Chapter 2

Theory and literature survey on Microwave Antennas

This chapter is intended for presenting the research carried out to find a radiating structure that fulfils all the requirements. In the following lines, a total number of 3 antenna structures are studied plus several other that will be just shown and ruled out because don’t achieve the objectives.

This chapter also contains a basic theory to explain the behaviour of the antennas. With this theory it will be easier to understand the results obtained in the simulations as well as the special properties of each antenna.

2.1. Microstrip antennas

All the studied antennas have a microstrip structure. Other structures like helix or conformal antennas have been ruled out due to several drawbacks. The helix antennas are difficult to integrate in an array and the conformal antennas are neither easy to build nor to integrate.

2.1.1. Introduction

A microstrip patch antenna consists of a very thin metallic patch (usually gold or copper) placed a small fraction of a wavelength above a conducting ground plane, separated by a dielectric substrate. Microstrip antennas have numerous advantages, they are lightweight, they can be designed to operate over a large range of frequencies (1-40 GHz), they can easily be combined to form linear or planar arrays, and they can generate linear, dual, and circular polarizations. These antennas are inexpensive to fabricate using printed circuit board etching, which makes them very useful for integrated active antennas in which circuit functions are integrated with the antenna to produce compact transceivers. This section includes a summary of rectangular microstrip antennas radiation mechanism, modelling, and design.
2.1.2. Rectangular Microstrip Antennas

Rectangular and square patches are the most commonly used type of microstrip antennas. They can be used in numerous types of applications including circular polarization, beam scanning, radiation pattern control and multiple frequency operation. The basic antenna element is a thin conductor of dimensions $L \times W$ on a dielectric substrate of permittivity $\varepsilon_r$ and thickness $h$ backed by a conducting ground plane. This configuration is shown below:

![Figure 2.1. Microstrip antenna configuration](image)

2.1.3. Radiation Mechanism

Radiation from a microstrip antenna is determined from the field distribution between the patch and the ground plane. This can also be described as the surface current distribution on the patch. A patch, which is connected to a microwave source, has a charge distribution on the upper and lower surface of the patch as well as the ground plane. The patch is half wavelength long at the dominant mode, which creates the positive and negative charge distribution shown in Figure 2.2.

![Figure 2.2. Microstrip antenna charge distribution and current density](image)
The repulsive nature of those charges on the bottom surface of the patch pushes some charges around the side to the top causing current densities $J_b$ and $J_s$. The ratio $h/W$ is small, therefore the strong attractive forces between the charges cause most of the current and charge concentration remains underneath the patch. But also the repulsive force between positive charges creates a large charge density around the edges. The fringing fields caused by these charges are responsible for radiation. Figure 2.3 shows the fringing fields in a microstrip patch.

Figure 2.3. Fringing fields for the dominant mode in a rectangular microstrip patch

2.1.4. Microstrip Antenna Analytical Models

There are various ways to model a microstrip patch. This modelling is used to predict characteristics of a microstrip patch such as resonant frequency, bandwidth, radiation pattern, etc. In this section the transmission line model and is presented. This model is based on some assumptions, which simplify the calculations at the cost of less accuracy. There are other models that provide more accuracy such as the cavity model and the full-wave model but are also more complicated to analyze.

The transmission line model is the simplest model and is restricted to rectangular microstrip antennas. This model considers the patch as a transmission line of width $W$ with two radiating slots on each end. For a desired frequency $f_0$, the width $W$ can be estimated using:

$$W = \frac{\lambda_0}{2} \sqrt{\frac{2}{\varepsilon_r + 1}}$$  \hspace{1cm} (2.1)

In this model the input impedance of a patch is the same as that of a transmission line with length $L$ and admittance $Y_c$. Each slot has an admittance of $Y_s = G_s + jB_s$ where the values for conductance $G_s$ and susceptance $B_s$ are given by:

$$G_s = \frac{W}{120\lambda_0} \left(1 - \frac{1}{24}(k_0 h)^2\right)$$  \hspace{1cm} (2.2)
\[ B_s = \frac{W}{120\lambda_0} \left(1 - 0.636 \ln \left(k_s h\right)\right) \quad (2.3) \]

