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**ESCUELA TÉCNICA SUPERIOR DE INGENIEROS DE
TELECOMUNICACIÓN**

DOCTORAL THESIS - TESIS DOCTORAL

**PLANAR RECONFIGURABLE ANTENNAS FOR SATELLITE
COMMUNICATIONS**

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Ingeniero de Telecomunicación

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**DEPARTAMENTO DE SEÑALES, SISTEMAS Y
RADIOCOMUNICACIONES**

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TESIS DOCTORAL Planar Reconfigurable Antennas for Satellite
Communications.

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Abstract

This thesis has been carried out in Grupo de Radiación of Señales, Departamento de Señales, Sistemas y Radiocomunicaciones from the ETSI de Telecomunicación of Universidad Politécnica de Madrid. The title of the thesis is “Planar Reconfigurable Antennas for Satellite Communications”. It has been developed by José Manuel Inclán Alonso, Electrical Engineer Msc. under the supervision of Prof. Manuel Sierra Pérez.

Traditionally, most of satellite terrestrial systems use a parabolic reflector as an antenna. Nowadays there is a need of new systems with lower profile and to be used on the move. On the other hand, the satellite frequencies are increasing. The last band to be occupied by satellite communications is the Ka band (20-30 GHz) and the frequencies are expected to grow up in the future. For these reasons, there is a need of new antennas and technologies with lower profile and with electronic reconfigurability in higher microwave bands. Satellite Communications On the Move (SATCOM) also requires cheaper system to be used for the general public.

In this way, three novel technologies are studied in this thesis:

The first technology is the suspended shielded stripline technology. This technology has been studied in the past but in lower frequency bands. In this thesis an antenna at X band (8 GHz) is proposed using this technology. A new phase shifter is also proposed to steer electronically this antenna. The aim of this phase shifter is to have a cheap device with a small size.

The second studied technology is the Substrate Integrated Waveguide (SIW). This technology consist of imitating the waveguide technology in a printed board. This way, the losses are lower than other printed technologies and the manufacturing process is much easier than traditional waveguide. In this thesis two antennas are designed using this technology. The first one is an antenna at ku band (12 GHz) with a low bandwidth. The second one has been designed at the up ka band (30 GHz) and has larger bandwidth (2 GHz).

The third technology studied in this thesis is the gap waveguide technology. This technology has even lower losses than the SIW technology although the size of the components designed in this technology is bigger. It is very suitable for high frequencies (even larger than the ka band). In this thesis, a small antenna at the up ka band using this technology is done. The antenna combines the traditional gap waveguide design with the printed board manufacturing process, obtaining a good performance with a lower price.

Finally, this thesis also shows the design of a traditional antenna but for a novel application. The traditional antenna is a phased array with a shaped beam in one of the planes. The novel application is the long range communications with an Unmanned Aerial Vehicle (UAV).

Resumen

Esta tesis ha sido realizada en el Grupo de Radiación del departamento Señales, Sistemas y Radiocomunicaciones de la ETSI de Telecomunicación de la Universidad Politécnica de Madrid. El título de la tesis es “Antenas Planas Reconfigurables para Comunicaciones por Satélite”. Ha sido desarrollada por José Manuel Inclán Alonso, Ingeniero de Telecomunicación Msc. , bajo la supervisión del profesor Manuel Sierra Pérez.

Tradicionalmente las antenas más usadas en sistemas de comunicaciones por satélite son reflectores parabólicos. Actualmente hay una necesidad de nuevos sistemas con menor perfil y para ser usados en movimiento. Por estas razones, hay una necesidad de nuevas antenas y tecnologías planas que permitan reconfiguración electrónica en frecuencias superiores a las actuales. Las comunicaciones por satélite en movimiento (SATCOM) también requieren sistemas más baratos para poder ser usados por el público en general. Por otro lado, las frecuencias usadas en satélites se están incrementando. La última banda en ser ocupada por comunicaciones por satélite es la banda ka (20-30 GHz) y se espera que las frecuencias usadas se incrementen aún más en el futuro.

Por consiguiente, tres nuevas tecnologías son estudiadas en esta tesis.

La primera tecnología es la línea triplaca suspendida apantallada. Esta tecnología ha sido estudiada en el pasado pero en bandas de frecuencia más bajas. En esta tesis se propone una antena en banda X (8 GHz) usando esta tecnología. También se propone un nuevo diseño de desfasador para apuntar electrónicamente la antena. El objetivo de este desfasador es tener un dispositivo barato de tamaño pequeño.

La segunda tecnología estudiada en la guía integrada en sustrato (SIW). Esta tecnología consiste en imitar la tecnología de guíaonda en un circuito impreso. De esta forma, las pérdidas son menores que en otras tecnologías impresas y la fabricación es más sencilla que la guíaonda tradicional. En esta tesis se diseñan dos antenas usando esta tecnología. La primera es una antena en la banda ku (12 GHz) con un ancho de banda bajo. La segunda se ha diseñado en la banda ka superior (30 GHz) y tiene un mayor ancho de banda (2 GHz).

La tercera tecnología estudiada en esta tesis es la guíaonda con gap. Esta tecnología tiene incluso menores pérdidas que la tecnología SIW aunque el tamaño de los componentes diseñados en esta tecnología es mayor. Es una tecnología muy apropiada para bandas de frecuencia altas (incluso mayores a la banda ka). En esta tesis se desarrolla una pequeña antena usando esta tecnología. La antena combina el diseño tradicional con

guiaonda con gap con la tecnología de fabricación de circuitos impresos, obteniendo un buen comportamiento con un precio menor.

Finalmente, esta tesis también muestra el diseño de una antena tradicional pero para una nueva aplicación. La antena tradicional es un phased array con un diagrama conformado en uno de los planos. La aplicación novedosa es la comunicación a larga distancia con un vehículo aéreo no tripulado (UAV).

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Dedicated to...

Chapter 1

Introduction and state of art

1.1 Motivation

Traditional antennas for satellite communications such as parabolic reflectors (see [1]) have been studied for decades. These antennas are widely used in satellite user terminals. The problem comes with the mobility. If the user is not static the antenna has to track the satellite. Parabolic antennas need a mechanical structure to move physically the antenna so it can track the satellite, making the system bulkier. This and other problems can be solved using active array antennas (see subsection 1.1.1).

Active array antennas have been used for military purposes for decades. Although active antenna systems have a great behavior in terms of adaptability they are not affordable for the general purpose. In this way, there is a need of developing new technologies that make affordable the use of array antennas for the general purpose. These technologies should be planar and take advantage of the usual printed circuits technologies. In lower bands (lower than 5 GHz), patch antennas and traditional microwave circuits are good candidates but as the frequency goes up, the losses in these technologies are too high. The solution proposed in this thesis is to use printed technologies that reduce the conductor losses or imitate the behavior of regular waveguides. Of course there is always a trade-off between losses and manufacturing media. The thesis is mainly focused on the design of the array antenna itself without the active circuits. However the designs are done to easily include the active circuits necessary for the reconfigurability.

1.1.1 Array antennas advantages

An array antenna is a set of small antennas that work together to improve the antenna radiation characteristics. The principal advantages are:

- **More directivity.** An array antenna has a larger area than a single element, therefore, it can focus more power in one single direction if all the elements are fed with the appropriate phase as show in Figure 1.1. Thus, the directivity is directly proportional to the area that all the elements occupy. However, there are also certain factors that reduce the antenna gain when many elements are used in an array antenna such as losses in the distribution network, aperture efficiency, etc.

FIGURE 1.1: Radiation pattern of one single element (a) and a group of elements (b).

- **More radiated power:** a set of N antennas can radiate N times the maximum power that supports one single element.
- **More distributed amplification.** Instead of one single high gain amplifier, the antenna can have a distributed amplification if a group of elements share an amplifier. This makes easier the design of the amplifier, the design of the cooling system and the antenna durability since the antenna can still work if several amplifiers are broken.
- **It allows an electronic steering.** As it is explained in Section 1.2.1, the antenna can be steered changing the phase of every element. The electronic steering has several advantage compared to traditional mechanical steering: lower profile, more accurate pointing, quicker switching beam time and multi beam capabilities. Examples of arrays with electronic steering can be seen in [2, 3]. Figure 1.2 shows the cell of the SARAS-GEODA project. SARAS-GEODA project is an example of an electronic steering array for satellite communications at S band, designed in the Radiation Group of Technical University of Madrid. The figure shows three possible pointing directions that can be selected electronically changing the phase

in every element. More references about the SARAS-GEODA project can be found in [4, 5].

FIGURE 1.2: Steered beams of the GEODA cell.

- The diagram can be dynamically conformed. Changing the phase and amplitude feeding of each element, the radiation pattern can be shaped. The possibility and variety of diagrams shapes are usually larger than traditional antennas such as parabolic dishes. One example is the cosecant squared pattern that is used in radar antennas to overcome the propagation losses with the distance (examples can be found in [6, 7]). Another example is the side lobe level reduction as shown in [8].
- The antenna can have several simultaneous beams. With the appropriate feedings the antenna can point to several directions with several beams. Moreover, depending on the circumstances, the beams can be grouped or modified to have a more directive beam or a specific pattern. Examples of multi beam antennas can be found in [9–11].
- Improvements in the gain over noise temperature ratio compared to traditional antennas. The noise in every element of the array is not coherent, thus, the noise is sum out of phase whereas the signal in every element is summed in phase. If every element has its own Low Noise Amplifier, the noise generated by every amplifier is also summed out of phase. Therefore, the global gain over noise temperature can be better than a system with only one antenna plus a single Low Noise Amplifier.
- Possibility of blocking interferences. If the signal in every element is digitalized (Adaptive array), the nulls of the radiation pattern can be moved to block certain interferences as shown in Figure 1.3. This process is done digitally, thus, the diagram can be conformed in all the possible ways. This is especially important in the electronic warfare when the enemy tries to blind the system through a high power signal. Examples of adaptive algorithms and antennas are shown in [12–14].

FIGURE 1.3: Example of adaptive antenna.

1.2 Array analysis

The total radiated power of an array antenna can be obtained summing the radiated power of each of the elements. The radiated power of each element depends on its geometry and feeding signal. In a first approximation, it can be supposed that the radiating field of each element is the same except for the feeding signal. Therefore, the array radiation pattern can be decomposed into two terms: the radiated field of one single element and the array factor that takes into account the feeding of each element as shown in Equation 1.1.

$$\vec{E}_{total} = \sum_i \vec{E}_{element} = \vec{E}_{isolated\ element} \cdot \sum A_i \cdot e^{j k \hat{r} \cdot r_i} \quad (1.1)$$

The information about the polarization yields in the first part of the equation since it is the only one with vectorial behavior. The second part of the equation represents the complex feeding of each element (A_i). Modifying these feedings, the radiation pattern can be conformed. Finally, the last exponential term takes into account the phase difference in the propagation caused by the displacement of each element. The summation part is usually called array factor, therefore, the radiation pattern can be decomposed into two terms: the radiation pattern of a single element plus the array factor.

In the next subsections, the most general conformations of the diagram pattern are explained. These conformations are based on changing only the phase among elements, changing only the amplitude among elements or changing both amplitude and phase among elements.

1.2.1 Phase only variation

The array factor of a lineal array with the same amplitude feeding but a different phase feeding is sketched in Equation 1.2. The distance among elements is equal and is represented with the letter d .

$$AF = \sum_i \exp(j \cdot (k_0 \cdot d \cdot i \cdot \cos(\theta) + \alpha_i)) \quad (1.2)$$

The maximum of the array factor occurs when all the contribution are summed in phase. This happens when the exponential of the equation is equal to one (the exponential phase equals to zero). This leads to Equation 1.3. The equation establishes that the antenna can be steered to a certain angle (θ_0) if there is a particular phase shift among elements.

$$\alpha_i = -k_0 \cdot d \cdot \cos(\theta_0) \cdot i \quad (1.3)$$

Figure 1.4 shows three array factor for three different phase shift among elements. As it is shown, when there is not phase shift among elements the antenna points to broadside (90°) and when the phase shift is changed the antenna starts to point to higher elevations.

FIGURE 1.4: Array factor changing the phase shift among elements.

Therefore, a phase shift among elements can be used to electronically steer the antenna. The antennas that use this principle are usually called phased arrays. Many references

of modern phase arrays can be found in the literature as for example in [3, 15–17]. Chapters 3 and 4 are based on phased array topics.

1.2.2 Amplitude only variation

A different amplitude in the feeding could be useful to reduce the side lobe levels. Usually, the power that receives some element in the array is larger when the element is closer to the array center. That is why this amplitude distribution is usually called tapering. The effect is similar to the windowing in digital signal processing.

Thus, since the array factor is equivalent to a Fourier transform of the feedings, the side lobes will be lower if the amplitude variation is higher and vice versa. There are a lot of possible synthesis and studies on how this amplitude variation can be. The most relevant are: triangular, cosine over pedestal, Taylor, Dolph-Chebyshev and exponential distributions.

Figure 1.5 shows a comparison of the feedings and array factors of several distributions. As it is shown, the triangular distribution has lower side lobe levels than the cosine over pedestal distribution since the tapering is more aggressive. The figure also shows a binomial distribution which is a distribution without side lobes. One effect that can be observed is that the beam-width is usually increased when the side lobe levels are reduced, thus, there is a trade-off between the side lobe levels and the radiation efficiency that can be achieved with an array antenna. Another concern is the realizability of the distribution. In the array factor calculations are not taken into account some effects like the coupling among elements, the maximum ratio between power dividers, the reflections in the distribution network... For these reasons, tapering with high decays are more theoretical than practical.

