ellipticity η are shown in Figure 4-38 a-b, respectively. At the resonance of 13 GHz, the azimuth rotation reaches a maximum of around $\theta = -3.66e-5^\circ$ and $\eta = 6.6e-5^\circ$. These values are indeed very small and as a result this structure does not exhibit negative refraction.

Figure 4-38: Results for the 2D chiral structure – a) Azimuth angle θ (degrees); b) Ellipticity angle η (degrees).

Figure 4-39 shows the chirality parameter κ , the refractive index for RCP n_+ and for LCP n_- , and the conventional refraction index n . They were all extracted from simulation results using the retrieval procedure.

Figure 4-39: Results for the 2D chiral structure – a) Real part of the chiral parameter κ (dimensionless); b) Real part of the refractive index for n_+ (RCP), n_- (RCP) and n (dimensionless).

The indices of refraction are positive in all the analyzed frequency range, and it is due to the small value presented by the chirality parameter κ . This implies that the refractive indices for RCP and LCP for $\kappa \approx 0$ is $n_{\pm} = \sqrt{\epsilon\mu} \pm \kappa = \sqrt{\epsilon\mu} > 0$, resulting in permittivity and permeability both positive.

Now, after the analysis of all metamaterial cells, we can introduce the metamaterial cover structure over the monopole antenna.

CHAPTER 5

5 ANALYSIS OF THE RESULTS

In this chapter, the results obtained from the numerical simulations of the complete structure for two different substrate thickness ($w = 1.6$ mm and $w = 0.7$ mm) are presented and discussed, showing the general improvements found in the system with the introduction of the metamaterial cover. It is also presented the experimental results for the substrate thickness $w = 0.7$ mm.

5.1 RESULTS OF THE NUMERICAL ANALYSIS

In Chapter 4, all parameters used to model and analyze the antenna, the metamaterials cells, and the entire structure was described. Now, we proceed to the numerical analysis of the system, which will be carry out with the HFSS software.

As has been discussed before, this software allows one to find several fundamental parameters for the antenna design, such as scattering parameters (S-parameters), gain, directivity, and efficiency.

5.1.1 THE PROPOSED STRUCTURE

With the antenna and the metamaterials cells defined, the next step is to model the complete structure, including the cover introduced around the radiating element (mnonopole antenna).

The complete structure can be seen in Figure 5-1, where the cells are assumed to be the 2D chiral metamaterials just for the sake of illustration. In all cases the substrate is modeled as a polyhedron with 8 sides and 25 mm in height. This has been assumed since the adopted substrates are all rigid materials.

Figure 5-1: Representation of the complete structure containing the monopole antenna surrounded by a 2D chiral metamaterial cover (used here just as an example).

Figure 5-2 presents the top view of the structure placed on the ground plane with the distances between the antenna and the metamaterial cover. The following parameters are adopted: $d = 4/5 \lambda_0$, $d = 3/4 \lambda_0$, $d = 1/2 \lambda_0$, and $d = 1/4 \lambda_0$, and λ_0 refers is the free-space wavelength. The operating frequency of the antenna is 8 GHz.

Figure 5-2: Distances from the monopole antenna to the metamaterial cover with respect to the wavelength (λ_0) for a) $d = 4/5 \lambda_0$, b) $d = 3/4 \lambda_0$, c) $d = 1/2 \lambda_0$, and d) $d = 1/4 \lambda_0$.

We numerically tested various dielectric substrates with a thickness $w = 1.6$ mm, such as FR4 (composite material made of glass fiber with epoxy resin, with a relative permittivity $\epsilon_r = 4.4$ and loss tangent $\tan\delta = 0.02$), widely used in printed circuit boards; Rogers RT/duroid® (composite material made of polytetrafluoroethylene – PTFE – reinforced with glass microfibers with $\epsilon_r = 2.2$ and $\tan\delta = 0.0009$), and aluminum oxide (popularly known as alumina, which is a chemical compound of aluminum and oxygen with $\epsilon_r = 9.4$ and $\tan\delta = 0.006$).

For different types of substrates with thickness $w = 1.6$ mm, the results of the reflection parameter S_{11} , gain, and radiation pattern are shown in Figure 5-3, Figure 5-4, and Figure 5-5, respectively. These results refer to the complete structure (using 2D chiral cells) for the three different types of substrate for the sake of comparison.

Figure 5-3: Reflection parameter S_{11} (dB) for different types of substrates with thickness $w = 1.6$ mm.

Figure 5-4: Gain (dB) for different types of substrates with thickness $w = 1.6$ mm.

Figure 5-5: Radiation pattern for different types of substrates with thickness $w = 1.6$ mm for the resonant frequency of 8 GHz.

The alumina substrate is particularly attractive for the present analysis, because of the appearance of a second useful resonant frequency (below the operating limit of -10 dB, see Figure 5-3). Therefore, this substrate will be adopted in all subsequent simulations. The promising results provided by this geometry (2D chiral metamaterials and 1.6 mm thick alumina substrate, discussed in the next section) allowed us to present this work in [99]. The results clearly demonstrate that the field rotation provided by the metamaterial inclusions around their respective axes opens the possibility of controlling the return loss (S_{11}), gain and input impedance. And it was also possible to obtain a second resonant frequency, which is quite attractive for telecom applications.

5.1.2 ALUMINA SUBSTRATE WITH THICKNESS $w = 1.6$ MM

In this section the metamaterial cover is defined only on a 1.6 mm thick alumina substrate, as this combination produces additional resonances.

First, the metamaterial structure inserted around the antenna is modeled at a distance $d = 4/5 \lambda_0$. The adopted cells are the 2D chiral metamaterial and the total number of cells is 160. The response of the reflection parameter S_{11} (return loss) is shown seen in Figure 5-6.

As can be seen in Figure 5-6, the red line represents the monopole antenna without any inclusion. The blue line is the inclusion of the new structure with the 2D chiral metamaterial in its original form, with the rotation angle $\alpha = 0^\circ$. When these cells are rotated on its own axis, as in $\alpha = 45^\circ$ and $\alpha = 90^\circ$, different responses are obtained, as can be seen green and gray lines.

Figure 5-6: Reflection parameter S_{11} for the monopole antenna surrounded by a 2D chiral metamaterial cover located at a distance $d = 4/5\lambda_0$ from the antenna.

A second resonance caused by the addition of the metamaterial cover also occurs. Interestingly, these results show that depending on the rotation angle of the cells, these resonances can be controlled. In the case where $\alpha = 0^\circ$, it is important to note that there is the possibility of working at two different frequencies, the first at 7.98 GHz and the second at 10.48 GHz. Because these two resonances are below the limit of -10 dB (threshold for an antenna to operate properly), there is the real possibility to operate as a transceiver, that is, an antenna that can transmit data on one frequency and receive on another. This double frequency operation suggests that this antenna could be used for instance in satellite applications, which often needs to send data to Earth at a frequency (*downlink*) and receive them on another frequency (*uplink*).

For the sake of comparison, Figure 5-7 shows the results for the gain of an ordinary monopole antenna gain (red line) together with the results when a 2D chiral metamaterial cover is inserted at a distance of $d = 4/5\lambda_0$ from this monopole antenna. The solid blue line represents the chiral in its original form with $\alpha = 0^\circ$, the dotted blue line with $\alpha = 45^\circ$, and the dashed blue line with $\alpha = 90^\circ$. Again, significant differences are noted between the structures, with a point of maximum improvement, near 7.5 GHz. In this point the gain difference between the ordinary monopole (red line) and the 2D chiral metamaterial cover antenna with $\alpha = 90^\circ$ (dashed blue line) is 6 dB. It is also noted that in almost all frequency range (6 to 12 GHz), the gain presented with the new inserted structure is greater than the gain of the conventional antenna.

Figure 5-7: Gain of a monopole antenna surrounded by a 2D chiral metamaterial cover located at a distance $d = 4/5\lambda_0$ of the antenna.

Table 5.1 lists some important figures-of-merit for a conventional monopole antenna and for a monopole antenna surrounded by a 2D chiral metamaterial cover for different cell rotation angles $\alpha = 0^\circ, 45^\circ$ and 90° .

Table 5.1: Some important figures-of-merit for the conventional monopole antenna and the monopole antenna surrounded by a 2D chiral metamaterial cover for $\alpha = 0^\circ, 45^\circ$ and 90° located at a distance $d = 4/5\lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
$\alpha = 0^\circ$	7.98 / -41.31	3.75	0.88	24.6	10.48 / -12.99	5.90	0.99	39.44
$\alpha = 45^\circ$	8.00 / -5.05	6.89	0.99	7.10	10.43 / -21.55	6.97	0.99	29.55
$\alpha = 90^\circ$	7.99 / -4.99	9.40	0.99	7.07	10.36 / -43.72	8.13	0.99	23.93

As can be seen, for the case where $\alpha = 0^\circ$ there is a significant improvement in the reflection parameter, although the gain is still close to that of a monopole antenna. The efficiency by its turn decreases by 12%, and the input impedance improves, approaching the input impedance of the system. The conventional monopole presents an impedance mismatch of 25% compared to the input impedance of the system while the metamaterial assisted structure with $\alpha = 0^\circ$ shows presents only a 1.6% mismatch. In addition, the improved structure also shows a new resonant frequency with good reflection parameters, gain, and efficiency, despite its 57.76% impedance mismatch.

For the cases in which $\alpha = 45^\circ$ and $\alpha = 90^\circ$, the system cannot operate at the first resonant frequency, since S_{11} is above the threshold of -10 dB. But these angles can still be used for frequencies around 10.4 GHz, because they present considerable gain and efficiency. The impedance mismatch is 18.2% for $\alpha = 45^\circ$ and only 4.28% for $\alpha = 90^\circ$.

The radiation pattern of the monopole antenna and the radiation patterns with the new 2D chiral metamaterials structure in their resonant frequencies [f_{r1} in *a*) and f_{r2} in *b*)] are shown in Figure 5-8. Due to the characteristic of the finite ground plane, the radiation pattern of the monopole antenna is slightly altered when compared with that of a dipole antenna, which generally produces the well known “8” pattern. When a monopole antenna is over a finite ground plane, the outer edges of this ground plane diffract the incident radiation in all directions and, consequently, modify the currents in the ground plane and the vertical element, changing its radiation pattern.

Figure 5-8: Radiation pattern of a monopole antenna and for a monopole antenna surrounded by a 2D chiral metamaterial cover located at a distance $d = 4/5\lambda_0$ - a) First resonant frequency; b) Second resonant frequency.