Figure 2.4. Equivalent admittance circuit model

At resonance these slots are 180° apart, which is achieved by adjusting the length \( L \) slightly smaller than half a guided wavelength \( \lambda_g \). This adjustment accounts for the fringing fields at the radiating edges so that the susceptance of the two slots cancel each other out, leaving a purely resistive input admittance, so that:

\[ Y_{in} = 2G_s \quad (2.4) \]

Therefore at resonance the input resistance at the edge of the patch is:

\[ R_{in} = \frac{1}{Y_{in}} \quad (2.5) \]

The microstrip patch has an inhomogeneous configuration (air above and dielectric below), which can be replaced by a homogeneous configuration. This is done by introducing a new medium with an effective permittivity \( 1 < \varepsilon_{eff} < \varepsilon_r \), which has the same electrical characteristics (impedance and phase velocity) as the original medium. For \( W/h > 1 \), the effective dielectric constant is:

\[ \varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + 12 \frac{h}{W}\right)^{\frac{1}{2}} \quad (2.6) \]
The waveguide wave length is:

\[ \lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_{\text{eff}}}} \]  \hspace{1cm} (2.7)

The fringing fields around the edges of the patch make it appear electrically larger than its physical length by \(2\Delta L\). This size increase can be seen in Figure 2.5.

![Figure 2.5. Two slots model](image)

The value \(\Delta L\) is determined from the following equation:

\[ \Delta L = \frac{0.412h(\varepsilon_{\text{eff}} + 0.3)(\frac{W}{h} + 0.264)}{(\varepsilon_{\text{eff}} - 0.258)(\frac{W}{h} + 0.8)} \]  \hspace{1cm} (2.8)

The effective length is given by:

\[ L_{\text{eff}} = (L + 2\Delta L) \]  \hspace{1cm} (2.9)

The resonant frequency \(f_0\) (for dominant mode \(TM_{10}\)) is:

\[ f_0 = \frac{c}{2L_{\text{eff}}\sqrt{\varepsilon_{\text{eff}}}} \]  \hspace{1cm} (2.10)
Equation (2.10) can be used to find the value of $L_{\text{eff}}$, and knowing $\Delta L$ from the equation (2.9), length $L$ will be found. The E-fields components of the far field radiation intensity can be found from the following equations:

$$E_\theta (\theta, \phi) = \frac{-\sin(W_k) \cos(L_k) (\varepsilon_{\text{eff}} - \sin^2(\theta)) \cos(\phi)}{W_k (\varepsilon_{\text{eff}} - \sin^2(\theta) \cos^2(\phi))}$$  \hspace{0.5cm} (2.11)$$

$$E_\phi (\theta, \phi) = \frac{-\sin(W_k) \cos(L_k) \varepsilon_{\text{eff}} \sin(\theta) \cos(\phi)}{W_k (\varepsilon_{\text{eff}} - \sin^2(\theta) \cos^2(\phi))}$$  \hspace{0.5cm} (2.12)$$

Where $W_k$ and $L_k$ are:

$$W_k = 0.5W \sin(\theta) \sin(\phi)$$  \hspace{0.5cm} (2.13)$$

$$L_k = 0.5k(L + 2\Delta L) \sin(\theta) \cos(\phi)$$  \hspace{0.5cm} (2.14)$$

2.2. Microstrip antenna structures overview

At the beginning of the work, several types of antennas have been studied in order to find the radiating structure having the best possibilities to achieve our objectives. The most interesting ones will be explained in the next points.

2.2.1. Small dual-band circularly polarized square microstrip antenna for GPS application

In [TAK08] a small dual-band circularly polarized square microstrip antenna for GPS application is proposed. The antenna consists of a square patch with slits and four T-shaped elements. The T-shaped element is loaded at each slit of the square patch. The geometry of this antenna is presented in the Figure 2.6.
The width of the square patch is $W_T$. In order to achieve right handed circular polarization in the dual bands, the widths of the T-shaped elements along the $x$ axis are wider than those along $y$ axis, $d_b > d_c$, and the lengths are both $d_a$.

The relative dielectric constant and thickness of the dielectric substrate are $\varepsilon_r = 2.6$ and $h = 2.4$ mm. The antenna is excited at $x_0, y_0$ around the diagonal on the patch by a coaxial feeder through the dielectric substrate.