The same side lobe reduction can be achieved with only phase synthesis or combining both, amplitude and phase (see [18]).

Side lobe level reduction based on amplitude tapering is used in Chapters 2, 5 and 6.

1.2.3 Phase and amplitude variation

The array factor can also be seen as a Z-transform. Equation 1.4 shows the equivalence for a linear array. Thus, the diagram can be expressed by its roots (the last part of the equation). The array factor will take its values from the radius one circumference ($|z| = 1$). The visible margin depends on the distance among elements and it can be a portion of the modulus one circumference or more than one spin.

FIGURE 1.5: Amplitude feedings (a) and array factors (b) of several weighting distributions.

$$AF = \sum_{i=0}^{N-1} A_i \cdot e^{jk_0 \cdot d \cdot i \cdot \cos(\theta)} = \sum_{i=0}^{N-1} A_i \cdot z^i = A_{N-1} \prod_{i=1}^{N-1} (z - z_i) \quad (1.4)$$

Therefore, the array synthesis can be done moving the zeros of the Z-diagram as it is done in filter theory. Figure 1.6 shows a simple example on how this process is performed in a linear array of 32 elements. In the left diagram the first three zeros have been moved anticlockwise a few degrees. In the second diagram 11 zeros have been taken out of the modulus one circumference.

FIGURE 1.6: Z-diagram for two cases.

The array factors of the Z-diagrams of Figure 1.6 are shown in Figure 1.7. The red array factor corresponds to the left Z-diagram. As shown, part of the side lobes in the right side are lower than the lobes in the left side because in this particular area the zeros are closer. The blue array factor corresponds to the right Z-diagram. In this case, the first zeros have been taken out and as a result there are not nulls in the array factor from 90° to 140° .

FIGURE 1.7: Array factor of the Z-diagrams of Figure 1.6.

Chapter 4 uses the synthesis using Z-diagram to synthesis a shaped beam.

There are a lot of different methods to synthesize an specific array factor, the most famous are: Fourier, Shelkunov and Wooden-Larsson methods. Many of them can be found in [8].

1.3 Low profile technologies for millimeter frequencies

As it has been established in 1.1, there is a need of new technologies that make affordable the use of active arrays in higher microwave bands. In this way, there are three candidates that set a trade-off between losses and easy manufacturing process. The first candidate is the shielded suspended stripline. This technology has been used in the past in lower frequencies as it is shown in [19]. The antenna in Chapter 2 uses this technology in the X band. The second and third candidates are the substrate integrated waveguide (Subsection 1.3.1) and the gap waveguide (Subsection 1.3.2) . These technologies are more novel and more suitable for higher bands.

1.3.1 Substrate integrated waveguide technology

Nowadays, one of the most interesting technologies for low profile antennas is the Substrate Integrated Waveguide (SIW). This technology consists in imitating a waveguide in a dielectric substrate. The waveguide walls are made using plated via holes. The technology has three advantages: it has lower losses than traditional printed technologies such as microstrip or stripline, the fabrication is not very complex and it allows to integrate monolithic devices very easily.

The classical SIW is presented in Figure 1.8. Several modifications can be done as seen in [20]. All the usual modifications to the standard SIW want to reduce the losses, maintaining the easy manufacturing process. In this way, to reduce the leakage the plated holes are substituted, for example, by plated slots and to reduce the dielectric losses multilayer structures are used, leaving the inner part empty (the substrate dielectric is substituted by air).

FIGURE 1.8: Substrate integrated waveguide with its main parameters.

Initial studies of this technology can be found in [21–23]. These studies establish design equations between SIW and a regular waveguide. The main concern is how to extrapolate a design in regular waveguide technology into SIW technology since the simulation of regular technology is much faster.

As far as the losses are concerned there are three losses types: the conductor losses, the dielectric losses and the radiation losses. The conductor losses are similar to the losses of a regular waveguide. The dielectric losses are similar to other printed technologies. Finally, the radiation losses are related to the leakage through the via-walls. These losses mechanisms are studied in [24, 25].

Most of the traditional passive microwave components have been designed in SIW technology, e.g., cavity filters [26–28], power dividers [29–31], couplers [32–34], six port junctions [35] and beam steering networks [36, 37]

The most common antennas used in SIW are cavity based antennas or slotted waveguide antennas. Examples of novel designs are shown in [38–41].

Active elements can easily be included in SIW technology. These elements are usually placed in a small microstrip network with two small SIW-microstrip transitions. Examples of the use of active circuits within a SIW are shown in [42, 43].

More details on how to simulate this technology besides several examples can be found in Chapters 5 and 6.

1.3.2 Gap waveguide technology

The dielectric and leakage losses in the Substrate Integrated Waveguide are too large in higher bands (above ka band). Although these losses can also be reduced in the SIW technology, it increases the structure complexity, making it more difficult to achieve good results in higher bands due to the manufacturing tolerances.

Gap waveguide is based on metamaterial structures. A simple scheme can be seen in Figure 1.9. The wave is propagated through two plates. The upper plate is a regular metal plate whereas the lower plate has a metamaterial surface that allows the propagation in the longitudinal direction and forbids the propagation in the transversal direction. Since the wave is propagated in the gap between the two plates, it mainly sees only air as a dielectric, thus, the dielectric losses are very low. Since the propagation in the transversal direction is forbidden, the leakage is also very low compared to SIW. Another advantage compared to other technologies is that a union between the two plates is not necessary. In the case of a regular rectangular waveguide, having a good metal union between the two halves is crucial, and in high bands this union could be very difficult to achieve.

Examples of the use of this technology in antenna design are shown in [44, 45].

Chapter 7 explains in more detail this technology and shows the design and results of a small ridge gap waveguide antenna in the ka band.

1.4 Methodology followed in the thesis

This thesis is focused on real prototyping of antennas. Thus, all the designs proposed are manufactured and tested. For this reason these designs are not very ambitious in terms of size and capabilities since they have to be manufactured, in many cases using only the university laboratories.

FIGURE 1.9: Scheme of a Gap Waveguide.

All the chapters share the same structure:

- The chapters begin with a small theoretical introduction, then, the antenna structure is explained and in some cases a first mathematical design is performed.
- Once the structure is sketched and the initial mathematical design is done, full wave simulations or optimizations of every part are done using CST Microwave Studio. The solver used in most of the designs is the Finite Integral Technique.
- The measurements versus simulations of a first prototype are presented. These results are analyzed and several tips on how to solve several manufacturing problems are shown.
- Finally, every chapter includes a small conclusion section in which the future the detected problems and future lines are sketched.

Chapter 2

Low profile man-pack antenna

2.1 Introduction

Nowadays low profile passive array antennas are being more and more used, substituting traditional parabolic antennas in satellite communications. This chapter describes the design of a modular low profile antenna for personal communications. These antennas are usually called man-packs because they are carried in a case with the necessary receptors and transmitters. The antenna must be deployed and steered to the desired satellite (Figure 2.1). The chosen frequency band is the X band. Man pack antennas are especially suitable for military applications because they are light weight, robust and easy to transport.

FIGURE 2.1: Man-pack antenna for satellite communications.

One of the main problems of low profile structures is that the losses are too high. This is especially a problem in large passive array antennas. In transmission the problem could be solved with more amplification from the transmitter but in reception the losses will

be represented as a big attenuator at the beginning of the chain increasing greatly the noise factor.

One solution to meet with the required efficiency is to use a radial line slot antenna [46]. The problem is that this kind of antenna only radiates with one circular polarization and is not versatile for future improvements. Another solution is to reduce the power distribution network length as much as possible using a serial distribution network [47]. The problem here is that the bandwidth is reduced and it is not possible to cover the whole X band.

The solution chosen in this chapter is to divide the array antenna in subarrays and use a low loss power distribution network to distribute the signal among these subarrays. Thus, the subarrays could be done in regular printed technology, allowing to have phase shifters, hybrid circuits and other devices on them. On the other hand, the power distribution network must be done in a low loss technology. The antenna must work in both circular polarizations simultaneously. This means that the antenna must receive and transmit in both circular polarizations. The main antenna requirements are summarized in table 2.1.

ANTENNA SPECIFICATIONS		
<i>Parameter</i>	<i>Specification</i>	<i>Unit</i>
Working bands		
Tx:	7.9 to 8.4	GHz
Rx:	7.25 to 7.75	GHz
Polarization	Dual circular polarization for Tx and Rx band	-
3 dB beamwidth	5	degrees
Maximum gain	26	dBi
Radiation efficiency	50-60	%
Axial Ratio	<2	dB
Isolation between polarizations	>20	dB
Maximum Side Lobe Level	-13	dB
Size	400x400x25	mm
Weight	<3.6	kg

TABLE 2.1: Main specifications of the man-pack antenna.

Three antennas prototypes have been designed:

- The first prototype is a single polarization antenna of 16x16 radiating elements grouped in 4x4 subarrays. This prototype served to prove the concept. The substrates used in this prototype are based on PTFE. One of the problem detected

is that there is not enough space in the subarray to include the other network for the second polarization.

- The second prototype is a dual polarized antenna of 12x12 radiating elements grouped in 3x3 subarrays. Since the antenna size is maintained, the subarray size has been increased. This way is possible to include the second network in the subarray.
- In the third prototype the cost of the antenna is reduced. To reduce the antenna manufacturing cost, the substrate is changed from PTFE to RO4350B. A new ad-hoc connector and a new multilayer stack-up are also used.

The first and second prototype are built and measured. The third prototype has been designed but not built. The third prototype subarray and a power distribution network demonstrator (with ad-hoc connectors) have been built and measured. The final antenna has not been manufactured because of a lack of funds.

The structure and the analysis of the array are shown in Section 2.2. Section 2.3 explains the design of every element in the subarrays. It also shows the results of the subarrays designed for each prototype. Section 2.4 shows the design of the power distribution network and the ad-hoc connectors. The results of each prototype are also given. The first two prototypes are presented in Section 2.5. Finally future work and improvements are depicted in Section 2.6.

2.2 Antenna structure and analysis

The first part of this section describes the antenna structure, detailing the materials used in every part. It is not possible to meet the specifications regarding the Side Lobe Levels (SLL) if all the elements are uniformly fed in the array. Thus, a tapering in the feed amplitude is used among subarrays, meaning that the central subarrays receive more power than the subarrays in the corners. The effect of this tapering is analyzed in Subsection 2.2.2.

2.2.1 Antenna structure

As it has been commented in Section 2.1, the antenna is divided in two main parts: the power distribution network and the subarrays. One of the advantages of dividing this way the antenna is that every part can be designed independently.

Figure 2.2 shows the antenna structure scheme. The subarray contains the radiating elements and a small microstrip network to distribute the power among its elements. The radiating elements are double stacked patches. The active patch is fed through a plated via from the microstrip network. The parasitic patch is printed in a thin substrate and is separated from the active patch by a foam sheet. To connect the subarrays and the power distribution network, pluggable connectors are used. The power distribution network is done in suspended stripline technology. This means that the line is printed in a thin substrate and suspended by metallic walls at a distance from the ground.

FIGURE 2.2: Structure scheme of the man-pack antenna.

The second prototype is shown in Figure 2.3. The central subarray would be plugged into the antenna just pressing carefully on the center.

FIGURE 2.3: Second prototype.

2.2.2 Antenna analysis

The analytical radiation patterns are shown in this subsection. All the analysis are done at the central frequency and for a $\phi = 0^\circ$ that corresponds to the maximum level of side lobes.

There are two classes of arrays to analyze. The first one is an array of 16×16 elements grouped in 4×4 subarrays with a distance among elements of 25 mm ($0.65 \cdot \lambda_0$). The

second one is an array of 12x12 elements grouped in 3x3 subarrays with a distance among elements of 33.3 mm ($0.87 \cdot \lambda_0$).

The first concern is the presence of grating lobes in the array with larger distance among elements. The array factor of 12x12 elements is shown in Figure 2.4. To simulate the radiation pattern of one single element a $\cos(\theta)$ function is used. In the Figure 2.4 is plotted the array factor, the single element radiation pattern and the result of the final 12x12 array. Grating lobes appear at $\theta = 78^\circ$ but they are not a huge problem since the radiating element does not radiate much power in this direction.

FIGURE 2.4: Array factor for the 12x12 element array.

As it has been commented before, to reduce the side lobe levels a tapering in the feeding of each subarray is performed. Figure 2.5 shows the normalized radiation pattern for the array of 16x16 elements, changing the feeding weights.

In Figure 2.5 is also included the gauge envelope of ITUR S-465-5 [48] for portable systems for satellite communications. The gauge envelope can be calculated as:

$$G(\theta) < \begin{cases} 55 - 10\log(D/\lambda) - 25\log(\phi) & \text{for } 100\lambda/D^\circ < \phi < 48^\circ \\ 13 - 10\log(D/\lambda) & \text{for } 48^\circ < \phi < 90^\circ \end{cases} \quad (2.1)$$

To estimate the antenna diameter (D), the average between the diagonal and the side has been taken.

The first tapering (A=1, B=0.8 and C=0.6) has been selected as a trade-off between a low first SLL and an easy design of the distribution network. The radiation pattern fulfills practically with the ITUR recommendation except for two far lobes at $\theta = 45^\circ$ and $\theta = 55^\circ$. In any case, these two far lobes are below 25 dB with respect to the main lobe. The directivity is reduced less than 0.15 dB with the selected tapering as it is presented in 2.5 (b).