As seen in Figure 5-8 *a*), there is no significant change in the radiation pattern when the new structure around the antenna is inserted, i.e., it is possible to operate with the new structure keeping the radiation pattern very close to that of a conventional monopole antenna. Small side lobes do appear in the second resonance frequency, which are not desirable. However, these side lobes usually appear mainly due to the geometric characteristics of the structure, the finite ground plane, and also due to the physical medium where the system is installed.

In Figure 5-9, it can be observed the S_{11} parameter (return loss) when the distance between the structure and the 2D chiral metamaterial to $d = 3/4 \lambda_0$ is changed. The total number of cells used is also 160.

Figure 5-9: Reflection parameter S_{11} for a monopole antenna surrounded by a 2D chiral metamaterial cover located at a distance $d = 3/4 \lambda_0$ from the antenna.

Table 5.2 lists some important figures-of-merit for a conventional monopole antenna and for a monopole antenna surrounded by a 2D chiral metamaterial cover.

Table 5.2: Some important figures-of-merit for a conventional monopole antenna and for a monopole antenna surrounded by a 2D chiral metamaterial cover with rotation angles $\alpha = 0^\circ$, 45° and 90° located at a distance $d = 3/4 \lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
$\alpha = 0^\circ$	8.39 / -22.60	4.27	0.78	26.6	11.16 / -9.51	5.01	0.75	50.11
$\alpha = 45^\circ$	8.15 / -8.54	5.18	0.90	11.64	11.09 / -13.96	5.84	0.89	37.44
$\alpha = 90^\circ$	8.20 / -6.48	7.35	0.98	9.03	10.97 / -11.75	4.86	0.70	41.9

Due to the proximity of the metamaterial cover to the antenna, there is a frequency offset with respect to the original antenna. Moreover, for $\alpha = 0^\circ$ it would be possible to operate only in the first resonance frequency ($f_{r1} = 8.39$ GHz). For $\alpha = 45^\circ$ and $\alpha = 90^\circ$, it would only be possible to operate in the second resonance frequency (approximately $f_{r2} = 11$ GHz). Besides the good reflection parameter S_{11} and gain, for $\alpha = 0^\circ$ there is also a low impedance mismatch (6.4%), although a decrease in radiation efficiency of 22% is observed. For $\alpha = 45^\circ$ and $\alpha = 90^\circ$, both operating in f_{r2} , good results are obtained for the gain at the expense of a lower efficiency and higher impedance mismatch, which is detrimental for practical use.

The radiation patterns for the conventional monopole and the 2D chiral metamaterials assisted monopole antenna is shown in Figure 5-10. The appearance of some side lobes, as

already explained, are common in a real system operated with a vertical element on a finite ground plane.

Figure 5-10: Radiation pattern of a conventional monopole antenna and a monopole antenna surrounded by a 2D chiral metamaterial cover. The cover is at a distance $d = 3/4\lambda_0$ - a) First resonant frequency; b) Second resonant frequency.

Modifying again the distance of the chiral metamaterial structure with respect to the antenna, i.e, assuming $d = 1/2 \lambda_0$, implies that the total number of cells added to the structure coupled to the antenna is reduced to 80. The results are shown in Figure 5-11.

Figure 5-11: Reflection parameter S_{11} for a monopole antenna surrounded by a 2D chiral metamaterial cover at a distance $d = 1/2 \lambda_0$ from the antenna.

The shorter distance between cover and antenna causes a frequency offset relative to the conventional monopole. But, in this situation, for the cell rotation angles $\alpha = 0^\circ$, $\alpha = 45^\circ$ and $\alpha = 90^\circ$, it is possible to work at two resonant frequencies (approximately $f_{r1} = 6.8$ GHz and $f_{r2} = 9.2$ GHz), since these resonances are below the threshold of -10 dB. It is interesting to note that the proposed system is capable of operating as a transceiver in all situations with the rotation angles mentioned above.

Table 5.3 lists some important figures-of-merit for each 2D chiral metamaterial surrounded monopole antenna and for a conventional monopole antenna for the sake of comparison.

Table 5.3: Important figures-of-merit of a conventional monopole antenna and of a monopole antenna covered with a 2D chiral metamaterial. The cell angles are $\alpha = 0^\circ, 45^\circ$ and 90° . The cover is located at a distance $d = 1/2 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
$\alpha = 0^\circ$	6.84 / -13.27	3.61	0.99	34.62	9.25 / -20.71	5.47	0.98	29.86
$\alpha = 45^\circ$	6.78 / -15.06	3.99	0.99	32.74	9.24 / -28.95	5.44	0.98	26.77
$\alpha = 90^\circ$	6.73 / -25.49	4.33	0.99	27.5	9.17 / -39.6	5.57	0.98	24.58

As can be seen, the radiation efficiency for the first resonance frequency f_{r1} for the three rotation angles, namely $\alpha = 0^\circ$, $\alpha = 45^\circ$ and $\alpha = 90^\circ$, is approximately $\eta_{fr1} = 0.99$, whilst for the second resonance frequency f_{r2} the efficiency is $\eta_{fr2} = 0.98$. The efficiency is therefore improved for both cases, almost converging to the efficiency of the conventional antenna pattern.

For $\alpha = 90^\circ$, excellent values were achieved for the reflection parameter S_{11} , gain, efficiency and also for the impedance mismatch (10% mismatch for f_{r1} and 1.68% for f_{r2}).

The radiation patterns for these cases are shown in Figure 5-12, where it is possible to observe that for f_{r1} the behavior is similar to that of a conventional monopole antenna (due to the small side lobes). For f_{r2} , the radiation pattern maintains similar characteristics to other cases previously described, with small side lobes, an intrinsic characteristic of radiator elements over a finite ground plane.

Next, the previous analysis is repeated but this time for a cover consisting of conventional metamaterials (SRR and Omega structures). Let's first assume the cover is located at a distance $d = 3/4 \lambda_0$ to the antenna. It is possible to notice some differences in the responses. The total number of cells used in this case is 160 for the single SRR (5 mm) and 160 for the Omega structure.

Figure 5-12: Radiation pattern of monopole antenna surrounded by a 2D chiral metamaterial cover located at a distance $d = 1/2 \lambda_0$ - a) First resonant frequency; b) Second resonant frequency.

It can be clearly noticed in the reflection parameter diagram, shown in Figure 5-13, that SRR-based covers (solid black line) can operate on both resonant frequencies, which are approximately $f_{r1} = 8.1$ GHz and $f_{r2} = 11$ GHz. However, Omega-based covers (pink line) show a broadening of the band around f_{r1} of approximately 1 GHz. Nevertheless, it would not be possible to operate at f_{r2} , since its return loss is above the -10 dB limit.

Figure 5-13: Reflection parameter S_{11} for a monopole antenna with a cover consisting of conventional metamaterial cells located at a distance $d = 3/4 \lambda_0$ from the antenna.

Table 5.4 lists some of the important figures-of-merit for a conventional monopole and a monopole surrounded by a cover of conventional metamaterial cells.

The Omega cell produces some improvement in the overall antenna performance, such as the gain. The efficiency is close to the ideal, and there is also a low impedance mismatch (only 1.8%). For the single SRR, the gain is improved by approximately 2.3 dB and there is also a low impedance mismatch (2.64%).

Table 5.4: Important figures-of-merit for a conventional monopole antenna and a monopole antenna surrounded by a conventional metamaterial cover located at a distance $d = 3/4 \lambda_0$.

Tipo	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Omega	8.03 / -15.54	4.35	0.96	24.55	11.06 / -9.06	6.55	0.99	57.17
SRR (5 mm)	8.13 / -14.46	6.11	0.95	25.66	10.98 / -32.61	2.36	0.47	26.18

It is interesting to note in Figure 5-14 that with the Omega cell, the radiation pattern is similar to that of a conventional monopole antenna. On the other hand, SRR cells produce undesirable upper lobes that appear due to the intrinsic geometrical characteristics of this metamaterial that affects the overall system electromagnetic response. Figure 5-14 b) shows the radiation patterns obtained only for the SRR-based cover at the frequency f_{r2} , where is also possible to observe small side lobes and upper lobes.

Figure 5-14: Radiation pattern of a monopole antenna surrounded with conventional metamaterials cover located at a distance $d = 3/4 \lambda_0$ - a) First resonant frequency (SRR and omega cells); b) Second resonant frequency (SRR only).

Continuing with the analysis of conventional metamaterials, but now with a cover located at a distance $d = 1/2 \lambda_0$ of the structure to the antenna, it is possible to observe even more differences in the antenna responses. The total number of cells used in this case is 80 for both the smaller SRR (5 mm) and the Omega structures.

In the reflection parameter shown in Figure 5-15, it is found that for the SRRs (solid black line and dashed black line) a new resonance frequency is produced. Now there are three frequencies for the SRR structure, due to the proximity of structures to the near field of the antenna, making the electromagnetic field interactions more intense.

Figure 5-15: Reflection parameter S_{11} for a monopole antenna surrounded by conventional metamaterial cells located at a distance $d = 1/2 \lambda_0$ from the antenna.

For the single SRR (solid black line) there is a broadening of the frequency band of approximately 1.1 GHz, which now ranges from 9.1 GHz to 10.2 GHz. The Omega structure (pink line), by its turn, can only operate at a resonance frequency $f_{r1} = 9.3$ GHz with a 0.8 GHz bandwidth.

Table 5.5 lists some important figures-of-merit for the conventional monopole antenna and for the monopole antenna surrounded either SRR- or Omega-based covers.

Table 5.5: Important figures-of-merit for the conventional monopole antenna and for the monopole antenna surrounded by a cover metamaterials (either SRR or Omega cells). The cover is located at a distance $d = 1/2 \lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Omega	9.3 / -17.08	5.41	0.99	33.07	---	---	---	---
SRR (5 mm)	6.48 / -26.86	4.47	0.95	27.15	8.42 / -17.52	2.59	0.87	31.14
Type	f_{r3} (GHz) / S_{11} (dB)	G_{fr3} (dB)	η_{fr3}	Z_{infr3} (Ω)				
SRR (5 mm)	9.20 / -26.67	5.42	0.94	23.35				

For the SRR cell, the efficiency values indicate operation at three distinct resonance frequencies. The gain at f_{r1} increases when compared to that of a conventional antenna. Moreover, the gains observed at f_{r2} and f_{r3} also indicate a useful operating point (with return loss < -10 dB). The impedance mismatch at f_{r1} and f_{r3} are 8.6% and 6.6%, respectively). The efficiency obtained with the Omega structure is almost equal to that obtained with the conventional monopole antenna, but its gain is about 2 dB higher, at the expense of a higher impedance mismatch which is not adequate for practical use.