The Figure 2.7 shows the return loss and the axial ratio of the antenna at the center frequencies of the L1 and L2 bands of GPS. The bandwidths (return loss $\leq -10$ dB with axial ratio $\leq 3$ dB) are very narrow (approximately 2 MHz).
Looking at the results, we have to reject this antenna structure due to the narrow bandwidth achieved. Moreover, there is no information about the radiation pattern (symmetry, beamwidth…) of the antenna, and therefore is not worth to carry out a deep research on this structure in order to achieve a wider bandwidth (increasing $h$ and decreasing $\varepsilon_r$).

This structure was apparently interesting because can achieve circular polarization without requiring hybrids. In the following points we will notice how all the structures that achieve circular polarization adding perturbations to the patches have a narrow axial ratio bandwidth.

2.2.2. Single-Feed Multi-Band Antenna formed by Patch and Ring Antennas for GPS applications

In [ZHA08] a microstrip antenna consisting in a concentric circular patch and ring antenna has been developed. The geometry of the described antenna can be seen in Figure 2.8.

In this configuration a patch antenna is a feeding radiator. The inner ring resonates at the L1 band and the outer ring resonates at the L2 band. The coupling elements play the role of exciting the outer ring elements. Furthermore, a perturbation $a_1$ and $a_2$ are added in order to produce RHCP at L1 band. Similarly, perturbations $b_1$ and $b_2$ are added in order to obtain RHCP at L2 band.

![Figure 2.8. Geometry of the antenna](image)

In the Figure 2.9 the return loss and the axial ratio for both L1 and L2 bands are presented. The bandwidth (return loss $\leq -10\text{dB}$ with axial ratio $\leq 3\text{ dB}$) is about 2 MHz in both bands. This performance is similar to the antenna described on 2.2.1. However the complexity to tune this antenna is bigger due to the coupling elements and the 4 perturbations used to generate RHCP.
In this antenna we find a complex structure in order to create a dual-band circularly polarized antenna. However the performance achieved in terms of bandwidth is not enough to fulfil our requirements. Like in 2.2.1, the utilisation of parasitic elements to create the RHCP provides a narrow axial ratio bandwidth. The bandwidth could be improved by using thicker dielectrics with a low dielectric constant. However, the dimensions of the antenna would increase resulting in a narrow beamwidth.

### 2.2.3. Dual Circularly-Polarized Stacked Patch Antenna for GPS/SDMB

The structure described in [JUN08] is a dual-port-feed antenna consisting in low-profile stacked radiators to provide dual-frequency circular polarization characteristic. To maintain concentric geometry for optimum circular polarization, two via-holes were embedded to excite each antenna.
As illustrated in Figure 2.10, the proposed antenna is comprised of a corner truncated circular lower patch having ears at diagonal direction for the global positioning system (GPS), and a corner truncated circular upper patch for satellite digital multimedia broadcasting (SDMB). The additional ears for lower patch are to play an important role in the circular polarization. Both patches show right-handed circular polarization. A stacked patch configuration is used in this antenna in order to reduce the volume, but undesired coupling degrades the bandwidth as well as the axial ratio.

![Simulated and measured return losses of the described antenna](image)

Figure 2.11. Simulated and measured return losses of the described antenna

The Figure 2.11 shows the return loss of the antenna in both GPS and SMDB bands. The impedance matching bandwidth is wide enough for our objectives. However, the axial ratio presented in the Figure 2.12 has a narrow bandwidth in the SMDB band.

![Simulated and measured axial ratio of the described antenna](image)

Figure 2.12. Simulated and measured axial ratio of the described antenna

The most interesting characteristic provided by this antenna is the dual-port feeding. We were very interested in this property so we contacted the authors of [JUN08]. The main question was the isolation between the two ports at the GPS and SMDB bands. The authors answered that the isolation between ports are -5 dB at 1.57 GHz (GPS band) and -15 dB at 2.64 GHz (SMDB band) [COM08].

This level of isolation is not enough to consider that both ports are decoupled. This idea is very important for our objectives but there is still work to do in order to achieve better isolation.
2.2.4. Circularly Polarized Microstrip Antennas and Techniques

In this section, two main techniques in order to achieve circular polarization are going to be discussed. We have to take into account our needs in terms of bandwidth in order to choose the most appropriate one.