FIGURE 2.5: Radiation pattern changing feeding weights.

A comparison between the 12x12 element antenna and the 16x16 element antenna is shown in Figure 2.6. Both antennas have the same tapering ($A=1, B=0.8$ and $C=0.6$). The behavior of the antenna with 16x16 elements is much better regarding the compliance with the ITUR recommendation. Therefore, the only reason for using a 3x3 subarray distribution is because there is more space among elements and the design process is easier.

FIGURE 2.6: 12x12 and 16x16 antenna with tapering comparison.

2.3 Subarray system

2.3.1 Subarray structure

The subarrays for the two first antenna prototypes share the same structure and stack-up. The subarray for the third prototype is made in a different low cost material and with a different stack-up.

There are two main parts in the subarrays: the radiating elements and the microstrip network.

The selected radiating element is a double stacked patch. The double stacked patch covers a wide bandwidth with a low profile. The active patch is fed by two plated vias separated by a 90° angle. These vias come from the microstrip network. The parasitic patch is printed in a thin substrate and separated from the active patch by a foam sheet.

The microstrip network contains power dividers, hybrid circuits and the SMP connectors. There are several kinds of power dividers used in the subarrays: regular T-dividers, multi-section T-dividers and Wilkinson dividers. The multi-section T-dividers have a better matching than regular T-dividers and the Wilkinson dividers are matched not only in the main arm but also in the two other ports. In first subarray prototype only regular T-dividers are used. In the second prototype Wilkinson dividers are used to improve the matching of the subarray since they help to cancel the coupling among elements. In the last prototype multi-section power dividers are used to improve the matching. Wilkinson power dividers cannot be used in the third prototype since the network for one of the polarizations is buried and resistances cannot be soldered.

To achieve a circular polarization in circular patches two orthogonal modes must be excited with a quadrature signal. The quadrature signal can be obtained with hybrid circuits and more specific with a branch line circuit. Moreover, the branch line circuit allows to work with the two circular polarizations simultaneously. One of the problems of traditional hybrid circuit is that they do not have a wide bandwidth. To increase the bandwidth, three branches hybrid circuit can be used. Another problem of the branch line hybrid circuits is that they take up a lot of space. The solution is to miniaturize these circuits, degrading the behavior but reducing the size.

In first and second prototype sequential rotation technique among every 2×2 elements is used to improve the axial ratio. This technique is detailed in Section 2.3.4. In third prototype, the sequential rotation technique is performed among subarrays instead of radiating elements to reduce the loops in the microstrip network.

Figure 2.7 shows the subarray layer scheme of the third prototype. The microstrip networks and the active patches are in a single stacked structure. The foam and the second patches are above this structure and they are pressed by nylon screws.

FIGURE 2.7: Subarray layer scheme.

2.3.2 Radiating element

As it has been commented before the selected radiating element is a double stacked patch. The first step is to simulate the patch antenna with open boundaries and optimize the matching and coupling in the desired bandwidth. The path is printed in a dielectric substrate, thus the height of the first patch is fixed by the available substrate heights. The height of the foam that separates the first patch and the second patch is also fixed to some values. Therefore the variables to be optimized are: the patch radii, the offset of the feeding for the active patch, the aperture in the ground plane for the plated hole and the pad in the microstrip. The plated hole radius can be also optimized, but the optimization gives values under the minimum radius for an easy fabrication. The optimization variables are depicted in Figure 2.8.

The simulation results for an isolated patch are shown in Figure 2.9. The matching and the coupling between probes are lower than -19 dB.

One of the problems of array patch antennas is the coupling among elements caused mainly by surface waves in the dielectric substrate. To take into account these couplings, the patch is optimized using periodic boundaries in a FIT simulator. The results of the optimized path are shown in Figure 2.10 (a). The matching for a linear polarization is

FIGURE 2.8: Patch Scheme.

FIGURE 2.9: S-Parameters of a isolated patch.

below -20 dB and the coupling between probes is lower than -15 dB. To simulate the real behavior of the patch, a circular polarization must be excited. This is easily done adding in the simulation ports a phase shift of $\pm 90^\circ$. The results for the circular polarization are shown in Figure 2.10 (b). Since the two ports are excited simultaneously, the receive signal in every port is formed by the reflected signal from the patch and the coupling from the other port. The output signals in every port are different because input signals do not have the same phase. As a curiosity, in lower frequencies the received signal in one port is above 0 dB. That is because the receive signal is composed by the reflected signal plus the coupling from the other port, and this combination is greater than the input signal in the port.

In a branch line hybrid circuit if the reflected signals in the output ports are similar, they go to the isolated port. This means that once the hybrid circuit is added to the patch, the reflected signal of the patch goes to the other polarization (isolated port of the hybrid circuit) and the coupling between patch probes goes to the input port. In other words, when the hybrid circuit is added to the patch, the coupling and matching of the patch

FIGURE 2.10: S-Parameters of a patch with periodic boundaries.

are interchanged with the matching and the coupling between circular polarizations of the structure. For this reason, there is not any point in optimizing only the matching if the coupling between lineal polarizations is not good enough. Figure 2.11 shows the optimization of a patch, optimizing the coupling between polarizations in the patch.

FIGURE 2.11: S-Parameters of a patch with periodic boundaries optimized in coupling.

The main difference between optimizing only matching and optimizing coupling is the achieved axial ratio. Figure 2.12 shows the axial ratio for broadside direction for the previous patches. The axial ratio for the patch optimized in coupling between probes is clearly better.

To check the behavior of the radiating element, two patches have been manufactured. One of them is optimized in matching and the other one in coupling. To get a circular

FIGURE 2.12: Axial ratios for broadside direction of patches optimized in matching and coupling.

polarization, the hybrid circuit is added. Therefore, the measurements show the coupling between circular polarizations and the matching of each polarization. Figure 2.13 shows that the results are better for the patch optimized in coupling (b). The patch optimized in matching (a) has a good isolation between polarizations but a worse matching compared to the other one.

FIGURE 2.13: Measured S-Parameters of a patch with hybrid circuit.

2.3.3 Microstrip elements

2.3.3.1 Dividers

To distribute the power among elements in the subarray, either T-dividers or Wilkinson dividers are used. The main differences between a regular T-divider and a Wilkinson divider are that in Wilkinson divider all the ports are matched and output ports are

isolated. Since it is a three port network, to match all the ports a resistance must be included in Wilkinson dividers. This makes it impossible to place Wilkinson dividers in internal layers since nothing can be soldered in internal layers.

Figure 2.14 (a) shows the model of a two section T-divider. The simulated matching is shown in Figure 2.14 (b). One section and two section T-divider are compared. As it is logical the two section T-divider has a larger bandwidth. The disadvantage is that it takes up more space.

FIGURE 2.14: Simulation of microstrip T-dividers.

The simulation of a Wilkinson divider is shown in Figure 2.15. The matching in all ports is better than -20 dB and the coupling between port 2 and 3 is lower than 20 dB.

FIGURE 2.15: Simulation of microstrip Wilkinson divider.

2.3.3.2 Transitions between layers

In the third subarray prototype, there are two microstrip networks, one for each polarization. The first network is similar to a standard microstrip and the second network is a microstrip located in an internal layer (see Figure 2.16). Thus, a transition between the first and the second network must be designed. There are two ways to design this transition. The first way is to couple the signal through a slot in the ground plane and the second way is to couple the signal through a plated hole. The advantage of the slot is that there is not a plated hole, the disadvantage is that the transition takes up more space.

FIGURE 2.16: Transition between layers in microstrip networks.

The slot length in Figure 2.17 (a) must be close to $\lambda_g/2$ to be at resonance. Since the line ends in an open circuit, to have a short circuit between the two lines, the distance from the slot till the end of the line is $\lambda_g/4$. The size of the transition can be reduced as shown in Figure 2.17 (b). The horizontal slot is changed by a H-slot, the currents in the H slot must also cover $\lambda_g/2$, hence, the H-slot outline must be $\lambda_g/2$. The extra line to achieve the short-circuit can be also bended.

FIGURE 2.17: Transition through slots.

The transition through a plated hole is shown in Figure 2.18. There are two main advantages over the slot transition. The first one is the size, now the transition area is the area of the plated hole plus the necessary pad. The second advantage is that the angle between the first network line and the second network line does not have to be 0° . Figure 2.18 shows three examples with angles of 0° , 90° and 180° . The variables to match the transition are the aperture in the shared ground and the plated hole pads.

FIGURE 2.18: Transition through plated hole.

The simulation results are presented in Figure 2.19. In the three cases the matching is below -24 dB at the X band although the bandwidth of the miniaturized slot and the plated hole transition are greater. As far as the losses are concerned they are also lower for the miniaturized slot and the plated hole transition.

FIGURE 2.19: Simulated S-parameters of transition between layers in microstrip.

Since the best simulation results are the ones for the transition with a plated hole, a through with two transitions is manufactured (see Figure 2.20). The measured matching is below -13 dB in the X band, taking into account that the matching of the connectors at X band is around -15 dB, it is not a bad result. The losses are around 1 dB. Since the losses in the connectors are around 0.3 dB for each one, the losses in the two transitions are around 0.4 dB, similar to the simulated results.

2.3.3.3 Hybrid circuits

To get a circular polarization in the patch, the two probes must be fed with a quadrature signal, meaning that the phase difference between outputs must be $\pm 90^\circ$. The hybrid

FIGURE 2.20: Measurement of a transition between microstrip networks.

branch line circuit is the most suitable circuit to achieve this. With this circuit the patch can be fed with two signals, one for each polarization, with a high isolation between them. The classic branch line circuit is presented in Figure 2.21 (a). The impedance of the branches are $70\ \Omega$ and $50\ \Omega$. More details and equations for designing branch line hybrid circuits can be found in [49]. The simulation results of the classic branch line are presented in Figure 2.21 (b). The isolation and matching are very similar and below $-15\ \text{dB}$ at the X band. The outputs are not exactly equal but they have a difference lower than $0.25\ \text{dB}$ at the X band. The output phases are not presented but they are exactly 90° in the center frequency and a bit lower/higher in lower/higher frequencies. To enhance the bandwidth of the branch line circuit more branches can be added. In this chapter, a three branch hybrid circuit has also been studied and tested.

FIGURE 2.21: Branch line hybrid circuit.

Each branch has a length of approximately $\lambda_g/4$. The size of the branch line can be

reduced approximating the $\lambda_g/4$ line to a short line with an stub in the middle, the stub is terminated in an open circuit as shown in Figure 2.22. This substitution is taken from [50]. More examples of miniaturized hybrid circuits can be found in [51].

FIGURE 2.22: Equivalence of a $\lambda/4$ line (a) and a line with an open stub (b).

The first miniaturized hybrid circuit is shown in Figure 2.23 (a). The problem with this circuit is that the distance between the stubs is very small, making it difficult the manufacturing process. To solve this problem the stub ends can be rounded and re-optimized as shown in Figure 2.23 (b).

FIGURE 2.23: Miniaturized hybrid circuits.

The measurements for the miniaturized hybrid circuit with rounded stubs are shown in Figure 2.24. The matching has moved a bit to lower frequencies. The phase difference between outputs is close to 90° in the frequencies where is well matched. As it has been proved, the behavior of the miniaturized hybrid circuit is similar to the classic hybrid circuit. In this case, the length of the branch must be reduced a little bit to move the working band to higher frequencies.

A three branch miniaturized hybrid circuit has also been manufactured and measured (see Figure 2.26). The achieved bandwidth is greater than the previous prototype as it is logical.

Figure 2.26 shows a comparison among the designed branch line circuits. The miniaturization pays off especially for the two branches hybrid circuit.

FIGURE 2.24: Measurements of the first miniaturized hybrid circuit.

FIGURE 2.25: Measurements of the three branches miniaturized hybrid circuit.

FIGURE 2.26: Size comparison among BLCs.

2.3.4 First and second subarray

The first and second subarrays are printed in a substrate based on PTFE. The advantage of PTFE is that it has low dielectric losses and a low permittivity. The disadvantage is that the plated holes are very difficult to make, especially the blind plated holes. The low

permittivity is good for the radiation efficiency but bad because the microstrip circuit size is bigger.

The microstrip network for the first subarray is shown in Figure 2.27 (a). The network is only for one polarization, thus only one of the inputs of the hybrid circuit is used, the other one is loaded with a $50\ \Omega$ resistance. The sequential rotation technique is used among 2×2 elements. This technique consists in rotating each element some angle and compensate this rotation with a phase shift in the network. This way the axial ratio is improved since the errors are summed out of phase and the copolar components are summed in phase.

One of the problems in the first prototype is that it is impossible to make room for the second network. For this reason, the size of the second prototype has been increased. Since the antenna size must not change, the number of subarrays in the second antenna prototype is 3×3 . Hence the second prototype has the two circular polarizations as shown in Figure 2.27 (b).

FIGURE 2.27: First (a) and second (b) subarray prototype.

The measured results for the first and second subarrays are shown in Figure 2.28. The matching is below $-11\ \text{dB}$ for both polarizations and prototypes. The isolation between polarizations in second prototype is better than $18\ \text{dB}$. The isolation between polarizations is much better than the matching since the patch is optimized in matching and not in coupling. The axial ratio is below $1\ \text{dB}$ in the whole X band for the use of sequential rotation. One of the polarizations in the second prototype has a lower gain because its

microstrip network is longer and it has more losses. The gain in the second prototype is greater because the area is larger (although it has the same number of elements).