The radiation pattern of each case is shown in Figure 5-16. Interestingly, in the third resonance frequency (Figure 5-16 c), there is an increase in the directivity, which can be observed by the narrower main lobe.

Figure 5-16: Radiation pattern of a monopole antenna surrounded by a conventional metamaterial cover located at a distance $d = 1/2 \lambda_0$ - a) First resonant frequency (SRR and Omega cells); b) Second resonant frequency (SRR only); c) Third resonant frequency (SRR only).

Finally, the responses of chiral metamaterials (cross-wired and curve-wired) covers located at a distance $d = 1/2 \lambda_0$ from the structure to the antenna, are presented. In the reflection parameter curves shown in Figure 5-17, it is possible to observe that for the curve-wired cells (dashed green line) there are two operable resonant frequencies ($f_{r1} = 6.68$ GHz and $f_{r2} = 9.14$ GHz).

Figure 5-17: Reflection parameter S_{11} for a monopole antenna surrounded by a cover of chiral metamaterials cells located at a distance $d = \lambda_0/2$ from the antenna.

Table 5.6 lists some important figures-of-merit for the conventional monopole antenna and for a monopole antenna surrounded by a cover of chiral metamaterial cells (cross- and curve-wired).

Table 5.6: Some important figures-of-merit for the conventional monopole antenna and for a monopole antenna surrounded by a cover of chiral metamaterial cells (cross- and curve-wired) located at a distance $d = 1/2 \lambda_0$.

Tipo	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Cross-wired	6.68 / -15.75	4.47	0.99	31.78	9.14 / -34.19	5.50	0.99	25.87
Curve-wired	9.27 / -19.32	5.45	0.99	30.93	---	---	---	---

The radiation patterns for these chiral metamaterials at f_{r1} are shown in Figure 5-18. The radiation pattern of the crossed wires (3.75 mm) in f_{r2} is shown in Figure 5-18 b). As can be observed, the radiation patterns are similar to those presented previously, with the appearance of small lateral lobes.

Figure 5-18: Radiation pattern of a monopole antenna surrounded by a chiral metamaterial cover located at a distance $d = 1/2 \lambda_0$ - a) First resonant frequency (cross- and curve-wired); b) Second resonant frequency (curve-wired only).

5.1.3 ALUMINA SUBSTRATE WITH THICKNESS $w = 0.7$ MM

The numerical results shown in this section are obtained for an alumina substrate with thickness $w = 0.7$ mm. It is also shown the experimental characterization when a metamaterial cover is located at a distance $d = 4/5 \lambda_0$ from the antenna.

First, we show the simulated and experimental measurement of a conventional monopole antenna on a 60 mm wide ground plane. The results shown in Figure 5-19 indicate the resonance in the simulated result, as already discussed in Section 4.1.1, occurs at 8.23 GHz, with a minimum S_{11} of -18.5 dB. For the measured result, the resonance occurs at 8.15 GHz with a minimum of -12.9 dB.

Figure 5-19: Reflection parameter S_{11} of the conventional monopole antenna over a finite ground plane with width $w_{gp} = 60$ mm.

Next, the modeled metamaterial cover is fabricated and inserted at a distance $d = 4/5 \lambda_0$ around the antenna. For all 2D chiral metamaterial covers, the total number of cells on the dielectric structure is 200 for $w_u = 5$ mm. The frequency response of the reflection parameter S_{11} can be seen in Figure 5-20.

The solid blue line represents the simulated reflection parameter and the red dotted line represents the reflection measurement of the monopole antenna with the metamaterial cover composed of 2D chiral cells with three different rotation angles ($\alpha = 0^\circ$, $\alpha = 45^\circ$ and $\alpha = 90^\circ$). For all cases, there is no agreement between simulated and measured results, where can also be seen a frequency shift of about 2 GHz. The resonance in the simulation occurs at around 8 GHz, but in the measurement it occurs at 9.7 GHz.

Figure 5-20: Reflection parameter S_{11} for a monopole antenna surrounded by a 2D chiral metamaterial cover with a) $\alpha = 0^\circ$, b) $\alpha = 45^\circ$, and c) $\alpha = 90^\circ$ located at a distance $d = 4/5 \lambda_0$ from the antenna.

The divergences may have occurred due to the difference in the substrate permittivity. In the simulations, an alumina substrate with relative permittivity $\epsilon_r = 9.6$ and loss tangent $\tan\delta = 0.006$ was used, while for the experimental tests these parameters could not be actually tested.

Although the resonances are shifted with respect to the simulated ones, their behavior are similar, as seen in Figure 5-20 c) (for $\alpha = 90^\circ$). This may indeed suggest the substrate parameters adopted in the simulations do not quite agree with those of the substrate utilized in the experiments.

Table 5.7 lists some important figures-of-merit numerically obtained for the conventional monopole antenna and for the monopole antenna surrounded by a 2D chiral metamaterial cover for different cell rotation angles $\alpha = 0^\circ, 45^\circ$ and 90° .

Table 5.7: Some important figures-of-merit for the conventional monopole antenna and the monopole antenna surrounded by a 2D chiral metamaterial cover for $\alpha = 0^\circ, 45^\circ$ and 90° located at a distance $d = 4/5 \lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
$\alpha = 0^\circ$	8.05 / -12.74	4.75	1	15.92	---	---	---	---
$\alpha = 45^\circ$	8.45 / -12.45	4.03	0.89	32.43	10.39 / -11.40	7.05	1	43.31
$\alpha = 90^\circ$	8.17 / -33.28	5.08	0.88	24.85	10.31 / -16.81	6.06	1	33.44

For $\alpha = 0^\circ$ there is no significant improvement in the overall system performance. For $\alpha = 45^\circ$ a second resonant frequency appears, with a gain of 7.05 dB. The best case here is for $\alpha = 90^\circ$, where there is a significant improvement in the reflection parameter, with a gain increase of 1.29 dB compared to the conventional monopole, and a better impedance match to the system. However, the efficiency decreases by 12%. For the same structure, it is also obtained a second resonant frequency (10.30 GHz) with good reflection parameters, 6.06 dB gain, and good efficiency. The simulated radiation patterns for the 2D chiral metamaterials cover are presented in Figure 5-21.

Table 5.8 lists the measured reflection parameters for the conventional monopole antenna and for the monopole antenna with 2D chiral metamaterial cover with $\alpha = 0^\circ, 45^\circ$ and 90° coupled to it.

Figure 5-21: Simulated radiation patterns for the monopole antenna with 2D chiral metamaterial cover at a distance $d = 4/5 \lambda_0$ with $\varphi = 90^\circ$ for both resonant frequencies f_{r1} and f_{r2} , a) $\alpha = 0^\circ$, b) $\alpha = 45^\circ$ c) $\alpha = 90^\circ$.

Table 5.8: Measured reflection parameter for the conventional monopole antenna and for the monopole antenna with the 2D chiral metamaterial cover with $\alpha = 0^\circ$, 45° and 90° coupled to it. The cover is located at a distance $d = 4/5 \lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)
Monopole	8.15 / -12.90
$\alpha = 0^\circ$	9.70 / -11.67
$\alpha = 45^\circ$	10.01 / -17.78
$\alpha = 90^\circ$	10.01 / -32.88

As shown in Figure 5-21, the radiation patterns for the first resonance f_{r1} (for all cell rotation angles) are very similar to that of a conventional monopole. Therefore, the metamaterial cover can be installed without any significant penalty to the system.

For the cross-wired cell with $w_u = 3.75$ mm the total number of cells inserted on the dielectric structure is 288, and for $w_u = 7.5$ mm, it is 72. The simulated and measured reflection parameter responses for these metamaterials can be seen in Figure 5-22. There is a good agreement between the simulated and measured results for these structures, but it can also be observed a frequency shift of about 250 MHz for $w_u = 3.75$ mm and 400 MHz for $w_u = 7.5$ mm.

Figure 5-22: Reflection parameter S_{11} for the cross-wired cell cover with a) $w_u = 7.5$ mm and b) $w_u = 3.75$ mm at a distance $d = 4/5 \lambda_0$ from the antenna.

For the curve-wired cell, the total number of cells inserted on the alumina structure is 72 for $w_u = 7.5$ mm. The simulated and measured reflection parameter responses are shown in Figure 5-23. Again, a frequency shift between the resonant frequencies of the simulated and measured results is observed, but their resonant frequency patterns are much closer in this case.

Figure 5-23: Reflection parameter S_{11} for the curve-wired cell cover at a distance $d = 4/5\lambda_0$ from the antenna.

Table 5.9 lists some important figures-of-merit simulated for the conventional monopole antenna and for the cross- and curve-wired metamaterial covers coupled to the antenna.

Table 5.9: Figures-of-merit for the conventional monopole antenna and for the cross- and curve-wired metamaterials covers located at a distance $d = 4/5\lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25
Cross-wired $w_u = 3.75$ mm	10.25 / -15.34	5.94	1	35.32
Cross-wired $w_u = 7.5$ mm	10.01 / -20.51	5.86	0.72	30.13
Curve-wired	9.36 / -13.42	5.53	0.89	38.27

The resonant frequencies of all three cases are blue-shifted (towards higher frequencies). It is interesting to note the increase in the gain for all inclusions. The efficiency is maintained for cross-wired cells with $w_u = 3.75$ mm, but decreases for crossed wires with $w_u = 7.5$ mm and for curve-wired cells with $w_u = 7.5$ mm, and this may be due to a low interaction of the electromagnetic fields, once the dimension of the cells is larger compared to the wavelength. Table 5.10 lists the measured reflection parameters for the conventional monopole antenna and for the cross- and curve-wired cell metamaterial covers coupled to the antenna.

Table 5.10: Measured reflection parameters for the conventional monopole antenna and for the cross- and curve-wired metamaterial covers, both located at a distance $d = 4/5 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	f_{r2} (GHz) / S_{11} (dB)
Monopole	8.15 / -12.90	---
Cross-wired $w_u = 3.75$ mm	10.01 / -15.33	---
Cross-wired $w_u = 7.5$ mm	9.6 / -16.40	---
Curve-wired	8.30 / -12.34	9.40 / -17.96

The agreement between measured and simulated results is very good. The resonance frequency for the cross-wired cell cover with $w_u = 3.75$ mm is red-shifted (towards lower frequencies) by 240 MHz. On the other hand, the resonance frequency for the cross-wired cell cover with $w_u = 7.5$ mm is blue-shifted by 410 MHz. The curve-wired cell cover by its turn shows an additional resonance frequency below the -10 dB limit at 8.3 GHz, as shown in Figure 5-23 (observe the two frequency dips). The simulated radiation patterns for the cross- and curve-wired chiral metamaterial covers are presented in Figure 5-24 for the first resonance f_{r1} .