The first method that is going to be explained is the Single Feed Patch with two induced orthogonal modes with equal amplitude and in-phase quadrature. This can be accomplished by slightly perturbing a patch at appropriate locations with respect to the feed. All the perturbation configurations for generating circular polarization operate on the same principle of detuning degenerate modes of a symmetrical patch by perturbation segments as presented in Figure 2.13. The fields of a singly fed patch can be resolved into two orthogonal degenerate modes, 1 and 2. Proper perturbation segments will detune the frequency response of mode 2 such that, at the operating frequency $f_0$, it is of the same amplitude but 90° out of phase with respect to mode 1. Hence, the two modes satisfy the required condition for circular polarization radiation. As the frequency moves away from $f_0$, the axial ratio rapidly degrades while the input match usually remains acceptable. The actual detuning occurs either for one or both modes depending on the placement of perturbation segments.

![Figure 2.13. Amplitude and phase of orthogonal modes for single feed circularly polarized antennas](image)

The second method studied in this section is the Dual-Orthogonal Fed Circularly Polarized Patch. The fundamental configurations of a dual-orthogonal fed circularly polarized patch using either an external power divider or an external 90°-hybrid are shown in Figure 2.14.
The patch is usually square or circular. The dual-orthogonal feeds excite two orthogonal modes with equal amplitude but in phase-quadrature. Several power divider circuits that have been successfully employed for CP generation include the quadrature hybrid, the ring hybrid, the Wilkinson power divider, and the T-junction power splitter.

The quadrature hybrid splits the input in two outputs with equal amplitude but 90° out of phase. Other types of dividers, however, need a quarter-wavelength line in one of the output arms to produce a 90° phase shift at the two feeds. Consequently, the quadrature hybrid provides a broader axial ratio bandwidth.

In order to achieve the 4.2% of axial ratio bandwidth required in the low band will be necessary to use dual-orthogonal feeding with a 90°-hybrid to create the circular polarization.

2.3. A Dual-Band Circularly Polarized Aperture-Coupled Stacked Microstrip Antenna

2.3.1. Introduction

In [POZ97] is described the design and testing of an aperture-coupled circularly polarized antenna for global positioning satellite applications. The antenna operates at both L1 and L2 frequencies of 1575 and 1227 MHz.

The aperture coupled microstrip patch antenna has a number of desirable features compared to other feeding techniques, such as doubled substrate area, a non-contacting feed transition and shielding of the feed network from the radiating aperture.
It has also been demonstrated that aperture coupling can be used to substantially increase bandwidth beyond the few percent that is typically obtained with traditional patch designs. Increased bandwidth can be achieved with a single patch using a relatively thick low-dielectric constant substrate, or by using two (or more) stacked patches. Impedance bandwidths ranging from 20% to 35% have been obtained with these methods.

Circular polarization can be generated from aperture coupled elements by using off-center coupling apertures or with crossed slots. In both cases, two orthogonal linearly polarized modes are independently excited with equal amplitudes and a 90° phase shift, achieving 2-dB bandwidths of 25%. This antenna uses two stacked microstrip elements in order to obtain the dual-band coverage of the L1 and L2 bands of GPS.

After analysing the results obtained by Pozar in [POZ97] we decided to start with the design of this type of antenna, adapting it to our requirements, in order to see the capabilities of this structure. Before explaining the design carried out in Ansoft Designer 4.0, a basic theory about the coupling mechanism is explained in the following points.

### 2.3.2. Aperture-coupled Feeding

In this technique the feed network is separated from the radiating patch by a common ground plane. Energy is electromagnetically coupled through an aperture in the ground plane. This aperture is usually centered with respect to the patch where the patch has its maximum magnetic field. For maximum coupling it has been suggested that a rectangular slot parallel to the two radiating edges should be used. Two very similar coupling mechanisms take place, one between the feed line and the slot and another between the slot and the patch. The coupling amplitude can be obtained by the following equation:

\[
Coupling = \int \int \int \int M H dv \equiv \sin(\pi x_0/L)
\]  
(2.15)