FIGURE 2.28: Measured axial ratio (a), Copolar and Crosspolar gain (b) and S-Parameters (c) of the first and second subarrays.

2.3.5 Third subarray

To reduce the price of the antenna the PTFE substrate has been changed to RO4350B. This substrate has more dielectric losses but the plated holes are much easier to do. Figure 2.29 shows the stack-up of the circuit with the microstrip networks and the active patches. Each polarization goes in a different microstrip network. Since they are only needed four layers (patches, 2nd network, ground and 1st network) there is not point in using more than 2 substrates. For this reason three prepegs are used as a dielectric between the line and the ground for the second network. The prepegs are used to stick substrates and in this case they have almost the same permittivity than the substrate. In this way, the price of the subarray is lower since the prepegs are much

cheaper than a substrate. All the plated holes are through plated holes to reduce the cost.

FIGURE 2.29: Stack-up of the multilayer board of the third prototype.

The manufactured third prototype is shown in Figure 2.30. Two branches hybrid circuits are used instead of three branches hybrid circuits because the line width of the three branches circuits are too small (the permittivity of RO4350B is higher than the permittivity of PTFE). To reduce the losses, the sequential rotation technique is used among subarrays instead of among elements, this way there is no need of loops in microstrip networks. The second network is not shown in the photo since it is embedded in the multilayer circuit.

FIGURE 2.30: Third prototype.

The measured results for the third subarray prototype are shown in Figure 2.31. The gain is around 16.5 dB, similar to the gain of the first subarray prototype. The differences between polarizations are lower than in previous prototypes. Regarding the polarization, the results are worse than in previous subarrays because sequential rotation is not used. The polarization purity of the final antenna prototype will be better since the sequential rotation technique will be used among subarrays. The patch used in this third model is optimized in coupling between probes. For this reason the matching is better than in the previous prototypes. The coupling between polarizations is not as good as the previous prototypes but is below 15 dB in the X-band.

2.4 Power distribution network system

This section describes the design of all the elements in the power distribution network. Subsection 2.4.1 explains the technology that have been chosen for the low loss power

FIGURE 2.31: Measured axial ratio (a), Copolar and Crosspolar gain (b) and S-Parameters (c) of the third subarray.

distribution network. Subsection 2.4.2 explains the design of transitions and T-dividers. A new ad-hoc connector system to improve the characteristics of the network is sketched in Subsection 2.4.3. The following Subsections (2.4.4, 2.4.5 and 2.4.6) describe three networks developed for the three antenna prototypes. Each network has an improvement compared to the previous one.

2.4.1 Stripline technology

One of the problems in traditional printed technologies is that the losses are very large. This is especially a problem in large distribution networks. Traditional waveguide technology could be an option for a low loss distribution network. The problem is that the width of waveguides is larger than the width of printed networks, making it difficult to design a complex network. Another problem of waveguides is that it is critical to assure a good metal contact among all parts, making the network more expensive and heavier.

Substrate Integrated Waveguides (SIW) could be an option. The problem at X band is that the ratio between the width of the waveguide for supporting the TE₀₁ mode and the height of the usual commercial substrates is large. This makes the design of vertical transitions more difficult. Moreover, the losses in SIW would be larger because all the wave is confined in a region with dielectric losses.

To reduce the conductor losses in printed technologies, the height or the distance between the line and the ground plane must be increased. To increase the distance between the line and the ground plane in microstrip technology, we can use either a higher substrate, increasing the weight and the price of the antenna, or a suspended microstrip, using foam to separate a thin substrate from the ground. The disadvantage of microstrip technology is that besides the conductor and dielectric losses, the line also has radiation losses. This sets a limit in how much we can reduce the losses in the structure as it is shown in Figure 2.32. Moreover, increasing the distance between ground plane and line, the coupling between lines and the influence of near objects are also increased. On the other hand, in stripline technology there are not radiating losses, therefore the total losses can be reduced more than in microstrip technology.

FIGURE 2.32: Losses vs height in printed technologies.

In Figure 2.33 is shown the general scheme of a suspended stripline. The line is printed in a thin substrate and supported either by metallic walls at some distance or by foam sheets. Since the ratio between the substrate height (h_s) and the distance between line and ground (h_1) is large, the effective permittivity is close to air permittivity. Moreover, the dielectric losses are small because most of the wave goes through the air. To have metal contact between the two ground planes, plated vias are used along the line.

The metallic walls and ground plane are made in aluminum. Since there is no need of high accuracy, all the plates can be cut with laser. The substrate used for the line is a PTFE-based substrate in the first prototypes and RO4350B in later prototypes because

of its easier and cheaper manufacturing process. In both cases the substrate height (h_s) is 0.254 mm.

FIGURE 2.33: Suspended stripline scheme.

As shown in Figure 2.32, the stripline losses can be reduced increasing the distance between the ground planes and the printed line. Figure 2.34 shows the simulated losses for a stripline varying this distance (h_1). As it is shown the losses can be reduced until certain limit in which increasing the line height does not reduce the losses significantly. For this reason a height of 2 mm has been chosen as a trade-off between low losses and low profile.

FIGURE 2.34: Losses vs height (h_1).

The main mode of the suspended stripline is a Quasi-TEM mode since there are two dielectrics (substrate and air). The line also supports TE modes, usually named as biplate modes. These modes are especially generated in transitions and some discontinuities as bends or T-dividers. The field distribution of these modes is shown in Figure 2.35. The TE mode presented in Figure 2.35 (b) has a cut-off frequency of 5 GHz.

To cut these modes, metal vertical walls are used. The distance between walls has to be larger enough for not interfering with the Quasi-TEM mode and smaller enough for cutting off the TE modes. The necessary minimum distance between walls can be calculated using the formula for the cut-off frequency of a TE₀₁ mode in a regular rectangular waveguide (equation 2.2). The ϵ_r in this case is the effective permittivity of

FIGURE 2.35: Supported modes in stripline.

the line that is close to 1. A cut-off frequency of 10 GHz is chosen for the first TE mode. For this frequency, the minimum distance between walls is 14 mm.

$$W = \frac{c_0}{2f_c \sqrt{\epsilon_r \mu_r}} \quad (2.2)$$

The line width for a specific impedance is detailed in Figure 2.36. Another advantage of using a low losses power distribution network is that the lines are wider. This means that high impedance lines can be used without having a limitation because of the manufacturing process (usually the limit is 0.2 mm line width). For a 50 Ω impedance the line width is 5.5 mm.

FIGURE 2.36: Line width vs impedance in stripline.

Table 2.2 summarizes the main parameters of the selected stripline.

Substrate height	0.254 mm
Distance between line and ground	2 mm
Line width for 50 Ω	5.5 mm
Distance between walls	14 mm
Minimum distance between lines to prevent coupling	14 mm

TABLE 2.2: Stripline summary.

2.4.2 Stripline design

In this section is explained the design of unitary elements such as vertical transitions or T power dividers.

2.4.2.1 SMA transition

TE modes are especially generated in vertical transitions as it is shown in Figure 2.37. In the Figure there is a stripline plus a vertical SMA transition. The Quasi-TEM excites a TEM mode in the SMA connector and a TE mode in front of the vertical transition. This TE mode increases the losses and the coupling among lines. These modes can be cut-off using a metal U-piece.

FIGURE 2.37: TE mode generation in vertical networks.

The designed transition is shown in Figure 2.38. The stripline ground planes are hidden to show the inner part. The inner conductor of the coaxial connector is soldered to the line. The transition is shielded with a metal structure. The dielectric of the coaxial connector (PTFE) is introduced inside the stripline until the substrate where the line is printed.

FIGURE 2.38: SMA transitions scheme.

Since the height of the stripline is electrically large, the transition from the coaxial TEM mode to the stripline Quasi-TEM mode has to be done smoothly. That is why the edge of the metallic piece that shields the transition has to be rounded as it is shown in Figure 2.39. When the metal piece is rounded the field coupled to the SMA is more uniform and more radial. The corners in the piece have to be also rounded in order to avoid

diffraction. A small segment of stripline with a different width is used at the beginning of the transition to improve the matching .

FIGURE 2.39: Field distribution in a SMA transition.

Figure 2.40 shows the matching of a SMA-stripline transition. The influence of the solder process is critical and has to be done carefully in order to have a good matching.

FIGURE 2.40: SMA to stripline transition.

2.4.2.2 SMP transition

In most of the cases the subarrays will be above the stripline. For this reason, a vertical transition or a transition to a SMP connector is necessary.

These vertical transitions have to be pluggable in some way because nothing can be soldered between the subarrays and the power distribution network since the distance should be small in order to maintain the low profile. Two approaches to connect subarrays and power distribution network have been followed. The first one is using commercial SMP coaxial connectors (pluggable connectors) and the second one is to design an ad-hoc vertical transition.

Usually SMP connectors are meant to be placed in regular microstrip technology with a small line width. Thus, it is necessary to design a transition between stripline and microstrip. Figure 2.41 shows one model of a SMP connector.

FIGURE 2.41: SMP connector.

This transition has to be designed very carefully since the ratio between the width of the stripline and the width of the microstrip is close to ten. This ratio causes to appear a capacitance between the line and ground that increases the reflection. Two ways has been followed to design the microstrip-stripline transitions.

The first way is shown in Figure 2.42. To compensate the capacitance that appears because the line width change, a small short high impedance stripline is introduced. This line represents an inductor that cancels the capacitance in the discontinuity. This can be easily done prolonging the microstrip line into the stripline as it is shown in Figure 2.42 (b).

FIGURE 2.42: First stripline to microstrip transition.

The measured results compared to simulations for the first kind of transition are shown in Figure 2.43 (b). The matching is below -20 dB in frequencies lower than 6 GHz and below -12.5 dB in frequencies between 6 and 10 GHz. The results are not good enough for the X band because the capacitance caused by the change of the line width increases with frequency. For this reason, it is not possible to compensate the capacitance with the small high impedance line in higher frequencies.

FIGURE 2.43: Realized first stripline to microstrip transition.

The second way of designing the stripline to microstrip transition is to reduce the line width gradually. The stripline width is reduced changing the impedance of the stripline from $50\ \Omega$ to $70\ \Omega$ as it is shown in Figure 2.44. In this way, the capacitance that appears with the change of the line width is reduced. Then, this change has to be reverted in the microstrip network where the SMP connector is soldered. To change the impedance a $\lambda/4$ adapter is used.

FIGURE 2.44: Second stripline to microstrip transition.

Several prototypes of the second transitions have been manufactured. The total losses of the transition can be divided in three parts: the SMP connector, the small microstrip line and the stripline. The measured losses for a stripline with two transitions (Figure 2.45 (a)) are 1.3 dB for 50 mm stripline length and 1.2 dB for 25 mm stripline length. This means that most of the losses are due to the SMP connectors and the small microstrip networks.

Therefore, the losses because of the transition to subarrays using SMP connector are around 0.6 dB (half of the losses of a through). The measured matching compared to simulations is shown in Figure 2.45 (b). The matching is better than 20 dB at the X band.

FIGURE 2.45: Results of the second stripline to microstrip transition.

2.4.2.3 Power dividers

Two things must be taken into account in the design of regular T-dividers in these networks. The first issue is that a V-shape cut must be done in the main arm of the T to match the divider as it is shown in Figure 2.46 (a). The second point is that multi $\lambda/4$ adapter is used instead of one single adapter. This is possible because the distance among outputs is usually bigger than λ_0 since the network feeds subarrays with several elements on them. The impedance variation used in the designs is a binomial distribution taken from [49].

In Figure 2.46 (b) is shown the simulated S-parameters of an asymmetric power divider. It has a good matching in a large bandwidth because of the use of multi-section $\lambda/4$ adapter.

FIGURE 2.46: Power divider.

2.4.3 Ad-hoc connector system

One of the problems detected in first prototypes is that the losses in commercial SMP connectors were too large. Thus, there are as many losses in the SMPs connectors and stripline-SMP transitions than in the whole stripline network. An ad-hoc connector is designed to reduce the losses. These connectors have to be pluggable since the two parts cannot be soldered because the space between the subarrays and the power distribution network is too small.

The connector has two parts. The first part is a coaxial line with an inner conductor of 3 mm diameter, this inner conductor is supposed to be soldered to the stripline. The second part is a smaller coaxial line with an inner conductor with a diameter of 1.3 mm, similar to commercial SMA connector, the inner conductor of the second part is supposed to be soldered to the subarray microstrip network. Both coaxial lines have PTFE as dielectric. A scheme of the connector is shown in Figure 2.47.

To achieve the continuity between the two inner conductors a virtual short circuit is used. Therefore, the smaller inner conductor surrounded with PTFE penetrates $\lambda_g/4$ inside the bigger inner conductor. Since at the end there is an open circuit, $\lambda_g/4$ before there is a virtual short circuit between the two inner conductors. To improve the matching, a small length of coaxial with high impedance is used. To achieve the high impedance line the smaller inner conductor is extended into the bigger coaxial connector.

The coaxial grounds could be coupled in the same way than the inner conductors. The only difference is that there is not PTFE between the coaxial grounds, so that the distance is now close to $\lambda_0/4$.

FIGURE 2.47: Ad-hoc connector model design.

The first manufactured connector prototype is shown in Figure 2.48. The prototype is a through with two connectors in both ends. The main body is made in aluminum and

the inner conductor is made in brass. The ends have been made mechanizing a standard SMA panel connector.

FIGURE 2.48: Ad-hoc connector prototype.