Figure 5-24: Simulated radiation patterns for the monopole antennas with chiral metamaterials covers at a distance $d = 4/5 \lambda_0$ with $\varphi = 90^\circ$ in the resonant frequency f_{r1} . a) Cross-wired with $w_u = 3.75$ mm b) Cross-wired with $w_u = 7.5$ mm, and c) Curve-wired.

Next, for the omega cell metamaterial cover the total number of cells inserted on the dielectric structure is 160 for $w_{u1} = 4.57$ mm and $w_{u2} = 5.3$ mm. The simulated and measured reflection parameter responses are presented in Figure 5-25. As can be seen, the resonance patterns are somewhat similar, but they have a considerable difference in the S_{11} frequency dip values. It is interesting to note that the measured results present a much larger frequency band than the simulated one, going from 7.40 GHz to 9.50 GHz, i.e., a bandwidth of 2.1 GHz. This may be an allusion to the unidentified alumina properties used in the experiment and the

high interaction between the electromagnetic fields and this structure. This antenna may find an application in broadband situations, since broadband antennas are capable of receiving a wide range of frequencies, while narrowband antenna receives a single-frequency.

Figure 5-25: Reflection parameter S_{11} for omega cell metamaterial cover at a distance $d = 4/5 \lambda_0$ from the antenna.

The single and double SRR cell covers, by its turn, present a total number of 200 cells with $w_u = 5$ mm. The simulated and measured reflection parameter responses for these two cases are presented in Figure 5-26. As can be seen, the single and double SRR metamaterial covers present a good agreement between the simulated and measured results, and both have a very similar frequency behavior, once they have the same size and geometry, with exception of the resonator inclusion on back side of the double SRR. The single SRR cell cover (Figure 5-26 *a*), shows a red-shift of its main resonance frequency (the first resonance is above the -10 dB limit and is discarded here). The first resonance frequency for the double SRR cell cover (Figure 5-26 *b*), by its turn, is blue-shifted while the second is red-shifted, both with respect to the simulated results.

Unfortunately, the conventional SRR metamaterial cover could not be fabricated due to its small dimensions (the width of the unit cell is 2.5 mm), which is not compatible with the resolution of the available fabrication process. The small cell dimension demands a much larger number of cells to be used in the metamaterial covers, namely, 720 cells. The simulated reflection parameter responses for this metamaterial is presented in Figure 5-27.

Figure 5-26: Reflection parameter S_{11} for a) single SRR and b) double SRR metamaterial cover at a distance $d = 4/5 \lambda_0$ from the antenna.

Figure 5-27: Reflection parameter S_{11} for conventional SRR metamaterial cover at a distance $d = 4/5 \lambda_0$ from the antenna.

Table 5.11 lists some important simulated figures-of-merit for the conventional monopole antenna and for the monopole antenna with three different metamaterial covers based, respectively, on omega, single and double SRR cells.

Table 5.11: Important figures-of-merit for the conventional monopole antenna and for the monopole antenna with three different metamaterial covers based, respectively, on omega, single and double SRR cells. The cover is located at a distance $d = 4/5 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Omega	7.55 / -16.23	5.70	0.96	21.53	8.64 / -11.89	4.25	0.92	42.05
Conventional SRR	7.27 / -11.25	4.90	0.98	16.96	8.43 / -17-81	5.98	0.97	19.32
Single SRR	10.08 / -18.65	5.47	0.98	19.81	---	---	---	---
Double SRR	10.10 / -23.23	3.95	0.68	21.8	---	---	---	---
Type	f_{r3} (GHz) / S_{11} (dB)	G_{fr3} (dB)	η_{fr3}	Z_{infr3} (Ω)				
Conventional SRR	10.56 / -13.55	8.47	1	38.08				

As can be seen, a second resonant frequency appears for the omega cell-based cover, while the conventional SRR presents three different resonances, which is really attractive for telecom applications. The gain increases with all metamaterial inclusions. The efficiency reaches almost 100% for all cases, except for the double SRR (drops to 0.68).

Table 5.12 lists the measured reflection parameters for the conventional monopole antenna and for the monopole antenna surrounded by a conventional metamaterial cover.

Table 5.12: Measured reflection parameters for the conventional monopole antenna and for the antenna with conventional metamaterial cover at a distance $d = 4/5 \lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)	f_{r2} (GHz) / S_{11} (dB)	f_{r3} (GHz) / S_{11} (dB)
Monopole	8.15 / -12.90	---	---
Omega	7.70 / -14.55	8.70 / -22.01	9.20 / -23.76
Single SRR	9.80 / -27.14	---	---
Double SRR	9.00 / -11.02	9.70 / -20.83	---

Regarding the measured results, for the omega cell it is possible to see a blue-shift of 150 MHz of the first resonance and of 60 MHz of the second resonance. Moreover, a third resonant frequency appears at 9.20 GHz for the omega cell, as shown in Figure 5-25. A 280 MHz red-shift is observed for the resonance frequency of the single SRR. The double SRR, by its turn, exhibits a 380 MHz red-shift in its resonance frequency (f_{r2}), along with the appearance of a new resonance at 9 GHz (f_{r1}), which agrees with the simulated results shown in Figure 5-26 b). The radiation patterns for the three conventional metamaterial covers are presented in Figure 5-28

Figure 5-28: Simulated radiation patterns for the monopole antenna with conventional metamaterial cover at a distance $d = 4/5 \lambda_0$ in all resonant frequencies f_{r1} , f_{r2} and f_{r3} - a) Omega, b) Conventional SRR, c) Single SRR, and d) Double SRR.

The next results refer to chiral metamaterial covers, but due to the lack of smaller alumina substrates only theoretical results are shown. The distance from the antenna to the 2D chiral metamaterial cover is now assumed as $d = 3/4 \lambda_0$, and the number of cells is 160 for all cases ($\alpha = 0^\circ$, $\alpha = 45^\circ$ and $\alpha = 90^\circ$). The results for the reflection parameter S_{11} of all three cells are shown in Figure 5-29.

Figure 5-29: Reflection parameter S_{11} for 2D chiral metamaterial cover at a distance $d = 3/4 \lambda_0$ from the antenna.

As can be seen, the S_{11} is basically the same for all cases, so these structures do not exhibit the desirable effects at this distance, such as the appearance of a second resonance, as should be expected. Table 5.13 lists some important figures-of-merit for a conventional monopole antenna and for a monopole antenna surrounded by a 2D chiral metamaterial cover.

Table 5.13: Figures-of-merit numerically obtained for a conventional monopole antenna and for a monopole antenna surrounded by a 2D chiral metamaterial at a distance $d = 3/4 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	$G_{f_{r1}}$ (dB)	$\eta_{f_{r1}}$	Z_{infr1} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25
$\alpha = 0^\circ$	8.17 / -15.20	4.90	1	17.82
$\alpha = 45^\circ$	8.16 / -14.98	5.37	1	17.62
$\alpha = 90^\circ$	8.16 / -14.81	5.22	1	17.54

As can be observed, the results are basically the same for all of these structures, but with an improved gain, maintaining the same efficiency and closely the same input impedance. The radiation patterns for these structures at the resonant frequency f_{r1} are shown in Figure 5-30.

Figure 5-30: Simulated radiation patterns for the monopole antenna for the monopole antenna surrounded by a 2D chiral metamaterial cover at a distance $d = 3/4 \lambda_0$ at the resonant frequency f_{r1} . a) $\alpha = 0^\circ$, b) $\alpha = 45^\circ$, and c) $\alpha = 90^\circ$.

It is interesting to note that the radiation patterns for the three 2D chiral metamaterials cases are very similar, and they are comparable to the monopole itself. These cells did not present any different effects, and the changes in the gain and the impedance can have been influenced by the substrate properties.

Next, cross- and curve-wired metamaterial cells are considered, but still keeping the distance $d = 3/4 \lambda_0$ from the antenna. This time the number of cross-wired cells is increased to 288 for the smaller cells and 72 for the larger ones. The number of curve-wired cells is also 72. The simulation results for the reflection parameter S_{11} for these three cell types are shown in Figure 5-31. The resonance frequencies for each of these cases, including the conventional monopole antenna, are shown in Table 5.14.

Figure 5-31: Reflection parameter S_{11} for a monopole antenna surrounded by a chiral metamaterial cover at a distance $d = 3/4 \lambda_0$ from the antenna.

Table 5.14: Simulated results for monopole antenna surrounded by a chiral metamaterial cover at a distance $d = 3/4 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25
Cross-wired $w_u = 3.75$ mm	10.55 / -13.21	6.26	1	38.98
Cross-wired $w_u = 7.5$ mm	10.44 / -22.75	2.25	0.49	28.86

As can be observed, the results are basically the same for the cross-wired structures, but the metamaterial with $w_u = 3.75$ mm is particularly interesting, because it has an improved gain (6.26 dB), maintaining the same efficiency, but with an increased mismatch of the input impedance. The curve-wired cell does not present resonances, so it is not of practical use. The radiation pattern for this case is shown in Figure 5-32.

Figure 5-32: Simulated radiation pattern for a monopole antenna surrounded by a cross-wired metamaterial cover at a distance $d = 3/4 \lambda_0$ at the resonant frequency f_{r1} . a) $w = 3.75$ mm and b) $w = 7.5$ mm.

Next, the analysis is carried out for conventional metamaterials maintaining the cover to antenna distance as $d = 3/4 \lambda_0$. The number of cells is 160 for omega, 720 for conventional SRR, and 160 for both single and double SRR. The simulated reflection parameters S_{11} for all these cases are shown in Figure 5-33.

Figure 5-33: Reflection parameters S_{11} for a monopole antenna surrounded by conventional metamaterial cover at a distance $d = 3/4 \lambda_0$ from the antenna.

Table 5.15 lists some important figures-of-merit numerically obtained for the monopole antenna and the monopole antenna surrounded by a conventional metamaterial cover.

Table 5.15: Figures-of-merit numerically obtained for a conventional monopole antenna and for a monopole antenna surrounded by conventional metamaterial cover a distance $d = 3/4 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Omega	7.68 / -13.76	5.39	0.96	18.95	---	---	---	---
Conventional SRR	7.37 / -17.16	4.23	0.91	21.00	8.51 / -22.92	5.97	0.97	21.68
Single SRR	10.38 / -21.63	5.45	1	21.19	---	---	---	---
Double SRR	10.41 / -23.24	3.84	0.69	21.79	---	---	---	---

The conventional SRR cell presents two resonant frequencies with acceptable gain, efficiency, and input impedance. The other structures, on the other hand, present improved gain maintaining practically the same efficiency (except for the double SRR, which had a decrease in the efficiency). The radiation pattern for the each of these cases is shown in Figure 5-34.