In this equation, \( x_0 \) is the offset of the slot from the patch edge. This technique has several advantages, which makes it suitable for widespread applications in communication systems. Also, the isolation of the feed network from the patch reduces the spurious radiations and provides more space for the feed network, being suitable for phased arrays. Wider bandwidth can be achieved by adjusting the width and length of the coupling slot and using a thicker patch substrate. The feed substrate is usually thin with high permittivity, whereas the patch substrate can be thick with low permittivity. The geometry of a basic aperture-coupled microstrip antenna is shown in Figure 2.15:

There are several methods to analyze aperture-coupled microstrip antennas such as the integral equation approach, cavity model, transmission line model, model expansion method, and a hybrid approach. Here, we discuss the transmission line model because of its ability to arrive at a simpler design.
A simplified equivalent circuit of a slot coupled microstrip patch antenna is shown below. The microstrip patch is characterized by admittance $Y_{\text{patch}}$ and the aperture by admittance $Y_{\text{ap}}$.
The slot interrupts the longitudinal current flow in the feed line and the patch, resulting in a coupling between them. The coupling between the patch, aperture, and the feed line can be described by two impedance transformers. The $1:n_1$ transformer has a turns ratio approximately equal to the fraction of patch current intercepted by the aperture to the total patch current, so that:

$$n_1 = \frac{L_s}{W}$$ (2.16)

and,

$$n_2 = \frac{J_0(k_{es}W_s/2)J_0(k_{em}W_s/2)}{k_{es}^2 + k_{em}^2} \left[ \frac{k_{em}^2k_2\varepsilon_{ef}^2}{k_2\varepsilon_{rf}^2 \cos(k_1d_p) - k_1 \sin(k_1d_p)} \right. + \left. \frac{k_{es}^2k_1^2}{k_2\varepsilon_{rf}^2 \cos(k_1d_p) + k_1 \sin(k_1d_p)} \right]$$ (2.17)

where $J_0(\cdot)$ is the zeroth-order Bessel function and,

$$k_1 = k_0 \sqrt{\varepsilon_{rf} - \varepsilon_{es} - \varepsilon_{efm}}$$ (2.18)

$$k_2 = k_0 \sqrt{\varepsilon_{res} - \varepsilon_{rem} - 1}$$ (2.19)

$$k_{es} = k_0 \sqrt{\varepsilon_{efm}} \quad k_{em} = k_0 \sqrt{\varepsilon_{efm}}$$ (2.20)

The transmission line equivalent circuit of microstrip patch is shown below:

![Figure 2.17. Transmission line equivalent circuit of a rectangular patch fed by a slot](image-url)
In this circuit $Z_1$ and $Z_2$ are impedances looking to the left and right of the aperture. The microstrip line has an impedance of $Z_{0m}$ and the slot has an impedance of $Z_{0s}$. The patch impedance is given by:

$$Z_{patch} = Z_1 + Z_2 = 1/Y_1 + 1/Y_2$$  \hspace{1cm} (2.21)

where,

$$Y_1 = Y_0 \left( \frac{G_r + jB_{open}}{Y_0 + j(G_r + jB_{open}) \tan(k_0L_1)} \right) + jY_0 \tan(k_0L_1)$$  \hspace{1cm} L_1 = x_0 \hspace{1cm} (2.22)

$$Y_2 = Y_0 \left( \frac{G_r + jB_{open}}{Y_0 + j(G_r + jB_{open}) \tan(k_0L_2)} \right) + jY_0 \tan(k_0L_2)$$  \hspace{1cm} L_2 = L_{patch} - L_1 \hspace{1cm} (2.23)

In these equations $(G_r + jB_{open})$ is the edge admittance of the patch and $(Y_0, k_0)$ characterize the patch as a transmission line of width $W$. Since the slot is electrically small the aperture admittance $Y_{ap}$ is inductive. $Y_{ap}$ explains the energy stored near the slot and is given by:

$$Y_{ap} = -2jY_0 \cot(k_0L_s/2)$$  \hspace{1cm} (2.24)

The input impedance of the antenna at the center of the aperture is given by:

$$Z_{in} \left( -\frac{n^2}{n^2 + 1} \right) + Y_{patch} + Y_{ap} = -jZ_{0m} \cot(k_0L_{stub})$$  \hspace{1cm} (2.25)

and the condition for resonance is given by:

$$B_{patch} \approx \frac{4W^2}{Z_{0m}/k_0L_s}$$  \hspace{1cm} (2.26)