The simulation and measurement of the first prototype are shown in Fig 2.49. The measured matching is below -18 dB at the X band. The maximum losses are 0.25 dB at the X band. The losses are lower than the commercial SMP connectors losses that are around 0.3 dB for each (0.6 dB for the through with two connectors).

FIGURE 2.49: Measurement of the first ad-hoc connector prototype.

The next step is to design transitions from the two coaxial parts of the connector to the power distribution network stripline and to the subarray microstrip.

2.4.3.1 Connector-stripline transition

To align the connector with the stripline a small swivel is made in the inner coaxial connector as is shown in Figure 2.50 (b). This swivel increases the mismatch. For this

reason a trade-off has been reached between an easy manufacturing and a good electrical behavior. The swivel dimensions are 1 mm for D and 0.5 mm for P . The transition from stripline to the ad-hoc connectors is very similar to the transition to SMA presented in Section 2.4.2.1. In this case, to match the transition and to compensate the effect of the swivel there are three variables sketched in Figure 2.50 (a).

FIGURE 2.50: Stripline with ad-hoc connector scheme.

In Figure 2.51 is shown the prototype of a through and a T-divider with ad-hoc connectors. The simulation and measurements are shown in Figure 2.52. The measured matching is a bit different from the simulation. The RX-band is better matched than the TX-band, even though the matching at RX-band is below -14 dB. The losses are lower than 0.2 dB at the X band. Compared to the through with SMP connectors, that has around 1.4 dB of losses, the behavior is much better.

FIGURE 2.51: Stripline with ad-hoc connectors.

2.4.3.2 Connector-microstrip transition

There are a few things to take into account in the microstrip transition design. The most important one is that the transition should be easy to manufacture, meaning that all the mechanical parts should be big enough. The second one is that it is important to

FIGURE 2.52: Results of the through with ad-hoc connectors.

have a good alignment between the part soldered to the stripline and the part soldered to microstrip, to achieve this, a swivel is also used in the conductor soldered to the microstrip.

In Figure 2.53 is shown the scheme of the transition with some details. To prevent the generation of higher modes the transition is shielded with through plated vias. The problem is that these vias go from the microstrip network until the patches. For this reason it is a good idea to reduce the number of vias. In Figure 2.53 (c) is shown the optimal via position for three vias. There are two variables to match the transition. The first one is the radius of the aperture left in the microstrip ground (Figure 2.53 (b)). The second one is a gap between the PTFE and the circuit board, this way there is a series coaxial line with air as dielectric with a different impedance (Figure 2.53 (d)). Finally a small window is opened in the coaxial ground for taking out the microstrip line.

The simulation results of a microstrip through with this transition are shown in Figure 2.54. Multiple simulations are presented varying the microstrip length to check the behavior independently from the stationary wave. The simulated matching is below -20 dB in the whole X-band.

2.4.3.3 Final full connector

One of the problems detected in first prototypes is that the size of the inner conductor soldered to the microstrip is very small, complicating the manufacturing process. To make easier the manufacturing process the coaxial dielectric is substituted by air,

FIGURE 2.53: Connector-microstrip transition scheme.

FIGURE 2.54: Connector-microstrip transition matching simulation.

this way the inner conductor diameter is increased if the coaxial external diameter is maintained.

Another improvement deals with the connector height. The connector height is determined mostly by the $\lambda_g/4$ distance necessary to achieve the virtual short circuit. One way to reduce the height is to change the dielectric material inside the bigger inner conductor by a material with higher permittivity. The new selected material is ertacetal C (POM-C) that has a permittivity of 3.8 and it is easy to manufacture. The ertacetal tube is extended till the microstrip network as it is shown in Figure 2.55 (a) . Thus there are two dielectrics (POM-C and air) in the part of the connector that goes to the microstrip. Since the POM-C only fills a small portion of the coaxial, the effective permittivity is very close to the air permittivity.

FIGURE 2.55: Final full connector scheme.

In Table 2.3, there is a comparison between POM-C and PTFE connectors according to the parameters sketched in Figure 2.55 (c). As it is shown, the inner diameters of the final model are larger than in previous prototypes, therefore the connectors are easier to manufacture. The length of the connector is also reduced in 3 mm for the use of POM-C instead of PTFE.

	D1(mm)	D2(mm)	D3(mm)	D4(mm)	D5(mm)	L(mm)
PTFE model	10	3	2	1.3	4.1	9.3
Air + POM-C model	10	4.3	3	2	5.4	6.3

TABLE 2.3: Comparison between PTFE and POM-C model.

Figure 2.56 shows the manufactured first prototype of the final connector model. Figure 2.56 (a) and (b) show the inner conductors and the POM-C tube (unplugged and plugged). In Figure 2.56 (c) and (d) is shown the stripline where the connector is soldered. The stripline transition has been re-optimized. Figure 2.56 (e) shows the microstrip in which is soldered one of the inner conductors. The mounted connector is in Figure 2.56 (f). To measure the connector a standard SMA connectors are necessary, for this reason there is a SMP plus a SMP to SMA transition in the microstrip. There is also a SMA at the opposite end of the stripline. To attach the connector two nylon screws are used.

The measurements versus simulation of the full ad-hoc connector presented in Figure 2.56 are shown in Figure 2.57. The measured matching is below -12 dB. The result is acceptable taking into account the mechanical precisions. The losses are below 1 dB at X-band. The losses include the losses of the SMA-SMP transition, the SMP connector, the small microstrip line, the ad-hoc connector and the small stripline network (see Figure 2.56 (f)).

FIGURE 2.56: Manufactured full connector.

2.4.4 First design

The first design is shown in Figure 2.58 (a). This network is used to distribute the power to an antenna of 4 x4 subarrays. The network is connected to the subarrays through SMP connectors. In this design only the transitions are shielded. Figure 2.58 (c) shows the measured outputs of one quadrant. As it can be seen there are peaks in the transmissions because TE modes are excited not only in the transitions but also in some other parts of the network.

To solve this problem, all the network is shielded as it is shown in Figure 2.58 (b). Figure 2.58 (d) shows that the peaks of the previous network have disappeared.

FIGURE 2.57: Results of full ad-hoc connectors.

FIGURE 2.58: First example of power distribution network.

2.4.5 Second design

The second design consists in two networks of 3x3 outputs with tapering. Each network is for one polarization. In this case, since all the outputs need to have the same phase,

meander lines have been used to compensate paths. To avoid the problems detected before with TE modes, the line has been shielded with metallic walls as shown in Figure 2.59. The input of this network is done through a SMA connector and SMP connectors are used to connect the network with the subarrays.

FIGURE 2.59: Second power distribution network.

As it is presented in Figure 2.60 (a), the matching for the two polarizations is better than 15 dB in the whole X band. Figure 2.60 (b) shows the outputs for one polarization. The transmissions show that the tapering is well done since the corner subarrays (C) receive in average 1 dB less than the edge subarrays (B) and 2 dB less than the one in the center (A).

FIGURE 2.60: Measured matching and transmissions for second design.

The losses of the second power distribution network for both polarizations are shown in Figure 2.61. The losses are below 1.5 dB for one polarization and below 2 dB for the other one.

FIGURE 2.61: Losses of the second power distribution network.

2.4.6 Third design

The third design is a network for an array of 2x2 subarrays. The size is 200 x 200 mm. There are two main differences with previous prototypes. The first difference is that in the third design the ad-hoc connectors presented in section 2.4.3 are used. The second one is that sequential rotation technique is used, meaning that each subarray is rotated 90° and the feeding network compensates this rotation with a phase shift of 90° as it is shown in Figure 2.62 (b). The sequential rotation technique is applied for improving the polarization purity of the antenna.

FIGURE 2.62: Third design.

The matching and losses of the network are presented in Figure 2.63 (a). The matching has been moved to lower frequencies. Even though, the matching is below -14 dB at the X band. The losses are lower than 0.6 dB at the X band. Compared to previous prototypes the losses are lower for two reasons: the use of ad-hoc connectors instead of commercial SMP connectors and the omission of the small microstrip network to place the SMP connector.

The transmission phases are shown in Figure 2.63 (b). The difference among each output is 90° at the center frequency.

FIGURE 2.63: Third design measurements.

2.5 Complete antenna

This section shows the results for the first two prototypes.

2.5.1 First prototype

The first prototype of the antenna is shown in Figure 2.64. The antenna weighs less than 3.5 kg. The subarrays have been protected with an extra foam sheets. The first prototype is a 4x4 subarray antenna with only one polarization.

FIGURE 2.64: First prototype.

The measurement of the first prototype are not as good as expected (see Figure 2.65). The main problem is the power distribution network that is not well matched. Even though the gain in the lower band is close to 24 dB and in the upper band is close to 21 dB. Figure 2.65 (c) shows the measured radiation pattern for $\phi = 90^\circ$ for low, central and high frequency. The lobe positions and levels for the central frequency are very similar to the ones calculated in Figure 2.6 for 16x16 elements. This first prototype does not meet with the requirements and it has only one circular polarization. However it helped to prove the concept and to learn about manufacturing processes for next prototypes.

FIGURE 2.65: Matching (a), axial ratio and gain (b) and radiation patterns for $\phi = 90^\circ$ (c) of the first prototype.

2.5.2 Second prototype

Figure 2.66 shows the second prototype mounted in the positioner of the UPM anechoic chamber. For the problems explained in Section 2.3.4, the second prototype has only 3x3 subarrays. It has both circular polarizations. The behavior of the subarrays and power distribution network is better than in the first prototype, thus the antenna behavior is expected to be better.

FIGURE 2.66: Second prototype in the anechoic chamber.

The measurements of the second prototype are shown in Figure 2.67. They are only presented the results for one of the polarizations since the results for the other one are very similar. The matching for each polarization is better than in the previous prototype. In this prototype, the power distribution network is better matched than the subarrays, hence the antenna matching is similar to the subarray matching. The coupling between polarization is better than 20 dB in the whole X band. The gain is around 26 dB in the low band and around 25 dB in the high band. The axial ratio is lower than 2 dB and meets with the specifications.

2.6 Conclusions and future work

The design and implementation of a man pack antenna for satellite communications at the X band has been presented. The main characteristics of the antenna is that it implements both circular polarization simultaneously, it has a modular design (easy to fabricate in serial production) and it is low profile.

A first prototype has been built showing some difficulties with the connector system and the power distribution network. In this first prototype the second polarization could not be included because the area in the subarray network is too small for the second polarization. For this reason in the second prototype the subarray size is increased,

FIGURE 2.67: Matching (a), axial ratio and gain (b) and radiation patterns for $\phi = 90^\circ$ (c) of the second prototype.

reducing the number of subarrays in the antenna. The power distribution network is also improved in the second prototype. The results for the second prototype meet with the requirements although some things are susceptible for improvements.

The main problem of the second prototype is the price. For this reason in the design of the third prototype new cheaper materials are used. The SMP connectors are substituted by an ad-hoc system. Although several parts of the third prototype has been proved, the final prototype is not built because of a lack of funds. The final ad-hoc connector did not have the expected behavior although the behavior was not that bad taking into account the mechanical precisions. For this reason the connector should be manufactured with higher mechanical precisions.

Therefore the first future line is to manufacture and test the final prototype. The expected results for the third prototype are better than in the second one and with a

cheaper antenna price.

The mechanical properties can also be improved. The antenna weight can be reduced substituting the metal plates by plated plastic. The screws can also be substituted by rivets.

Since the antenna has a modular design (all the subarrays are equal) antennas with different shapes can be designed only by changing the power distribution network. As an example a larger antenna with a rectangular shape for radar applications could be designed.

The antenna is supposed to be steered to the satellite with a mechanical system. The problem is that it is very complex to steer a high gain antenna to a satellite since the beam width is very narrow. For this reason, an electronic steering system can be added. A full electronic steering system could be very expensive, thus, a mechanical-electronic steering system could be helpful and cheap. In this way, the antenna can be steered through a mechanical system with some pointing error and electronically correct this error. The electronic steering angle should not be very large, $\pm 5^\circ$ at most.

The proposed system is based on phase shifters. There would be a phase shifter in each of the subarrays, thus the phase is changed among each 4x4 elements. So to speak it is like a phase array of 4x4 elements with a distance among elements equal to the subarray size. The limit in the steering angle that can be achieved is fixed by the tolerable level of the first grating lobe. To include the phase shifters in the design there are two main options. The first one is to use a commercial phase shifters. The problem is that losses in commercial phase shifters are large, for this reason a T/R module must be included to compensate these losses. This T/R module is complicated to design since the antenna has to work in RX and in TX in all the band and simultaneously. The second option is to design an specific phase shifter with fewer losses similar to the one presented in Chapter 3.

The Array factor for $\phi = 0^\circ$ can be calculated according to equation 2.3 where α is the phase shift among subarrays.

$$AF(\theta) = \sum_{m=0}^{N/4-1} e^{jm(4dk_0 \sin(\theta) + \alpha)} [1 + e^{jk_0 \sin(\theta)} + e^{j2k_0 \sin(\theta)} + e^{j3k_0 \sin(\theta)}] \quad (2.3)$$

The steering direction is :

$$\theta = \sin^{-1} \left(\frac{-\alpha}{4dk_0} \right) \quad (2.4)$$

FIGURE 2.68: Radiation patterns changing the phase of each subarray.

And the first grating lobe is in the angle :

$$\theta = \sin^{-1} \left(\frac{2\pi - \alpha}{4dk_0} \right) \quad (2.5)$$

Figure 2.68 shows the radiation pattern of the antenna changing the phase shift among subarrays. If a maximum grating lobe level of 10 dB is accepted, then the maximum steering direction is $\pm 4.5^\circ$. For an steering direction of 10° the grating lobe has almost the same level than the main lobe.