Figure 5-34: Simulated radiation patterns for the monopole antenna surrounded by conventional metamaterial cover at a distance $d = 3/4 \lambda_0$ from the antenna at both resonant frequencies f_{r1} and f_{r2} . a) Omega, b) Conventional SRR, c) Single SRR, and d) Double SRR.

Now, the cover to antenna distance is changed to $d = 1/2 \lambda_0$. The analysis is carried out first for three 2D chiral metamaterial cover configurations. The total number of cells is 120 for all three cases ($\alpha = 0^\circ$, $\alpha = 45^\circ$ and $\alpha = 90^\circ$). It is possible to observe in the reflection parameter S_{11} shown in Figure 5-35 the presence of only a single resonant frequency around 9 GHz, with a bandwidth of approximately 800 MHz. Table 5.16 lists some of the figures-of-merit for the analyzed structures.

Figure 5-35: Reflection parameter S_{11} for the monopole antenna surrounded by a 2D chiral metamaterial cover at a distance $d = 1/2 \lambda_0$ from the antenna.

Table 5.16: Figures-of-merit for the conventional monopole antenna and the monopole antenna surrounded by a 2D chiral metamaterial cover at a distance $d = 1/2 \lambda_0$.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25
$\alpha = 0^\circ$	9.11 / -14.70	5.40	1	36
$\alpha = 45^\circ$	9.14 / -22.04	5.29	1	29.18
$\alpha = 90^\circ$	9.09 / -36.97	5.63	1	25.72

In all cases, the responses are similar, but the best case occurs for $\alpha = 90^\circ$, for which the smallest reflection of -36.97 occurs. This case also shows the greatest gain (5.63 dB), almost 2 dB higher than that of a conventional monopole antenna. Yet, it maintains the same efficiency with a better impedance matching to the system. The radiation patterns for these cases are shown in Figure 5-36, but only for the resonant frequency f_{r1} .

Figure 5-36: Simulated radiation patterns for the monopole antenna surrounded by 2D chiral metamaterial cover at a distance $d = 1/2 \lambda_0$ for the resonant frequency f_{r1} . a) $\alpha = 0^\circ$, b) $\alpha = 45^\circ$ c) $\alpha = 90^\circ$.

Next, cross- and curve-wired metamaterials are considered, still for distance $d = 1/2 \lambda_0$ from the antenna. The total number of cells is 192 for the cross-wired width $w_u = 3.75$ mm and 48 the cross-wired with $w_u = 7.5$ mm, while for the curve-wired the number of cells is 48. The reflection parameters S_{11} for these cases are shown in Figure 5-37. Table 5.17 lists some of the important figures-of-merit for these cases.

Figure 5-37: Reflection parameter S_{11} for the monopole antenna surrounded by chiral metamaterial cover at a distance $d = 1/2 \lambda_0$ from the antenna.

Table 5.17: Figures-of-merit for the conventional monopole antenna and for the monopole antenna surrounded by a chiral metamaterial cover at a distance $d = 1/2 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Cross-wired $w_u = 3.75$ mm	9.03 / -70.61	5.67	1	25	---	---	---	---
Cross-wired $w_u = 7.5$ mm	8.85 / -17.55	5.89	0.98	20.15	9.48 / -13.95	4.60	0.62	33.09
Curve-wired	6.41 / -13.11	5.02	0.99	36.28	8.52 / -39.19	5.71	0.95	24.02

Type	f_{r3} (GHz) / S_{11} (dB)	G_{fr3} (dB)	η_{fr3}	Z_{infr3} (Ω)
Curve-wired	8.82 / -39.19	7.02	0.77	24.47

All cases show similar response, but the curve-wired in particular shows two operating frequencies, the first one at 6.41 GHz and the second one covering a bandwidth from 8.38 to 8.96 GHz. The gain increases in all cases, maintaining approximately the same efficiency and, in most cases, a better impedance matching to the system. The radiation patterns for these cases are shown in Figure 5-38.

Figure 5-38: Simulated radiation patterns for a monopole antenna surrounded with chiral metamaterials cover at a distance $d = 1/2 \lambda_0$. The resonant frequencies are f_{r1} , f_{r2} and f_{r3} . a) cross-wired width $w_u = 3.75$ mm, b) cross-wired with $w_u = 7.5$ mm, and c) curve-wired.

Now, the analysis is carried out for conventional metamaterial covers still located at a distance $d = 1/2 \lambda_0$ from the antenna. The total number of cells is 120 for omega, 480 for conventional SRR, and 120 for both single and double SRR. The reflection parameters of these four cases are presented in Figure 5-39.

Figure 5-39: Reflection parameter S_{11} for a monopole antenna surrounded by a conventional metamaterial cover at a distance $d = 1/2 \lambda_0$ from the antenna.

Table 5.18 lists some important figures-of-merit numerically obtained for these four structures along with those for the conventional monopole antenna. The corresponding radiation patterns are shown in Figure 5-40.

Table 5.18: Figures-of-merit for the conventional monopole antenna and for the monopole antenna surrounded by a conventional metamaterial cover at a distance $d = 1/2 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Omega	6.22 / -26.20	3.27	0.78	26.08	8.16 / -10.65	7.07	0.94	24.08
Conventional SRR	8.12 / -13.37	4.88	0.91	17.20	9.53 / -43.87	4.82	0.94	24.89
Single SRR	6.43 / -16.49	5.37	1	31.63	8.81 / -34.03	3.27	0.83	25.39
Double SRR	8.53 / -23.33	3.67	0.84	27.01	9.00 / -19.40	3.62	0.85	24.02

Type	f_{r3} (GHz) / S_{11} (dB)	G_{fr3} (dB)	η_{fr3}	Z_{infr3} (Ω)
Omega	8.74 / -10.51	7.96	0.95	13.65
Single SRR	8.98 / -24.16	5.78	1	22.30
Double SRR	9.18 / -17.04	5.12	0.94	31.32

Figure 5-40: Simulated radiation patterns for the monopole antenna surrounded with a conventional metamaterials cover at a distance $d = 1/2 \lambda_0$. The resonant frequencies are f_{r1} , f_{r2} and f_{r3} . a) Omega, b) Conventional SRR, c) Single SRR, and d) Double SRR.

The cover to antenna distance is now reduced to $d = 1/4 \lambda_0$, and the analysis is carried out for 2D chiral metamaterials. The number of cells is 40 for all cases ($\alpha = 0^\circ$, $\alpha = 45^\circ$ and $\alpha = 90^\circ$). The reflection parameters S_{11} corresponding to these cases are shown in Figure 5-41. Since all curves are above the -10 dB limit, these structures are of no practical use.

Figure 5-41: Reflection parameter S_{11} for a monopole antenna surrounded by a 2D chiral metamaterial cover at a distance $d = 1/4 \lambda_0$ from the antenna.

Still at a distance of $d = 1/4 \lambda_0$, the analysis is now carried out for cross- and curve-wired chiral metamaterials. The number of cells is 96 for the smaller cross-wired cells, 24 for the larger ones, and 24 for the curve-wired cells. The reflection parameters S_{11} for these structures are shown in Figure 5-42. These configurations are also of no practical use, as the all perform below the -10 dB limit.

Figure 5-42: Reflection parameter S_{11} for a monopole antenna surrounded by a chiral metamaterial cover at a distance $d = 1/4 \lambda_0$ from the antenna.

The cover to antenna separation distance is now changed to $d = 3/4 \lambda_0$, and the analysis is carried out for conventional metamaterials. The number of cells is 40 for the omega structure, 240 for the conventional SRR, and 40 for both single and double SRR. The reflection parameters S_{11} for these cases is shown in Figure 5-43. Table 5.19 lists some important figures-of-merit for these structures.

Figure 5-43: Reflection parameter S_{11} for a monopole antenna surrounded by a conventional metamaterial cover at a distance $d = 1/4 \lambda_0$ from the antenna.

Table 5.19: Figures-of-merit for a monopole antenna surrounded by a conventional metamaterial cover at a distance $d = 1/4 \lambda_0$ from the antenna.

Type	f_{r1} (GHz) / S_{11} (dB)	G_{fr1} (dB)	η_{fr1}	Z_{infr1} (Ω)	f_{r2} (GHz) / S_{11} (dB)	G_{fr2} (dB)	η_{fr2}	Z_{infr2} (Ω)
Monopole	8.23 / -18.50	3.79	1	31.25	---	---	---	---
Omega	7.38 / -29.73	4.77	1	25.48	7.80 / -22.45	5.56	0.94	28.03
Conventional SRR	8.53 / -23.08	6.19	0.98	28.37	---	---	---	---
Type	f_{r3} (GHz) / S_{11} (dB)	G_{fr3} (dB)	η_{fr3}	Z_{infr3} (Ω)				
Omega	8.33 / -12.85	2.77	0.93	38.27				

As can be seen, only the omega structure and the conventional SRR perform below the -10 dB limit. They present an improved gain, maintaining nearly the same efficiency and improved input impedance for f_{r1} and f_{r2} . The radiation patterns corresponding to these structures are shown in Figure 5-44.

Figure 5-44: Simulated radiation patterns for a monopole antenna surrounded by a conventional metamaterial cover at a distance $d = 1/4 \lambda_0$ for the resonant frequencies f_{r1} , f_{r2} and f_{r3} . a) Omega, b) Conventional SRR.

This chapter presented a detailed analysis of the monopole antenna assisted by a metamaterial cover of different geometries. It was shown how these structures affect important antenna parameters such as gain, return loss (S_{11}), and radiation efficiency. Also, it was shown how the impedance matching, a fundamental parameter to design an antenna system, can be favored in the design of the monopole antenna.

The worst results occurred when the structure was placed near the irradiating element ($d = 1/4 \lambda_0$), i.e., when the structure is in the region immediately surrounding the antenna. In this region the fields are predominately reactive, which means the electric and magnetic fields are out of phase by 90 degrees to each other. And for propagating or radiating fields, the fields are orthogonal (perpendicular) to each other, but in phase.

When the structure was placed slightly distant ($d = 1/2 \lambda_0$, $3/4 \lambda_0$ and $4/5 \lambda_0$) from the irradiating element, good results in terms of gain, efficiency and impedance matching were observed. And it was also observed the appearance of new operating resonance frequencies, which is certainly attractive for telecom applications where different frequencies are desirable with just a single antenna.