This shows that by increasing $L_s$ the resonant frequency decreases. It has been recommended to follow the following rules of thumb in choosing the slot width and length:

$$L_s > W_f + nd_f \hspace{0.5cm} \text{with} \hspace{0.5cm} n > 6 \hspace{0.5cm} ; \hspace{0.5cm} W_s/L_s = 1/10$$  \hspace{1cm} (2.27)
2.3.3. **Stacked Patch Microstrip Antennas**

Stacked microstrip patches can be used to achieve dual or multiple frequency operation. Also it has been shown that stacking microstrip patches result in an increased impedance bandwidth over a single layer radiator. In this configuration two microstrip patches with slightly different sizes are stacked on top of each other on a common ground plane. These two patches can be fed separately in a piggyback configuration, or together by the means of an aperture-coupled feeding, probe feeding, or microstrip line feeding. The upper patch is coupled to the excited lower patch and may radiate at a second resonant frequency. In this configuration each patch can use a thin substrate layer to maintain a low profile design and to minimize the excitation of surface waves. Also the designer can take advantage of several extra degrees of freedom such as element offset, size, and substrate thicknesses. The aperture-coupled configuration has an additional advantage; it isolates spurious feed radiation by the use of the common ground plane. An exploded view of a multilayered linear polarized aperture-coupled stacked microstrip antenna is shown in Figure 2.18:

![Figure 2.18. Aperture-Coupled Stacked Microstrip Antenna Exploded Geometry](image)

In order to analyze the coupling between elements, the electric current distribution on the patches or the field distribution between them can be studied. Two modes of operation are possible depending on the currents phase. The in-phase currents, known as the even mode, happen at a lower frequency and are suitable for radiation; while the out-of-phase currents, known as the odd mode, happen at a higher frequency and are not suitable for radiation. The electric and magnetic field distribution for the even mode of operation is shown in the following figure. The patches are coupled through the magnetic field.
2.3.4. Antenna design geometry

This section addresses the design procedure of a dual-band aperture-coupled stacked patch antenna. The parameters of the antenna have been taken from [POZ97]. Therefore, this antenna will not be adjusted to the required frequency bands of Galileo. Basically, the objective of this section is to learn how to simulate the aperture-coupling feeding in a stacked patch structure in Ansoft Designer 4.0. Moreover, the antenna is studied in deep in order to know the possibilities of this geometry.

In Figure 2.20 the antenna geometry is presented. The patch sizes are 8.35 cm square for the bottom patch and 7.90 cm for the top patch. The length of the slot is 5.3 cm and the width is 1 mm. The stub length is $\lambda/4$ at 1.4GHz, which means 4.94 cm. The dielectric material used in the design is the Rogers Duroid 5870 with a dielectric permittivity $\varepsilon_r = 2.33$. Furthermore,
there is a gap of air between layers in order to increase the thickness of the antenna with a minimum effective permittivity. To increase the bandwidth of the antenna is important to have a thick dielectric with low permittivity.

![Antenna perspective images](image)

Figure 2.21. Top (a) and bottom (b) perspectives of the antenna

### 2.3.5. Antenna design results

In this section the results of the simulations are going to be presented. We should have in mind that Ansoft Designer works supposing an infinite ground plane as well as infinite dielectrics layers. Therefore the field results are not precise. For example, the results do not have into account the field diffraction in the ground plane.

The Figure 2.22 presents the reflection coefficient of the antenna in two ways; the image (a) shows a rectangular graph with the results in dB whereas the image (b) shows a Smith Chart. We can see two resonances, one at about 1.36 GHz and the other at 1.62 GHz. It is important to point out that the impedance of this antenna is not matched, so the reflection coefficient is poor but enough to distinguish the resonances.

![Reflection coefficient graphs](image)

Figure 2.22. S11 depicted in a (a) Rectangular Graph and in a (b) Smith Chart

In the Smith Chart the resonances appear as two frequency loops. The higher frequency loop is due to the coupling of the lower patch and the slot, while the lower frequency loop is due to the coupling between the upper and lower patch.
The next picture shows the gain of the antenna. Both LHCP and RHCP gains do not take into account the impedance mismatch whereas the input gain represents the global gain having into account the impedance mismatch. From this picture can be pointed out how the frequency rejection between the two resonances is 3 dB as maximum. This level is too slow for our requirements. However, the level of cross-polarization is almost 20 dB or greater in both resonance bands.