Chapter 3

Reflective phase shifter based on varactors at X band

3.1 Introduction

This chapter describes the design and manufacturing of a novel configuration of a reflective phase shifter. Phase shifters are mostly used in electronically steered antennas. To steer electronically an antenna, there are two main solutions: phase shifters and steering networks. The main disadvantages of steering networks (see [52]) are the size and the complexity when the number of radiating elements and steering directions are high. On the other hand, phase shifters usually take less space and they can be placed in each radiating element, allowing to have larger array size and a quasi-continual steering scanning.

There are two main phase shifter families: those that commute among paths and those that change electronically the physical behavior of something in the circuit.

The first family commutates among paths with different electrical lengths. In this way, every path phase-shifts a different quantity. To commute among paths, PIN based switches are usually used. The size of these phase shifters can be large if they are implemented in a common technology, although if MMIC technology is used, they can fit in a small chip.

The second family changes the physical behavior of the microwave line or the behavior of some component. The line behavior can be changed either changing the permittivity or the permeability. One example of a phase shifter that changes electronically the permittivity is shown in [53]. It uses liquid crystal that changes the orientation of its molecules depending on the static voltage that is applied between its borders. The

permeability can be changed by using ferrites, the magnetic field inside a ferrite is modified with a solenoid controlled by a static current. The ferrite permeability varies according to the magnetic field (see [54]). The microwave line parameters can be also altered using metamaterials. The line capacitance per length can be modified using Microelectromechanical Systems (MEMS) (see [55]). The MEMS are placed as parallel elements between a coplanar line and the ground. Depending on the MEMS state, the parallel capacitance is modified, changing the line capacitance per length and the line phase constant. The drawback of MEMS is that their life time is not very high and they require high lab capabilities to be manufactured.

Varactor diodes can also be used to change the phase of a signal. These diodes change their capacitance value depending on the inverse voltage that is applied to them. If they are used in a reflective phase shifter, the phase reflection depends on the capacitance value. An example of reflective phase shifter is shown in [56]. The drawback is that the phase range does not cover 360° , hence, it is necessary to use several stages, increasing the size and losses. The proposed phase shifter overcomes this drawback with a series-shunt varactor configuration of only one stage.

Section 3.3 introduces the concept of the reflective phase shifter based on varactors and proposes some ideas to extend the phase shift by placing more diodes in series. Section 3.4 explains the design and different optimizations of series-shunt varactor configurations. A first prototype is shown in Section 3.5.

3.2 Reflective phase shifters based on serial varactor diodes

The basic scheme of a reflective phase shifter is presented in Figure 3.1. Equation 3.1 shows the S parameters of an ideal branch line circuit. If the two output ports in the branch line circuit (port 2 and 3) are loaded with the same impedance (same reflection coefficient), the reflected signals in these ports are sum out of phase in the input port (port 1) and in phase in the isolated port (port 4).

$$[S]_{BLC} = \begin{pmatrix} 0 & j & 1 & 0 \\ j & 0 & 0 & 1 \\ 1 & 0 & 0 & j \\ 0 & 1 & j & 0 \end{pmatrix}, \quad [a] = [S] \cdot [b] \quad \text{with} \quad \left. \begin{array}{l} a_2 = \rho \cdot b_2 \\ a_3 = \rho \cdot b_3 \\ a_4 = 0 \end{array} \right\} \Rightarrow \begin{array}{l} b_4 = -j\rho \cdot a_1 \\ b_1 = 0 \end{array} \quad (3.1)$$

FIGURE 3.1: Reflective phase shifter basic scheme.

If the reflection coefficient phase can be controlled and the reflection coefficient amplitude is close to one, the device will work as a phase shifter. Hence the matching of the circuit will be related to the isolation coefficient of the branch line circuit plus the similitude between varactors.

3.2.1 Varactor lumped element model

The lumped element model of a real varactor is shown in Figure 3.2. The series inductor is related to the package. In some cases, this series inductor could be higher than the optimal series inductor to obtain the maximum phase shift with the varactor. The series resistant represents the losses in the varactor. The lower the capacitance range is, the larger this resistance is. However, to obtain the maximum phase shift, low capacities are desirable. The parallel capacitance is added to the varactor capacitance, therefore is related to the minimum value of the capacitance range.

FIGURE 3.2: Varactor lumped element model.

The lumped element parameters provided by the manufacturers for every varactor analyzed in this chapter are summarized in Table 3.1

Model	Cap. range (pF)	Rs (Ω)	Ls (nH)
SMV1430	0.3-1.2	3.15	0.4-0.7
SMV1405	0.6-2.7	0.8	0.4-0.7
SMV1408	1-4.1	0.6	0.4-0.7
MA46580	0.2-1.1	x	x

TABLE 3.1: Lumped element parameters of commercial varactors.

3.2.2 Analysis of the phase shift produced by the varactor

For a first analysis the reflection coefficient of a LC network (Figure 3.3) is obtained (Equation 3.2).

FIGURE 3.3: LC model.

$$\rho = \frac{1 - LC\omega^2 + jZ_0C\omega}{1 - LC\omega^2 - jZ_0C\omega} \quad (3.2)$$

The phase shift in the reflected signal is related to the phase of this reflection coefficient. Figure 3.4 shows the maximum achievable phase shift that can be obtained with four commercial varactor models in function of the series inductor. The frequency has been set to the central frequency at X band. As shown, the phase shift is larger when the capacitance range is lower. Furthermore, the optimal series inductor is also larger when the capacitance range is lower. This inductance is not the resonant value for the central capacitance value but it is closer to the resonant inductor for the upper capacitance value.

FIGURE 3.4: Phase shift for several varactor models.

To take into account the losses, the internal resistance in the diode must be included, as depicted in Equations 3.3 and 3.4.

$$\rho = \frac{R_s\omega C - Z_0C\omega - jLC\omega^2 - j}{R_s\omega C + Z_0C\omega - jLC\omega^2 - j} \quad (3.3)$$

$$Losses(dB) = -20 \cdot \log(|\rho|) \quad (3.4)$$

The losses for every diode are shown in Figure 3.5. The abscissa does not have a fix value since the capacitance range is different depending on the diode. The diode with lower capacitance range (larger phase shift) has the largest losses. For this reason the second diode would be a good trade-off between losses and achievable phase shift. Figure 3.5 takes into account only the losses because of the varactor, the final phase shifter will also have losses due to the microstrip lines and the hybrid circuit.

FIGURE 3.5: Losses depending on the varactor diode.

To extend the achievable phase shift, the capacitance value can be reduced placing some varactors in series. This way, the capacitance of a series union is the capacitance of one single varactor divided by the number of varactors. Figure 3.6 shows the phase shift combining some varactors in series. The minimum package inductance that has this varactor is between 0.4 nH and 0.7 nH, therefore the varactors will have more or less the largest phase shift without the need of synthesizing and additional inductance.

The losses of several varactors in series are shown in Figure 3.7. As it is logical they increase directly proportional to the number of varactors.

The microstrip lines between diodes should be added to have a more realistic model (Figure 3.8). These lines have to be longer enough to make easier the soldering/bounding process. In a first step the distance between diodes is taken constant.

FIGURE 3.6: Phase shift for several SMV1405 and SMV1430 varactors in series.

FIGURE 3.7: Losses for several varactor in series SMV1405.

FIGURE 3.8: Model for including lines between diodes.

The impedance before and after every diode can be obtained according to Equation 3.5.

$$\begin{cases} Z_i = Z_{i-1} + j\omega L_{diode} + \frac{1}{j\omega C_{diode}} + R_{diode} \\ Z_{i+1} = Z_0 \cdot \frac{Z_i + Z_0 \cdot \tanh(\gamma l)}{Z_0 + Z_i \cdot \tanh(\gamma l)} \end{cases} \quad (3.5)$$

The achievable phase shift changing the distance between diodes is plotted in Figure 3.9.

According to the graph, there is an optimal length for obtaining the maximum phase shift. This length is larger enough for the soldering/bounding process.

FIGURE 3.9: Phase shift for four SMV1405 in series changing the line length among them.

Eventually, to have a more realistic model the branch line circuit is added. This circuit can be added as an ideal S-parameter matrix or as a real S-Parameter matrix taken from a simulation. The second option is chosen using the arbitrary S-parameter matrix concatenation taken from [57]. The port model for the calculation is shown in Figure 3.10. It has three S-matrix: the 4x4 matrix of the branch line circuit and the two matrices corresponding to the varactor chains. The branch line circuit matrix is obtained through a full wave simulation and the varactor matrices are taken from the previous calculations.

FIGURE 3.10: Phase shifter S parameter model.

The non-null terms of the connexion and S-parameter matrices are shown in Figure 3.11.

The phase shifter matching for every state is shown in Figure 3.12 (a). This matching is especially related to the matching of the hybrid circuit since the impedance of the varactor chains does not have a strong influence as long as this impedance is equal for both chains. The transmissions for every state are shown in Figure 3.12 (b). The losses are similar to the ones calculated in Figure 3.7 for three diodes, adding the losses of the branch line circuit.

FIGURE 3.11: Model matrixes for obtaining the S-parameters.

FIGURE 3.12: Matching and S21 for three SMV1405 in series.

Figure 3.13 shows the result for concatenating three SMV14030 diodes instead of SMV1405 diodes. As it has been explained before, these diodes have larger losses but with a lower capacitance, therefore the achievable phase shift is larger. Although the losses are a bit large, they are not higher than the losses of a commercial phase shifter. Even so, the phase shift does not cover 360° .

One solution to cover the 360° with the previous designs is to use a mix varactor-PIN diode phase shifter as shown in Figure 3.14. The PIN diodes would be use as switches. This way, the switches would commute among paths with a phase difference of 0° , 90° , 180° and 270° . Therefore, the varactors would provide a phase shift from 0° to more than 90° and the quadrant of the 360° circumference would be selected with the switches. If the selected varactor configuration covers more than 180° , it would only be necessary a switch that commutes between paths with 0° and 180° .

FIGURE 3.13: S_{21} and phase shift for three SMV14030 in series.

FIGURE 3.14: Varactor plus switch model.

3.3 Reflective phase shifter based on serial-shunt varactor configuration

A novel series-shunt configuration is proposed (Figure 3.15). The idea is to cover the 360° with only two varactors in the chain. From now on, the selected varactor diodes are Macom MA46580 since they are similar to the Skywork ones and can be easily ordered from Spain.

FIGURE 3.15: Series-shunt varactors.

The impedance in every point of Figure 3.15 is shown in Equation 3.6. The characteristic impedance of every line is the same although these impedances can be distinct if more degrees of freedom were needed.

$$\left\{ \begin{array}{l} Z_1 = Z_0 \cdot \tanh(\gamma \cdot l_1) \\ Z_2 = Z_1 // Z_{varac} \\ Z_3 = Z_0 \frac{Z_2 + Z_0 \cdot \tanh(\gamma \cdot l_2)}{Z_0 + Z_2 \cdot \tanh(\gamma \cdot l_2)} \\ Z_4 = Z_3 + Z_{varac} \end{array} \right. \quad (3.6)$$

The reflection coefficient of the chain can be obtained from Z_4 . Then this coefficient can be inserted into the algorithm that concatenates S-matrices. An optimizer is added to the routine. This optimizer is based on a genetic algorithm. The main parameters to be optimized are: the line length between diodes (l2) and the line between the diode joint and the ground (l1). The inductances associated to every varactor could be also taken as a optimization parameters. These inductances can be variated by adding a short high impedance lines, but previous calculations show that the optimal inductances are in the range of the varactor case inductances. The line characteristic impedances are set to 50 Ω .

Two different goal scenarios are defined for the optimizations. The first scenario optimizes the phase shift to the maximum value. Although only 360° are needed, the simulated maximum phase shift can be reduced in the prototype due to unexpected manufacturing problems, or problems in the varactor characterization. For these reasons if the simulated phase shift is larger than 360° it is more plausible that the prototype has at least 360° .

In the second optimization the goal is to reduce the losses to the minimum value but with a phase shift of at least 360° .

The results for both optimizations are shown in Figure 3.16. The phase shift for the first case is larger than 500° (more than enough for a phase shifter). Two cases are taking into account for the loss calculation, depending on the minimum and maximum resistance of the varactor. The losses for the optimization based on phase-shift improvement are between 1.5 and 6 dB. For the optimization based on loss reduction, the losses are between 0.8 and 2.25 dB. The line lengths are 1 and 3 mm for the first case and 1.1 and 6.5 mm for the second one.

FIGURE 3.16: Results for the two serial-shunt optimizations at 7.75 GHz.

The case optimized for minimum losses and with minimum varactor series resistant is simulated in a full wave simulator (Figure 3.17 (a)). The losses and phase shift is very similar to the ones calculated before, showing that the ad-hoc simulator can be used for designing these phase shifters. The phases for each capacitor value are presented in Figure 3.17 (b). The phase is not very lineal at the lower capacitance values. This could lead to some phase errors, especially when the antenna transmits a broadband signal. The achieved phase shift varies from 300° at the highest frequency and 400° at the lowest frequency. However, the simulation shows a good behavior if the phase shifter is used in the middle frequencies.

FIGURE 3.17: CST phase simulation of the phase shifter based on series-shunt varactors.