6 ADDITIONAL ANTENNA DESIGN

In this chapter a novel patch antenna design is introduced using the concept of concentric rings applied to the ground plane. This approach shows great potential for the design of antennas with multiple resonances.

6.1 PATCH ANTENNA DESIGN

The demand for microstrip antennas has increased rapidly in the past decades, mainly due to their attractive properties, such as low profile, compactness, low cost, light weight, and easy of fabrication [100]. The selection of operating frequencies is essential for antenna design. Antennas presenting narrow bandwidths have been widely used in the military and in commercial applications. This narrowband characteristic is highly desirable in government security systems, high speed missiles, rockets and weaponry [100]. Also, they are used in commercial applications, such as radio services and the Global Positioning System (GPS).

To achieve these narrow bandwidths, different diode switch combinations are often used [43]. With a simpler design and without additional components, we present a novel elliptical multi-resonant patch antenna with Fresnel Zone Plate (FZP) inspired concentric rings ground plane. In this approach the resonant frequencies are controlled simply by rescaling the radii of the FZP [101]. Simulation and experimental results show that this antenna design has potential not only for military but also for civil applications.

The layout of the proposed elliptical patch antenna is shown in Figure 6-1 *a*), where the minor radius is $r_a = 7.5$ mm and the major radius is $r_b = 10$ mm. The substrate is FR-4, with relative permittivity $\epsilon_r = 4$ and thickness $w = 1.6$ mm. The substrate length is $L = 40$ mm and width $W = 35$ mm. The feed line length is $L_f = 3.5$ mm and width $W_f = 6.77$ mm.

Each radius r_n (see Figure 6-1 *b*) and *c*)) in the ground plane is calculated using the FZP equation, given by [102]

$$r_n = \sqrt{n_z f_l \lambda_0 + \left(\frac{n_z \lambda_0}{2}\right)^2} \quad (43)$$

where n_z is the zone number, f_l is the focal length ($f_l = \lambda_0/4$), and λ_0 is the wavelength at the operating frequency $f_o = 7.5$ GHz.

Figure 6-1: Layout of the proposed antenna: a) Elliptical patch antenna (front side); Concentric rings ground plane (back side) - b) for $n = 9$, and c) for $n = 17$.

The simulation results show that the performance of the antenna can be more easily improved if the FZP radii are rescaled as follows: for $n = 17$, the new radii are $r_n/17$, and for $n = 9$, the radii are $r_n/11$. This guarantees that both cases will have approximately the same external diameter on the FR-4 substrate.

The numerical analysis is carried out with the HFSS software. It is used lumped port element to feed the transmission line with a resistance of 50Ω and reactance of 0Ω . All metal claddings are set as a sheet of perfect electric conductor. The frequency range of interest goes from 10 MHz to 20 GHz.

6.2 PATCH ANTENNA FABRICATION

To fabricate these structures, a heat transfer technique is used to transfer the antenna design to the printed circuit board, then iron perchlorate is used as a corrosion agent to

remove copper and maintain the antenna design, and finally SMA connectors are soldered to the circuit and used as feeding element for the antenna. More details about the fabrication process can be found in APPENDIX D, which is the same process used to fabricate the metamaterials. Figure 6-2 presents the fabricated patch antennas.

Figure 6-2: Fabricated patch antenna on FR-4 substrate with copper cladding: a) Elliptical patch (front side); Concentric rings ground plane (back side) - b) for $n = 9$, and c) for $n = 17$.

A vector network analyzer (VNA) is then used to measure the reflection parameter (S_{11}). More information regarding the measure can be found in APPENDIX D. The VNA is used in Trace mode to measure the radiation pattern. The measuring setup is shown in Figure 6-3, which includes a horn antenna as the transmitter (T_x), and the receiving antenna (R_x) or the antenna under test (AUT) placed on a tower with a turntable, to collect data at any desired angle.

Figure 6-3: Receiving antenna tower with turntable.

6.3 PATCH ANTENNA RESULTS

The simulated and measured return loss (S_{11}) of the elliptical patch antenna with conventional ground plane is presented in Figure 6-4 a), and the corresponding radiation pattern shown in Figure 6-4 b).

Figure 6-4: a) Return Loss (S_{11}) versus frequency (simulated and measured results) for the elliptical patch with conventional ground plane. In the inset is shown the magnitude of electric current density J ; b) E-plane and H-plane radiation pattern for $f_r = 7.47$ GHz.

It can be observed a good agreement between the simulated and measured results for the elliptical patch with conventional ground plane, especially in the lower frequency bands, from 0 to 10 GHz. Above this range, numerical and experimental results disagree due the transmitter antenna used (adequate for the X-band only).

In the inset of Figure 6-4, it is possible to see the magnitude of the electric current

density J , which represents the amount of current flowing through the area of the antenna. In this particular case, we can observe the current flowing at $f_r = 7.47$ GHz, showing the antenna is radiating in this resonant frequency.

Simulated and experimentally measured return loss (S_{11}) results for the elliptical patch antenna with $n = 9$ concentric rings ground plane is shown in Figure 6-5 a), and its corresponding radiation pattern shown in Figure 6-5 b). As can be seen, the agreement between simulated and measured results is good for almost the entire frequency range.

Figure 6-5: a) Return Loss (S_{11}) versus frequency (simulated and measured results) for an elliptical patch with concentric rings ground plane for $n = 9$. The inset shows the magnitude of electric current density J ; b) E-plane and H-plane radiation pattern for $f_r = 7.98$ GHz.

In the inset of Figure 6-5 it is presented the magnitude of the electric current density J . We can observe the current flowing at $f_r = 7.98$ GHz and it is possible to see that the concentric rings placed on the ground plane are working as radiating elements.

Next, the number of concentric rings on the ground plane is increased to $n = 17$. The simulated and experimentally measured return loss results (S_{11}) are shown in Figure 6-6 a), with the corresponding radiation pattern shown in Figure 6-6 b).

Figure 6-6: a) Return Loss (S_{11}) versus frequency (simulated and measured results) for the elliptical patch with concentric rings ground plane for $n = 17$. In the inset is the plot of the magnitude of electric current density J ; b) E-plane and H-plane radiation pattern for $f_r = 7.51$ GHz.

In the inset of Figure 6-6 it is possible to see the magnitude of the electric current density J . We can observe the current flowing at $f_r = 7.51$ GHz, showing a strong constructive interaction between the concentric rings placed on the ground plane, since they are closer to each other; and they are also working as radiating elements.

Even in this case, the agreement between simulated and measured results is still good, especially in the 0 to 10 GHz range (well within the designed operating band of the transmitter antenna). Above this range, the results are not expected to show a good convergence, since it is not within the operating band of the transmitter antenna.

Table 6.1 summarizes the main results for the proposed antenna at the resonant frequencies (f_r) shown in Figure 6-4, Figure 6-5 and Figure 6-6. This table helps visualize the improvements achieved with this design.

Table 6.1: Results of the proposed antennas.

Ground plane type	f_r (GHz)	Gain at f_r (dB)	Efficiency at f_r (%)
Conventional	7.47	4.21	52.3
Concentric rings ($n = 9$)	7.98	7.47	86.6
Concentric rings ($n = 17$)	7.51	7.23	80.6

For example, comparing the elliptical antenna with the conventional ground plane and with concentric rings ($n = 9$), it can be observed that for approximately the same frequency (7.47 GHz for conventional ground plane and 7.98 GHz for the concentric rings), the antenna gain increases by 3.02 dB, which represents an improvement of almost 50% in terms of power. Also, an increase of 28.3% in the efficiency (from 52.3% to 80.6%) is observed, which represents a 54.11% improvement of the original value.

Therefore, the results show that the frequency resonances of the proposed patch antenna depends only on the number of radii n of the concentric rings, making it a simple, low cost, easy of fabrication, and innovative controllable system to achieve different frequency requirements in antenna designs.

CHAPTER 7

7 CONCLUSIONS

In this work, it has been proposed a new approach to improve the performance characteristics of a conventional antenna, particularly the monopole antenna. The approach is based on the design of a metamaterial cover to be inserted around a monopole antenna.

In order to optimize the design, different types of metamaterial cells were investigated, such as conventional cells (single SRR, double SRR, omega), and chiral cells (2D chiral, cross- and curve-wired cells). These metamaterial cells were then arranged on a substrate to form an eight-sided cylinder-type cover to be inserted around the antenna.

Intensive numeric efforts were carried out to obtain a nearly optimized structure capable of improving the performance characteristics of the monopole antenna operating in the microwave regime. Different substrates and substrate thicknesses were carefully investigated. In addition, different metamaterial cover diameters were investigated in order to obtain the adequate geometry. Numerical simulations indicate the alumina substrate with both thicknesses can be appropriate choices for the antenna type and operating frequency adopted.

It was shown that when the metamaterial cover was placed near the antenna ($d = 1/4 \lambda_0$), no improvements was observed in the overall system performance. However, when the cover was placed slightly distant ($d = 1/2 \lambda_0$, $3/4 \lambda_0$ and $4/5 \lambda_0$) from the irradiating element, it was clearly observed very good results in terms of gain, return loss (S_{11}), efficiency and impedance matching, due to the constructive interaction of the electromagnetic fields radiated by the antenna with metamaterial cover. It was also possible to detect new operating resonances, which is certainly attractive for telecom applications, where different frequencies are desirable with the use of just a single antenna.

In the experimental characterization, the cover was placed at the maximum distance from the antenna. The measured results showed good agreement when compared with the simulated ones, but we could also observe a frequency shift (relative to the simulated results), that may be due to the substrate characteristics and some measurement errors.

We showed that it is possible to control the response of the monopole antenna with the correct choice of metamaterial, which is directly related to its shape, size, number of inclusions, and substrate type.

We also investigated a new patch antenna design composed of multiple concentric rings whose radii are obtained directly from Fresnel Zone Plate equations, with the concentric rings defined directly on the ground plane. The results showed that the frequency resonances of the proposed patch antenna depends only on the number of radii n of the concentric rings, making it a simple, low cost, easy of fabrication, and innovative controllable system to achieve different frequency requirements in antenna designs.

Finally, we hope that the application of this new metamaterial cover and concentric ring-based ground plane can be successfully applied to other types of antennas in order to improve their performance characteristics.

7.1 FUTURE PERSPECTIVES

In this section we present some suggestions to this work that has the potential to contribute to improve the performance of monopole and other types of antennas. The idea is to design a structure that can potentially change the antenna response and at the same time protect it from the external environment. To this aim, we suggest the metamaterial radome, as presented in Figure 7-1.