![Antenna Gain](image)

**Figure 2.23. Antenna Gain**

In the figure below we can see how the axial ratio is under 2 dB in both bands. This level of axial ratio is with ideal phase shifts in the inputs. It will be degraded with the insertion of the 90°-hybrids plus the 180°-hybrid.

![Antenna Axial Ratio](image)

**Figure 2.24. Antenna axial ratio**
In Figure 2.25 and Figure 2.26 the high band and low band radiation patterns are presented respectively. Both are calculated without taking into account the impedance mismatch. They are very directive having a gain in boresight around 9 dB. Therefore the gain at 85° from boresight is rather low, less than -10 dB.

Using a material with a dielectric constant around 3.5 instead of air would make narrower the radiation patterns and will decrease the bandwidths. By doing this change together with adjusting the patch dimensions will be possible to approach our specifications.
2.4. A Dual Frequency Microstrip Patch Antenna for High-Precision GPS Applications

2.4.1. Introduction

In this section, a design of a dual-band circularly polarized antenna has been carried out. The antenna explained is a quadruple feed stacked shorted annular circular patch antenna operating at L1 and E5a-E5b bands of Galileo. In this geometry, to shorted circular annular patches are concentrically printed on two stacked substrates separated by an air gap. The main characteristic of this antenna is that the radiation pattern can be adjusted by varying the relation between the external and internal radii. Furthermore, the stacked shorted annular circular patch can be designed to avoid surface wave emissions reducing back and lateral radiations and improving efficiency. In the following point the coaxial feeding is explained in order to better understand the obtained results in the simulation.

The idea of using this structure is obtained from [BOC04]. In this paper, the circular polarization is generated using only one probe feeding with elliptical patches instead of circular ones. However, the axial ratio bandwidth is not enough to fulfil our objectives so we have used circular patches with four probes feeding the upper patch. Hence, the phase difference must be generated with two 90° Hybrids and one 180° Hybrid. The geometry of the antenna is explained in 2.4.3.

This second design is intended to learn how to use Ansoft HFSS 11 and to simulate antennas with coaxial feeding and via-holes. Moreover, we are interested to check in the simulations if a wide radiation pattern can be obtained with this geometry.

2.4.2. Coaxial feeding Feed/Probe Coupling

One of the basic mechanisms of transferring power to a microstrip patch is by using a probe passed through the substrate and soldered to the patch. This probe can be the inner conductor of the coaxial line or a strip line through a slot in the ground plane. The location of the slot should be at the point where the best impedance match is achieved. The coupling of the feed current $J_z$ to the $E_z$ field results in excitation in the patch. The coupling can be obtained by the following equation:

\[
\text{Coupling} = \iiint \nabla \cdot E_z J_z \, dv \equiv \cos(\pi x_0 / L) \tag{2.28}
\]

where $x_0$ is the slot offset from the edge of the patch. It can be seen that the coupling is maximum at the edge of the patch. The following figure shows the top and side views of an N-type coaxial connector:
The main advantage of the probe feeding technique is that the input impedance level can be easily adjusted and positioning of the feed is quite simple. There are limitations associated with this kind of feeding. For thick substrates a long probe is required, which causes additional spurious radiation, feed inductance and surface wave.

### 2.4.3. Antenna design geometry

The antenna geometry consists in a two concentrically stacked annular circular patches shorted to ground. This antenna was designed so that the first band located at L1 has an impedance matching bandwidth of 32 MHz (2%); and the second band located at E5a-E5b has an impedance matching bandwidth of 51 MHz (4.2%). It is also desirable to achieve a high rejection of the frequencies between those bands. This antenna is also circularly polarized (RHCP) and the axial ratio bandwidth (AR < 3dB) should be enough to cover all desired frequency bands. There is no step-by-step design procedure for this kind of structure due to the number of interacting design parameters.