The simulated losses and matching are presented in Figure 3.18. The behavior in terms of losses is very similar to the previous simulations. There is a peak in the losses at 0.32

pF. This peak is also present in the ad-hoc simulations (Figure 3.16 (b)). The matching is better than -8 dB in all the X band and for all the phase shifter states.

FIGURE 3.18: CST S11-S21 simulation of the phase shifter based on series-shunt varactors.

3.4 Biasing of the phase shifter

The varactor diodes have to be polarized in inverse, hence, the positive voltage has to be applied in the cathode. They could be polarized independently or with the same voltage. For the phase shifter purpose, there is not any advantage in polarizing them independently since the maximum-minimum achieved phase is got when both have the maximum/minimum capacity. For this reason the voltage is applied in the diode joint. The anode from both diodes is put to the ground. To have a RF choke the bias lines have a distance of $\lambda_g/4$ from the RF line until the ground. Finally, high value capacitors are used when a ground is needed for RF but not for DC.

The diodes change the capacitance depending on the applied inverse voltage. The phase shift is not linear with this voltage so it has to be calibrated and stored in a look-up table. The easiest way to provide a variable DC voltage is to use a voltage divider with a potentiometer. This potentiometer can be controlled directly from a I2C bus. Since the varactors do not consume DC power, most of the consumed DC power is due to the dissipation in the resistances. Therefore the fix resistance value and the potentiometer range should be large enough for minimizing the DC current consumption.

The final configuration is shown in Figure 3.19.

The RF choke is shown in Figure 3.20. It consists of a high impedance line with a length of $\lambda_g/4$ terminated in a short. This short is isolated from RF by a high value capacitor.

FIGURE 3.19: Bias scheme.

The impedance that is seen in RF is higher than $15\text{ K}\Omega$ at the central frequency (Figure 3.20). The impedance value is not that high at the extreme frequencies. This could affect the phase shifter behavior at these frequencies since the reflection coefficient could be a bit influenced by this parallel impedance.

FIGURE 3.20: CST bias simulation.

3.5 First prototype

The photograph of the first prototype is shown in Figure 3.21. This prototype corresponds to the design for minimum losses with a phase shift of at least 360° . Several boards were manufactured changing the distance between the diode joint and the ground. This distance controls the optimum point in which the phase difference is maximum. Although this point has been adjusted in simulation to be at central frequency, there is a lot of things that are not taking into account in the simulation such as the real diode

parameters, the influence of the soldering process, the alignment of all the diodes... The results presented in this section corresponds to the board with the best behavior.

The diodes are hardly visible in the photograph since their dimensions are 0.13×0.27 mm. Since the university does not have the equipment to properly bound these diodes to the board, they are pasted with conductive epoxi. In this way, the phase shifter can be connected to the network analyzer and the diode can be moved until the response is more or less the expected result. However, this is just a way to prove the phase shifter behavior although the performance would be much better if proper bounding process were used.

FIGURE 3.21: First prototype photograph.

The measured phase is shown in Figure 3.21. Two things must be taken into account in the interpretation of these measurements. The first one is that the varactor capacitance are highly not linear with the voltage, the variation with lower voltage is higher than with larger voltages. The second thing is that the varactor capacitance and the inverse voltage are inversely proportional. As shown, the phase does not cover the 360° but it is above 300° in all the frequencies. The difference in the achievable phase shift at the X band (7.25-8.4 GHz) is 35° . Thus, the measurement show that the achievable phase shift of the traditional reflective phase shifters based on varactors can be extended using this series-parallel configuration.

The matching and S21 of the phase shifter for some states are shown in Figure 3.23. The matching for the best case is lower than -10 dB, the worst measured case has a matching better than -8 dB. This matching should be better taking into account that the reflections in each output of the hybrid circuit should go to the phase shifter output and not to the input output. This would be true if the varactors were identical and

FIGURE 3.22: Measured phase of first prototype.

identically placed. Thus, the results are very influenced by the soldering capabilities. As far as the losses are concerned, it seems that most of the states have minimum losses at the same frequency where the phase shift is maximum. However, there are some states in which the losses are higher and the loss minimum is at frequencies below the frequencies with maximum phase shift. For this reason, it is better to have the phase shifter tuned to have the maximum phase shift in higher frequencies so that the losses were not that high. The losses are lower than 5 dB in the whole X band. The simulated losses were about 2.5 dB. The main difference causes could be: higher series resistance in the varactor (the varactor series resistant specified by the manufacturer has been measured in lower frequencies) and the not optimal bounding process that leads to a relatively bad matching. The figure also includes a reference losses that are the hybrid circuit losses without the varactors.

FIGURE 3.23: Measured matching and S21 of first prototype.

3.6 Conclusions and future lines

A novel design of a reflective phase shifter based on varactors has been presented and tested. The main advantage of the proposed solution is that is capable of obtaining more than 360° with only one stage. Two designs have been proposed: one with minimum losses and one with larger phase shift. The first prototype corresponds to the design with minimum losses but at least 360° phase shift. The measurements show that the losses are higher than in the simulations. Nevertheless, the losses are below 5 dB which are lower than in commercial phase shifters. The measured phase shift does not cover 360° but it is above 300° . In future prototypes the optimizations can be done to cover 400° more or less since this phase shift will be reduced in the real prototype.

The first future line is to re-manufacture the phase shifter but with more accurate bounding process. Then, they should be integrated in a small array to show the capabilities. The phase shifters has been designed for the X band, but the varactors used in the design can work in frequencies above the ka band. Moreover, nowadays there is not a commercial phase shifter that works in the ka-band, thus, they can be used in ka band antennas where available options for phased arrays are lower.

Chapter 4

Phased array for UAV communications

4.1 Introduction

The objective of this antenna is to provide a high capacity system for emergency situations. The first studied scenario is a wildfire. The communications will be between an advance control post and different aerial assets (helicopters, Unmanned Aerial Vehicles ...). These aerial assets will be in charge of capturing information and taking pictures of the disaster-affected area. The antenna presented in this chapter is the antenna of the advanced control post.

The selected communication standard is IEEE 802.16 or WiMAX (see [58]) at the 5.4 GHz band. The reference area of a wildfire is 10x10 km and the maximum distance between the control post and an aerial asset is 40 km. Thus, the usual range for WiMAX mobile communications must be extended.

The communication system in the aerial assets cannot be modified. Thus, to improve the range it is necessary to improve the equivalent isotropically radiated power (EIRP) and reception sensitivity (G/T) in the control post. Both parameters can be improved either increasing the antenna gain or the transmitted power and the receptor sensitivity (receptor noise figure).

- To improve the antenna directivity, the beam-width must be decreased, reducing the antenna angular coverage. To keep the link, the antenna should be provided with steering capability.

- Increasing the transmitted power is not very problematic but reducing the noise figure in the receptors is not that simple. It would be necessary to refrigerate the receptors to cryogenic temperatures, making the system much more complicated and expensive.

The first option is chosen. An array antenna with a 24 dBi directivity is proposed. The improvement compared to conventional antennas is around 8 dB, making it possible to establish a link with an aerial asset at a distance of 50 km. For shorter distances, the link capacitance would be higher since more efficient WiMAX modes can be used. Figure 4.1 shows the necessary antenna gain for every possible WiMAX modulation, the more efficient the modulation is, the more antenna gain is needed.

FIGURE 4.1: Necessary antenna gain for 50 km range depending on the WiMAX mode.

In the horizontal plane, the antenna is steerable so that the main beam can be switched to track down aerial assets or to connect with several assets multiplexing the signal in time. In the vertical plane, there is a shaped radiation pattern with the maximum pointing to the horizon. Hence, the antenna gain decay is compensated by the propagation losses decrease in directions different from the horizon. Examples of the design and manufacturing of shaped beams can be found in [59] and [60]. Figure 4.2 shows the communication with two Unmanned Aerial Vehicles (UAVs). One of them is closer so that the antenna has less gain. The other one is at the horizon so that the antenna gain is the highest.

The initial antenna specifications are summarized in Table 4.1.

Two main options can be applied for steering the antenna electronically: a mutibeam antenna and commercial phase shifters. Multibeam antennas have several beams available all the time. If there is only one base station, the pointing beam can be selected

FIGURE 4.2: Shaped beam scheme.

ANTENNA SPECIFICATIONS

<i>Parameter</i>	<i>Specification</i>	<i>Unit</i>
Frequency range	5.4 to 5.7	GHz
Maximum directivity	24	dBi
Horizontal steering range	± 40	$^{\circ}$
Horizontal beam width	12	$^{\circ}$
Horizontal beam width pointing to 50°	20	$^{\circ}$
Vertical beam width $^{\circ}$	10	$^{\circ}$
Polarization	Vertical linear	
Number of elements	8x8	
Vertical distance among elements	40	mm
Horizontal distance among elements	30	mm

TABLE 4.1: Main antenna specifications.

through a commutation matrix. One example of a cylindrical EBG structure antenna with several beams is shown in [61]. The problem with these antennas is that they do not have the necessary directivity since each beam only uses a reduced number of radiating elements. Another way to have a multibeam antenna is using a steering network such as a Butler network (see [62] and [63]). These networks can have similar losses than commercial phase shifters but they take up more space. Moreover, the number of beams are related to the number of radiating element rows. If more beams are required, a multibeam lossy network can be also used. In this kind of network, part of the steering network outputs are loaded with 50Ω . The other outputs feed the radiating elements. Since sixteen beams are necessary and there is only eight rows of radiating elements, half of the steering network outputs would be loaded, thus, there would be a loss of 3 dB in transmission. An example of the design and realization of these lossy networks is shown in [64].

The second option to steer the antenna is to use commercial monolithic phase shifters. These phase shifters are placed in every row. The problem with commercial phase shifter

is that they have a lot of losses. To overcome these losses a T/R module has to be used in every row. Since WiMAX technology is a very common standard, multiple commercial T/R modules can be found. The advantage of monolithic phase shifters is that the available beams are much higher, therefore, the pointing losses are reduced. Another advantage is that they take up a small space, thus, they can be easily integrated in the antenna.

All the different antenna subsystems are summarized in Section 4.2. The radiating element and shaped beam designs are shown in Section 4.3 and 4.4. The selected monolithic devices and the gain and noise studies are presented in Section 4.5. The antenna is controlled through an I2C bus for changing among pointing directions, the estimated control speed and all the elements used to control the antenna are depicted in Section 4.6. Each phase shifter and attenuator has a different phase and module depending on its state, thus, it is necessary to establish a calibration scheme to have the best antenna performance. This calibration scheme is shown in Section 4.7. The final prototype and its measurements are presented in Section 4.8. Finally, the conclusions and some future lines are explained in Section 4.9.

4.2 Antenna structure

The scheme of the antenna system is depicted in Figure 4.3. This scheme includes the antenna and the WiMAX base station since the antenna has to be controlled by this station.

The antenna is formed by seven subsystems:

- Radiating subsystem: row of eight radiating elements fed by a passive distribution network. The radiation pattern is a shaped beam in the vertical plane and has a beam width of 90° in the horizontal plane.
- T/R subsystem. These subsystem is formed by commercial T/R front-end modules plus switches to commute between input and output. The T/R module includes an internal switch, a LNA, two power amplifier stages and a power detector.
- Phase shift subsystem. Commercial phase shifter are used to change the phase among rows. Four bits are needed to control each phase shifter.
- Amplitude control. The T/R amplifiers could have a gain difference of 1 dB among them. Phase shifters have also a different attenuation depending on the state. For these reason it is necessary to equalize every row of the array. To control the

FIGURE 4.3: Antenna scheme.

amplitude of each row 6 bit digital attenuators are used. These attenuators are also used in the cases in which the base station gives more power than needed. In these cases, the attenuator will prevent the saturation of the power amplifiers. Eventually, they can also be used to attenuate all the rows except one. In this way, the antenna would have an omnidirectional mode in the horizontal plane.

- **Control subsystem.** The control subsystem is based on expanders that are controlled by an I2C bus. The i2C bus is expanded into 40 bits for each expander, thus at least three expanders are necessary for controlling all the elements in the antenna. T/R modules are also controlled with five bits to change among the different Tx and Rx operation modes. The commutation between Tx and Rx has to be done faster to maintain the link. Therefore, these five bits can be also controlled independently by the base station.
- **Amplitude sampling.** Although the different attenuation of each state of the phase shifter can be calculated statically, the difference among amplifiers cannot be calibrated unless they are measured one by one. Moreover the amplifiers can vary their gain over the time. For this reasons, T/R modules include power detectors

that give different voltage depending on the output power level. This voltage levels have to be digitalized with 8 Analog Digital Converters. These converters can be controlled with the I2C bus.

- Power supply. The antenna is fed through a single line from the base station. Since the T/R modules need a different voltage than the rest of components, different DC-DC converters are used. Moreover, the T/R feeding can be divided in power amplifier feeding and low noise amplifier feeding. This way the power amplifier can be shut down, reducing the consumed power when the antenna is only listening.

4.3 Radiating element and antenna analysis

This Section shows the design of the radiating subsystem. It includes the design of the radiating element and the feedings to obtain the shaped beam. Two radiating elements are compared in the simulations: a double stacked patch inside a cavity and without cavity.

4.3.1 Radiating element

The selected radiated element is a double stacked circular patch. The first patch is printed in a Ro4350B substrate and it is stacked with the RF and control layers. The patch substrate is isolated from the other layers by an internal stacked ground plane. The second patch is printed in a thin substrate and separated from the active patch by a foam sheet.

To reduce the influence of the coupling when the antenna is steered, metal cavities that surround the patch can be used. In this Section both cases (with and without cavities) are studied.