Figure 7-1: Example of metamaterial radome structure covering a monopole antenna.

The radome protects the antenna without affecting its electromagnetic properties. The radome is an acronym for the combination of the words “radar” and “dome”, which is a cover

placed on the antenna. Since it is completely sealed, it can hide the antenna from human sight. Generally, it is fabricated with materials that minimally attenuate the electromagnetic signal transmitted or received by the antenna, such as fiberglass and PTFE-coated, being transparent to radio frequency waves.

The design of an antenna radome using metamaterials can enhance the antenna gain, focus the microwave transmitted by the antenna, reducing the side lobes, and also compensate the effects of radome Boresight Error (BSE), which is a bending of the angle of arrival of a received signal relative to its actual angle of arrival, and it stems primarily from distortions of the electromagnetic wave front, as it propagates through the dielectric radome wall [103].

Therefore, with the proper design and the proper materials and metamaterials, this structure has the potential to be implemented for different applications such as ground, maritime, vehicular, aircraft and missile systems.

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APPENDIX A – PARAMETER RETRIEVAL

A.1 PARAMETER RETRIEVAL OF CONVENTIONAL METAMATERIALS

After obtaining the transmission S_{21} and reflection (return loss) S_{11} coefficients, we can obtain the impedance z of the metamaterial and the refractive index n . They are calculated with the following equations [46]:

$$z = \pm \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (44)$$

$$n_{\pm} = \frac{i}{k_0 w_u} \log \left[\frac{1}{S_{21}} \left(1 - \frac{z - 1}{z + 1} S_{11} \right) \pm 2m\pi \right], \quad (45)$$

where $k_0 = \frac{2\pi f_0}{c}$ is the wave number, w_u is the width of each metamaterial cell, and m can be any integer.

The sign of the square root in (44) and the multi-branches in (45) need to be chosen carefully, according to the energy conservation principle [46]. Since the metamaterial is assumed here as a passive medium, the signs in (44) and (45) are determined by the requirement that the real part of the impedance z and the imaginary part of n must be positive [46],[65], i.e.,

$$\text{Re}\{z\} \geq 0 \quad (46)$$

$$\text{Im}\{n\} \geq 0. \quad (47)$$

Once this requirement is fulfilled, z and $n_{\pm}$ can be obtained. Then, the dielectric permittivity ε and magnetic permeability μ of these metamaterials are easily found, using the relations given by [66]

$$\varepsilon = \frac{n}{z} \quad (48)$$

$$(49)$$

$$\mu = n \cdot z.$$

These equations are then implemented using a Matlab routine. Thus, it is possible to trace all the curves for the four parameters of equations (44), (45), (48), and (49) to verify the electromagnetic response of each cell.

A.2 PARAMETER RETRIEVAL OF CHIRAL METAMATERIALS

A similar method is used to extract the parameters from chiral metamaterials. After the simulations using HFSS, we obtain solutions of the electromagnetic wave in chiral media. These solutions are two circularly polarized electromagnetic waves, i.e., the right-handed circularly polarized wave (RCP, represented by the signal +) and the left-handed circularly polarized wave (LCP, represented by the signal –) [45].

From the transmission results (S_{21}) acquired in HFSS, four linear transmission coefficients (T_{xx} , T_{yx} , T_{xy} , and T_{yy}) are obtained. Figure A-1 presents how the waves are polarized in HFSS with respect to x and y to obtain the linear transmission coefficients.

Figure A-1: Linear transmission coefficients.

Then, these coefficients are converted into circular transmission coefficients (T_{++} , T_{+-} , T_{-+} , and T_{--}) using the following matrix [45]:

$$\begin{pmatrix} T_{++} & T_{+-} \\ T_{-+} & T_{--} \end{pmatrix} = \frac{1}{2} \times \begin{pmatrix} (T_{xx} + T_{yy}) + i(T_{xy} - T_{yx}) & (T_{xx} - T_{yy}) - i(T_{xy} + T_{yx}) \\ (T_{xx} - T_{yy}) + i(T_{xy} + T_{yx}) & (T_{xx} + T_{yy}) - i(T_{xy} - T_{yx}) \end{pmatrix} \quad (50)$$

These coefficients are obtained to fully characterize the response of the chiral cell. But it is important to note that the cross coupling transmission (T_{+-} and T_{-+}) are negligible [45], and we will work with both coefficients $T_{++} = T_{RCP}$ and $T_{--} = T_{LCP}$. Now, from equation (50), we have that

$$T_{++} = \frac{1}{2} \left((T_{xx} + T_{yy}) + i(T_{xy} - T_{yx}) \right) \quad (51)$$

$$T_{--} = \frac{1}{2} \left((T_{xx} + T_{yy}) - i(T_{xy} - T_{yx}) \right) \quad (52)$$

The impedance z and the refractive index n can be calculated with the following equations [46]

$$z = \sqrt{\frac{(1 + S_{11})^2 - T_+ T_-}{(1 - S_{11})^2 - T_+ T_-}} \quad (53)$$

$$n_{\pm} = \frac{i}{k_0 w_u} \log \left[\frac{1}{T_{\pm}} \left(1 - \frac{z - 1}{z + 1} S_{11} \right) \right] \quad (54)$$

where n_+ refers to RCP and n_- to LCP, $T_{\pm} = T e^{\pm i \kappa k_0 d}$, S_{11} is the reflection coefficient.

Now, it is possible to calculate the chiral parameter and the refractive index [46]:

$$\kappa = \frac{n_+ - n_-}{2} \quad (55)$$

$$n = \frac{n_+ + n_-}{2}. \quad (56)$$

The equations used to find permittivity and permeability are the same shown in equations (48) and (49), respectively.

APPENDIX B – SCATTERING PARAMETERS

The S-parameters use the reflection and transmission coefficients of a system based on a model that serves as a basic building block: the two-ports network, as shown in Figure B-1. The incident signal on each port is referred to a , and the output signal of each port is referred as b . The first (P1) and second (P2) ports, have inputs a_1 and a_2 , and outputs b_1 and b_2 , respectively.

Figure B-1: Representation of the two-port network.

As can be seen in Figure B-1, the S-parameters have subscript numbers, where the first number refers to the output port and the second number refers to the incident port (input). For example, S_{21} represents the response of P2 due to a signal from P1. For this two-ports network, the S-parameters can be represented by a matrix that indicates a relation between the transmitted and reflected waves, given by [81]

$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix}. \quad (57)$$

Expanding the matrix in linear equations, we have

$$b_1 = S_{11}a_1 + S_{12}a_2 \quad (58)$$

$$b_2 = S_{21}a_1 + S_{22}a_2. \quad (59)$$

Equations (58) and (59) relate the incident and reflected waves at each port of the network in terms of S-parameters (S_{11} , S_{12} , S_{21} and S_{22}). A wave incident on a_1 results in waves that can come out on b_1 or b_2 . The same occurs with a wave incident on a_2 .

However, according to the definition of S-parameters, when port 2 is terminated in a load equal to the system impedance (Z_0), b_2 is completely absorbed, leading to $a_2 = 0$. This can be proven by the maximum power transfer theorem. Thus, it follows that

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} \quad (60)$$

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} \quad (61)$$

Following the same reasoning, when port 1 is terminated in a load Z_0 , then, $a_1 = 0$, so

$$S_{22} = \left. \frac{b_2}{a_2} \right|_{a_1=0} \quad (62)$$

$$S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0} \quad (63)$$

In the case of metamaterials, when a plane wave is incident inside these structures, the reflection coefficients (S_{11} and S_{22}) and transmission coefficients (S_{12} and S_{21}) are obtained, as shown in Figure B-2.

Figure B-2: Representation of a two-port network model with metamaterial inclusion; a) Incident wave in P1: S_{11} and S_{21} parameters; b) Incident wave in P2: S_{12} and S_{22} parameters.

B.1 DEFINITIONS USING SCATTERING PARAMETERS

Some parameters of antenna theory can be obtained using the scattering parameters. Considering a two-port network in which port 1 is the input and port 2 is the output, which is the most common convention, some of these parameters can be obtained as listed below.

a) Complex linear gain (G)

The complex linear gain is simply the voltage gain as a linear ratio of the output voltage divided by the input voltage, and is given by

$$G = S_{21} \quad (64)$$

b) Scalar linear gain ($|G|$)

The linear scaling gain has the same principle of the complex gain, but the phase is not relevant. This gain is given by

$$|G| = |S_{21}| \quad (65)$$

c) Scalar logarithmic gain (g)

This gain is most used than the scalar gain and the quantity is given in decibel (dB). A positive value is understood as a gain, effectively. However, a negative value is understood as

a loss. The expression that defines this gain is given by

$$g = 20 \log_{10} |S_{21}| \quad (66)$$

d) Input Return Loss (RL_{in})

This is one of the most important and one of the most used parameters in antenna-related researches. The input return loss can be defined as the measure of how close is the value of the impedance of the device under test with respect to the nominal impedance of the measuring equipment, such as a vector network analyzer. It is so called because it is analyzed in the input port (port 1) of the equipment, and one can get the S-parameter responsible for reflection or return loss (S_{11}). This parameter is expressed in dB and is given by

$$RL_{in} = 10 \log_{10} \left| \frac{1}{S_{11}^2} \right| = -20 \log_{10} |S_{11}| \quad (67)$$

e) Output Return Loss (RL_{out})

This loss has the same definition of the RL_{in} , but is analyzed at the output port (S_{22}). Thus, it follows that

$$RL_{out} = 10 \log_{10} \left| \frac{1}{S_{22}^2} \right| = -20 \log_{10} |S_{22}| \quad (68)$$

APPENDIX C – CONFIGURATIONS ADOPTED IN THE SOFTWARE

As already mentioned, metamaterials are formed by arrays of resonator elements. To simulate this effect, periodic boundary conditions (PBC) are set in the software. First of all, it is modeled a single metamaterial cell in HFSS. Then, these periodic boundary conditions are applied, resulting in an analysis of an infinite array of double periodicity (in both x - y and z - y planes), as will be explained later.

C.1 CONVENTIONAL METAMATERIALS

The SRR was used as an example of the characterization of conventional metamaterials, and the presented configurations are used for all conventional cells adopted in this work. As already clarified, SRR is composed of two conductive elements, resonator rings on one side and a wire on the other. They are designed to achieve a band with a negative refractive index at microwave frequencies. In Figure C-1 it is presented the material used for the simulation of a SRR unit.

Figure C-1: Example of the structure and the materials used in a metamaterial cell.