There are two layers of dielectric separated by a thin layer of air. The antenna is fed using four coaxial probes. The phase in the probes are 0°, 90°, 180°, 270° respectively in order to produce right hand circular polarization. The antenna geometry presented in Figure 2.28.

![Figure 2.27. Coaxial probe fed microstrip patch antenna](image1)

![Figure 2.28. Perspective of the antenna geometry](image2)
In the figure above we can see the two patch radiators and the four feeding probes. It has to be pointed out how the antenna is shorted in the center. There is a hole in the center of the antenna. The walls of the hole connect the patches to the ground.

![Side view of the antenna](image)

Figure 2.29. Side view of the antenna

In Figure 2.29 the side view of the antenna is presented. In this picture is easy to identify the coaxial feedings and the layer structure. Both dielectrics have the same thickness $h = 3.2\,\text{mm}$. The thickness of the air gap is $t = 0.7\,\text{mm}$. The two dielectric layers have different permittivity. The permittivity of the top and bottom dielectrics are $\varepsilon_{r2} = 8$ and $\varepsilon_{r1} = 4.5$ respectively.

In the picture below the top view of the antenna is exposed. The dimensions of the patches and the positions of the probes are shown in this figure. The dielectric layers are square with a dimension of $d = 130\,\text{mm}$. The top patch is smaller than the bottom patch, therefore the top patch resonates at the high band and the bottom patch resonates at the low band. The top patch and bottom patch diameters are $a = 98\,\text{mm}$ and $b = 86\,\text{mm}$ respectively. The inner hole has a diameter of $c = 45\,\text{mm}$. The four probes are placed at $\rho = 34\,\text{mm}$ from the center with a phase shift of 90° between them.

The phase of the probes is shown in the figure. It can be pointed out how the phase is distributed from 0° to 270° in the four probes. With this distribution a right hand circular polarization is achieved.

![Top view of the antenna](image)

Figure 2.30. Top view of the antenna
2.4.4. **Antenna design results**

In this section the antenna simulation results are presented. This simulation has been carried out with Ansoft HFSS 11. The antenna has been adjusted in order to resonate at the frequencies of interest and to have an axial ratio below 3 dB in the bands of interest. Once these objectives have been fulfilled, the gain and the radiation patterns have been checked to know whether the values achieve our requirements. The antenna has still to be matched in order to improve the impedance bandwidth. In the pictures below the S11 and the antenna gain are presented.

![Figure 2.31. Reflection coefficient of one coaxial port](image)

In both figures the bands of interest are bounded by two green dotted lines. In the S11 graph we can check how the antenna resonates at the desired frequencies. We can also see how the antenna impedance needs to be matched in order to improve the resonance in the low band.
In the Figure 2.32 both resonances can be seen again in the gain plot. This gain is calculated at $\theta = 0$ and $\varphi = 0$. The realized gain takes into account the impedance mismatch; therefore the value of this gain is always lower than the normal gain. Is important to point out how the value of the gain RHCP and the directivity are the same at the bands of interest. Hence, the efficiency of the antenna would be close to 1 with a good impedance matching. Another interesting behaviour of this antenna is the good frequency rejection between both bands. Moreover the level of cross polarization is extremely low, indicating that there is a very good axial ratio along the whole band.

![Figure 2.33. Axial Ratio against Frequency](image)

In the image above we can see how the Axial Ratio is below 2.1 dB in both bands. This good behaviour of the antenna in the axial ratio is due to the quadruple feeding method. With this method the simulated phases are perfect along the entire frequency band. The axial ratio will get worse with the insertion of the hybrids since the phase and the transmission coefficients of the hybrid will not be ideal.

![Figure 2.34. High band radiation pattern](image)
In the Figure 2.34 and the Figure 2.35 the radiation patterns of the antenna at the frequencies of interest are presented. All the curves are calculated using the realized gain (taking into account the impedance mismatch). We have to point out the difference between both radiation patterns. The low band radiation pattern fulfils our objectives because the gain at 85° in elevation is almost -4.5 dBi. However, the high band radiation pattern is too narrow for our requirements.

It can also be seen the good symmetry of the radiation patterns at both frequency bands. This feature is possible thanks to using the symmetrical balanced feeding.

![Radiation Pattern: 1.189GHz](image)

*Figure 2.35. Low band radiation pattern*