For the conventional metamaterial cells, boundary conditions were used with walls of perfect electric conductor (PEC) tangential to the electric fields, and perfect magnetic conductor (PMC) materials, as presented in Figure C-2.

Figure C-2: Boundary conditions used in conventional metamaterials; (a) PEC walls; (b) PMC walls.

The PEC boundary condition forces the electric field to be perpendicular to the selected surface. This boundary condition can represent metal surfaces, an antenna ground plane, cavity walls, among others. On the other hand, the PMC boundary condition forces the electric field to be tangential to the surface and, consequently, forces the magnetic field to be perpendicular to the surface. To better understand this arrangement, Figure C-3 illustrates how the electric field acts on each of these boundary conditions.

Figure C-3: Electric field in the boundary conditions: (a) PEC; (b) PMC.

A plane wave is used as the excitation source on the side walls of the conventional metamaterial cells, as presented in Figure C-4. This wave is used on the sides, because it excites the electromagnetic field correctly.

Figure C-4: Excitation ports (plane wave type) in conventional metamaterials.

C.2 CHIRAL METAMATERIALS

The structure known as crossed-wire, presented by Zhou et al. [45], is used as an example of the characterization of chiral metamaterials. The configurations shown in this section are used for all chiral cells adopted in this work. Figure C-5 shows the materials used for the metamaterial simulation, which are similar to conventional metamaterials.

Figure C-5: Example of the structure and the materials used in a chiral metamaterial cell.

For chiral metamaterials, Floquet ports were used as excitation. This port type is used exclusively for planar periodic structures [58]. The Floquet theory states that the fields are equal in the boundary directions, providing periodic continuity. In HFSS, Floquet port type should be used only with “Master/Slave” walls. These boundary conditions are necessary to define an “infinite” space in the directions that will be considered periodic. In the case of the unit cell shown in Figure C-6, it has periodicity in both y and x direction.

Figure C-6: Boundary conditions used in chiral metamaterials; (a) Master/ Slave 1 (periodicity in y); (b) Master/Slave 2 (periodicity in x).

The Floquet excitation ports are allocated on the upper and lower walls parallel to the unit cell, as presented in Figure C-7. Therefore, it simulates a plane wave at normal incidence on the metamaterial.

Figure C-7: Floquet excitation ports allocated on chiral metamaterials.

C.3 MESHING

HFSS generates an initial mesh that includes an approximation of the surface according to the structure modeled in software. If necessary, automatic repairs are generated to obtain an accurate mesh representation of the created model [88]. Using the mesh result, the software computes the electromagnetic fields that exist within the structure when it is excited with the operation frequency. This automatic and adaptive refinement is effective at generating an array of finite elements to discretize the structure. An example of the mesh generation in the crossed wire structure is presented in Figure C-8.

Figure C-8: Example of mesh generation in a structure in HFSS. The coarse resolution shown is just an illustration.

When is computationally possible (in terms of computing resources), the mesh used for the metamaterial cells can be assigned using HFSS. To restrict the length of the tetrahedral element inside or on the object, first one needs to select this object and type the maximum length of elements (MLE). The default value is set to 20% of the maximum edge lengths of the bounding boxes of each selected object's faces. The software refines the element edges inside the object until they are equal to or less than this value.

C.4 THE HFSS SOLUTION PROCESS

To calculate the S-matrix associated with a structure with ports, HFSS processes the solution as following:

- 1) Divide the structure into a finite element mesh;
- 2) Compute the modes on each port of the structure;
- 3) Compute the full electromagnetic field pattern inside the structure, assuming that one mode is excited at a time;
- 4) Compute the generalized S-matrix from the amount of reflection and transmission that occurs.

The resulting S-matrix allows the magnitude of transmitted and reflected signals to be computed directly from a given set of input signals, reducing the full 3D electromagnetic behavior of a structure to a set of high frequency circuit parameters.

APPENDIX D – FABRICATION AND MEASUREMENTS

This appendix will briefly describe the fabrication process of the antenna and the metamaterials, and also the measurements realized in the Laboratory of Microwave at the Engineering School of São Carlos / University of São Paulo.

D.1 MONOPOLE ANTENNA

After the design of the antenna was numerically validated, it was possible to begin the manufacturing process. For this purpose, the following materials were used:

- Copper wire (diameter 0.5 mm, height 8.1 mm);
- SMA (SubMiniature Version A) connector;
- Substrates used as ground plane (FR4 with one metallic copper layer with width 60 mm x 60 mm).

With the materials available in the Laboratory of Microwave, it was possible to fabricate the antennas. In Figure D-1, one can observe the monopole antenna fabricated with the ground plane on FR4 substrate with a copper layer in one side.

Figure D-1: Monopole antenna with FR-4 ground plane (with copper layer in one side).

D.2 METAMATERIALS

Planar metamaterials designed to operate in the microwave range (GHz) have the advantage of being easy to fabricate. Due to the millimeter magnitude order of the wavelength in this range, the unit cells and metallic resonators present dimensions of the same magnitude and, therefore, such materials can be fabricated by traditional manufacturing processes of Printed Circuit Boards (PCB). In this technique, flat plates of dielectric substrates with surfaces covered by a thin metal layer are used, such as the FR-4.

The fabrication process of a metamaterial plate is basically done in two steps: thermal transfer and chemical corrosion. First, the design of the metamaterial structures is printed in a laser printer using a specific transfer paper. As the layout image is reversed with the transfer, the layout should be mirrored prior to printing. The designed pattern (layout) of the metamaterials is transferred to the metal using the thermal transfer technique. In the case of the alumina substrate, it does not have a metal copper on it and there were difficulties in depositing copper material in alumina substrate. Regardless of this, we found a solution to this problem: to use a tape comprising flat copper coated with adhesive in the back. This is an extremely interesting solution as it is a commercial ribbon, the same used for grounding, and electromagnetic shielding in equipment and components. Then, we used an adhesive copper tape to transfer the drawing for the metal using thermal transfer. The thermal transfer can be done using a heat press, as showed in

Figure D-2, where we placed face to face the metamaterials in the printed paper with the copper tape, during approximately 60 seconds with a temperature of 190°C. At the end of this process, the drawings are transferred to the copper and the metal is partially covered by the design of the metamaterial structures. Finally, we have the metamaterials ready to be placed on the substrate. In the case of biplane metamaterials, with the structures in both sides of the plate, the same transfer procedure must be done for the opposite side.

Figure D-2: Heat press used for thermal transfer.

With this step accomplished, it is necessary to place the metallic tape on the substrate. We used an alumina plate of the width of 25 mm x 25 mm. Finally, we have the alumina plate with copper tape, and we did the same procedure for the conventional metamaterials, for the 2D chiral metamaterials and for the chiral metamaterials, as can be seen in Figure D-3, Figure D-4 and Figure D-5, respectively.

Figure D-3: Alumina plate with adhesive copper tape with conventional metamaterials: a) SRR structure, b) Omega structure.

Figure D-4: Alumina plate with adhesive copper tape with 2D chiral metamaterials: a) $\alpha = 0^\circ$, b) $\alpha = 45^\circ$, and c) $\alpha = 90^\circ$.

Figure D-5: Alumina plate with adhesive copper tape with chiral metamaterials: a) Cross-wired structure with $w = 3.75$ mm, b) Cross-wired structure with $w = 7.5$ mm, and c) Curve-wired structure.

Subsequently, the alumina plate, now with the metal and the metamaterials layout, is subjected to a chemical etching process in which the plate is immersed in an acid solution of ferric chloride, as shown in Figure D-6. The acid is corrosive to the exposed copper and does not affect the substrate nor the printing ink.

Figure D-6: Alumina plates immersed in an acid solution of ferric chloride.

This process lasts from thirty minutes to one hour and when it is done, only the remaining metallic structures are placed on the substrate, where the paint was initially. Figure D-7 shows the SRR structures, just to illustrate how the alumina plate is after the chemical corrosion.

Figure D-7: Alumina plate with metamaterials layout after the chemical corrosion, for illustration.

This process was carried out on eight alumina plates with all metamaterials layout. Then we arranged these plates to form an octagon, as shown in Figure D-8 for the conventional metamaterials, Figure D-9 for the chiral metamaterials and Figure D-10 for the 2D chiral metamaterials. In order to maintain this structure in the correct position, it was necessary to use an adhesive tape on the opposite side of the metamaterial structures.

Figure D-8: Alumina octagon cover: a) Double SRR cells, b) Single SRR cells and c) Omega cells.

Figure D-9: Alumina octagon cover: a) Cross-wired cells for $w_u = 3.75$ mm, b) Cross-wired cells for $w_u = 7.5$ mm and c) Curve-wired cells.

Figure D-10: Alumina octagon cover with 2D chiral metamaterials cells: a) $\alpha = 0^\circ$, b) $\alpha = 45^\circ$ and c) $\alpha = 90^\circ$.

This octagon composed of metamaterials was inserted on the ground plane of the monopole antenna, as shown in Figure D-11, for illustration purposes.

Figure D-11: Cover inserted on the ground plane of the monopole antenna, for illustration. a) Double SRR metamaterials, b) Zoom of the monopole antenna and the 2D chiral metamaterials cover for $\alpha = 90^\circ$.

For experimental measurements of the reflection parameter (S_{11}), it was used the Rohde & Schwarz Vector Network Analyzer (VNA) model ZVA40 available in the Microwave Laboratory, as shown in Figure D-12. This analyzer covers the frequency range 10 MHz to 40 GHz and can also be used to measure the transmission parameters.

Figure D-12: Rohde & Schwarz ZVA40 vector network analyzer.

In addition to the VNA, calibration kits and compatible cables are required so the measured system is less susceptible to errors. In our laboratory we have available a calibration unit ZV-Z54 (Figure D-13 a) that covers the same frequency range as the VNA and performs the calibration automatically. The cable used has SMA (Subminiature version A) connector (Figure D-13 b).

Figure D-13: a) Calibration kit and b) Cable with SMA connector.

Before starting the testing procedures with the VNA, it is necessary to perform its calibration. First, the operating frequency band is selected. Then, we plug the calibration kit, which performs the calibration of the Port 1 (P1) and Port 2 (P2), for both reflection and transmission parameters.

Once the calibration is performed, the characterization of the antenna can be done. For testing procedures, the setup shown in Figure D-14 was used. As can be seen, the antenna under test (AUT) is connected to one of the network analyzer ports and then, the results can be collected for further analyzes.

Figure D-14: a) Setup for the antenna characterization and b) S-parameters measurement in the Laboratory with the omega structure cover, just for illustration.