The active patch is fed through a plated hole. Periodic boundaries are used in the simulation to take into account the effect of surrounding patches. The radome is also included in the simulations, this way, the antenna is optimized taking into account the real structure. The radome is made of ABS plastic and has a height of 2 mm. The radome and the parasitic patches are separated by a 2 mm foam sheet. To overcome the inductance caused by the active patch feeding, a printed capacitor is used as shown in Figure 4.4 (a). The variables to be optimized besides this printed capacitor are shown in Figure 2.8. To extend the metal cavity through the substrates, plated holes are used as shown in Figure 4.4 (a)

FIGURE 4.4: Double stacked patch.

The simulated results for one patch in a periodic grid are shown in Figure 4.6. As it is logical, the achieved bandwidth for the patch with metal cavities is larger. The problem comes with the radiation patterns. The horizontal radiation pattern corresponds to the pattern where the antenna is steered, both horizontal radiation patterns have a beam-width larger than $\pm 40^\circ$ (taken at -3 dB). The vertical radiation pattern correspond to the shaped beam radiation pattern. Since the distance among patches in vertical direction is larger, the best option is to have a radiating element that does not radiate power in directions far from broadside to prevent grating lobes. The vertical radiation pattern for the element inside a metal cavity is not very directive, having less than 2 dB difference between broadside and endfire. For this reason, if cavities are used there will be grating lobe problems, thus, cavities will not be used although the matching and coupling will be worse.

FIGURE 4.5: Patch simulation results.

4.3.2 Antenna steering

A row of radiating elements can be steered changing the phase of the feeding ports in the simulation model. Figure 4.6 shows the radiation patterns for five pointing directions. As shown the gain decay for the 40° pointing direction pattern is less than 2 dB as it was expected. The side lobe directivity is increased by 0.5 dB at the 40° pointing direction pattern, making the side lobe level to increase 2.5 dB.

FIGURE 4.6: Radiation pattern of a row pointing at several directions for $\theta = 0^\circ$.

The matchings for every patch in the row for broadside and 40° pointing direction are shown in Figure 4.7. The matchings for broadside are not equals since the coupling with other patches depends on the patches that surround every patch (it is not the same to have 1 patch at the left side and 6 at the right side than to have 4 at the left side and 3 at the right side). For the maximum pointing direction the matching in every port has moved to lower frequencies, even so, the matching in every port is below -8 dB at the highest frequency.

4.3.3 Vertical radiation pattern

The distance from the antenna to the UAV will change according to the elevation as shown in Figure 4.2. Therefore, the distance to the UAV will be larger when the UAV is at broadside direction and smaller for higher elevations. With a shaped beam radiation pattern the antenna has more gain in directions close to broadside and less gain in directions with higher elevations. This way, the free space loss increment is compensated by and increment of antenna directivity. Hence, the shaped beam radiation pattern has

FIGURE 4.7: Matching of every patch in a row.

a small decay in positive elevations and a high decay in negative elevations since these elevations correspond to the ground.

A real cosecant radiation pattern could not be synthesized with only eight elements. Hence, the radiation pattern is modified so that there are not any nulls in positive elevations and the directivity decay is more or less smooth. Figure 4.8 (a) shows the Z diagram of the synthesized radiation pattern. The zeros that correspond to positive elevations close to broadside have been taken out of the module one Z circumference, thus, the radiation pattern has not zeros in these elevations and its decay is smooth. To reduce the level of the side lobes that point to the ground, the zeros of negative elevations close to broadside have been approached to the broadside direction. This will cause an increment in the lobe levels far from broadside but this will not be a problem since the radiating element does not radiate much power in directions far from broadside.

The array feedings for the Z diagram of Figure 4.8 (a) are presented in Figure 4.8 (b). The most important restriction in array feeding is the difference between the highest and the lowest amplitude. This difference is not larger than 10 dB in the proposed feeding, therefore, the feeding network will be easy to design and manufacture.

Figure 4.9 shows the simulated vertical radiation patterns. This simulation is done using an electromagnetic simulator, changing the feeding of each patch port. Cavity and non cavity patches are compared. The radiation pattern for the elements inside a cavity has a higher lobe levels in directions far from broadside as it has been commented in previous section.

FIGURE 4.8: Array feeding for the shaped beam pattern.

FIGURE 4.9: Vertical radiation pattern.

4.4 Vertical network

The vertical network to achieve the feeding for the shaped beam radiation pattern is presented in this Section. The power is divided using regular T-dividers. Seven different power dividers are required as shown in Figure 4.10. There are two choices to feed each output with a different phase. The first one is to add an extra meander line at the end of each port to add a phase shift. The second way is to move the position of each T-divider to change the length of each arm. This way, there is no need of extra meander lines and the total line length is the minimum possible. Hence, the second option is chosen. The divisions for each arm are presented in Table 4.2. The maximum difference

between arms is 5.3 dB. The width of the narrower arm is larger than the etching process minimum width.

FIGURE 4.10: Feeding network.

	$T1$	$T2$	$T3$	$T4$	$Tc1$	$Tc2$	Tc
1st arm (dB)	-1.7	-2.6	-1.5	-2.7	-1.8	-2.7	-1.1
2nd arm (dB)	-4.8	-3.5	-5.4	-3.3	-4.7	-3.3	-6.4

TABLE 4.2: T-divider divisions.

The necessary transmissions for each port are depicted in Table 4.3. The output phases are normalized so that all the values are positive. As it has been commented in Section 4.3.3, the maximum difference among outputs is 10 dB.

	$P1$	$P2$	$P3$	$P4$	$P5$	$P6$	$P7$	$P8$
Phase ($^{\circ}$)	55	0	28	218	11	18	42	75
Module (dB)	-13	-15.9	-11.7	-10.8	-9.8	-5.9	-6.6	-7.2

TABLE 4.3: Port outputs.

Figure 4.11 (a) shows a comparison between the full wave simulation of a row of patches exciting every patch independently and a row of patches plus the network. This way, the influence of the coupling in the network and the etching tolerances are taking into account. As it is shown, the radiation pattern with the network has a small gain decay at $\theta = 25^{\circ}$. This is not a problem since at this angle the UAV would be close and would be able to communicate with an omni-directional antenna mode. Furthermore, the figure shows the radiation pattern for a vertical row, since there are eight of these rows, the total gain directivity will be around 9 dB higher. Another problem of the network could be the differences in the feeding phases with the frequency. Figure 4.11 (b) shows the radiation pattern for the central and extreme frequencies, as it is shown there are not significant differences.

A first demonstrator of this network is shown in the left of Figure 4.12 (a). The measured losses and matching are presented in Figure 4.12. The losses include the losses of the input and output SMP connector that are around 0.5 dB, therefore, the losses of the network are close to 1 dB.

FIGURE 4.11: Vertical radiation pattern.

FIGURE 4.12: Vertical network.

4.5 Active circuit

In this section all the active elements are depicted. The analysis of the RF chain is also performed.

4.5.1 Monolithic devices

The RF chain from the radiating element to the base station is show in Figure 4.13.

FIGURE 4.13: RF chain from antenna row to combination network.

The first element of the RF chain from the WiMAX station is the digital attenuator. This attenuator is used to equalize the output powers in all the rows. It is also used to prevent the power amplifier saturation in the case in which the base station gives more power than needed. The main causes of differences in the output powers of every row are the amplifier drifts and the different attenuation in every state of the phase shifters. The differences among power amplifiers and phase shifter states can be calibrated and stored in a look-up table. However the output powers can also be checked dynamically since the T/R modules include a power detector. One thing that should be taken into account is that digital attenuators do not have the same phase for each attenuation, thus, a trade-off has to be reached among having the appropriate phase in each row or equalizing the power among rows.

The next element of the chain is the digital phase shifter. There are commercial phase shifters of 4 and 6 bits. The price of 6 bit phase shifter is close to three times the price of the 4 bit one and considering that one phase shifter costs more than all the T/R modules, it is advisable to determine the achieved gain difference between both phase shifters. The maximum phase error in the feeding is given by Equation 4.1, where N is the number of bits.

$$\Delta\phi = \frac{\pi}{2^N} \quad (4.1)$$

The losses because of the phase discretization can be calculated rounding the necessary phase feeding for a pointing direction to the available phase steps. Then, the new radiation patterns with the rounded phases can be calculated. Figure 4.14 (a) shows the main beams of several radiation patterns changing the steering direction. As it is shown, there are some differences depending on the phase errors. Figure 4.14 (b) shows the directivity at the desired steering directions, as presented, the directivity decay caused by the phase discretization is less than 0.1 dB, thus, there is no need of using 6 bit phase shifters.

The next elements are the switch and the T/R module. The T/R module is a monolithic device with reception (Rx), transmission (Tx) and antenna pads, thus, it contains an

FIGURE 4.14: Effect of the phase discretization in phase shifters.

internal switch that commutates between reception and transmission. An additional switch is added to commute between the Tx and Rx pad since the attenuator and phase shifter must be common for transmission and reception. The T/R module has four operation modes, two for Tx and two for Rx. The Tx modes are the high linearity mode and the low linearity mode. The high linearity mode has a larger maximum output power but with higher DC consumption than the low linearity mode. The Rx modes are the bypass and the LNA mode. Bypass mode is used in cases where the LNA could be saturated or just to protect the LNA if the received signal has a high power.

The selected components are summarized in Table 4.4.

<i>Component</i>	<i>Model</i>	<i>Company</i>
T/R module	SE5012T	Skyworks
Switch	SKY13286	Skyworks
Phase shifter	MAPS-010145	Macom
Digital attenuator	MAAD-000523	Macom

TABLE 4.4: Selected components.

4.5.2 Power analysis

The maximum and minimum gain in Tx and Rx of the antenna is calculated in Table 4.5. In reception, the chain gain is close to zero, thus, the losses are covered by the LNA. In transmission the gain is more than necessary, thus, depending on the output power of the WiMAX base station the attenuators should increase the attenuation to

avoid saturating the power amplifiers. Since the maximum output power of the power amplifier is 21 dBm, the maximum antenna input power is 10 dBm for the best scenario.

TRANSMISSION/RECEPTION		
	Max	Min
Distri. net. (dB)	-1.5	-1.5
Att (dB)	-2.5	-2.5
PS (dB)	-5	-6.5
SW (dB)	-1.5	-1.8
T/R (dB)	33/12	28/11
Vert. net. (dB)	-1	-1
Chain gain (dB)	21.5/0.5	14.7/-2.3
Directivity (dBi)	23.8	23.8
Gain (dB)	45.3/25.8	40/21.5

TABLE 4.5: Gain study.

4.5.3 Noise analysis

The noise factor of the chain is presented in Table 4.6. The first element in the chain is the vertical network of Section 4.4. The second element is the LNA of the T/R module. The rest of the chain is grouped just in one single element since the rest of devices only have losses. This calculation gives an example of the noise factor of the active chain but not the antenna. To estimate the signal to noise ratio improvement, it should be taken into account that the signals in every array element are summed in phase but the noise are summed out of phase. Therefore there is an improvement of this ratio related to the number of the elements in the array.

<i>Component</i>	<i>Att./Gain (dB)</i>		<i>Noise Factor (dB)</i>	
	Min	Max	Min	Max
Vert. network	1	1	1	1
LNA	12	11	2.5	2.5
Rest	10.5	12.3	10.5	12.3
Total	0.5	-2.3	7.39	8.59
Equivalent Noise Temperature			1300 K	1800K

TABLE 4.6: Equivalent noise temperature calculation.

There are two ways of studying this improvement. The first way is using the circuit antenna representation (Figure 4.15).

All the elements in the chain are grouped in one single box with an amplification and a equivalent temperature. If all the gains and equivalent temperatures are equal, the signal to noise ratio can be calculated as Equation 4.2. Hence, the improvement would be directly proportional to the number of elements in the array. However there are

FIGURE 4.15: Array model

several things that are not taken into account. The first thing is that the signal in all elements are not necessary summed with the same amplitude and phase since there could be some tapering, coupling among elements, polarization losses... thus, it would be more convenient to represent S_t as $S_i \cdot G_{antenna}$. Moreover, we are assuming that the noise at every antenna (T_{ai}) is uncorrelated, but in fact there is some correlation among received noise in every element. This correlation depends on the specific environment where the antenna is placed and cannot be easily determined without measuring it.

$$\left. \begin{aligned} S_t &= \left[\sum_i |s_i| \right]^2 \cdot G \\ N_t &= \left[\sum_i N_{e_i} \right] \cdot G = \left[\sum_i K(T_{e_i} + T_{a_i})B \right] \cdot G \end{aligned} \right\} S_t/N_t \propto \frac{N_{ele}^2}{N_{ele}} = N_{ele} \quad (4.2)$$

For the previous reasons, it would be more convenient to express the improvement in terms of G/T (Equation 4.3).

$$\frac{G_{directive}}{T_{chain} + T_{antenna}} \quad (4.3)$$

To determine the antenna temperature is necessary to integrate the noise weighted by the array radiation pattern (Equation 4.4). Since the antenna is pointed to the horizon the brightness temperature will be very close to Earth temperature (290 K). However, in some occasions, the sun could be just behind the aerial asset. In this cases the antenna temperature could be three orders of magnitude higher.

$$T_{antenna} = \int_{\Omega} Directivity \cdot T_{brightness} d\Omega \quad (4.4)$$

Summarizing, although the noise factor of the RF chain is a bit high, the signal to noise ratio will be reduced in a factor close to the antenna directive gain. For most

occasions, the antenna noise temperature will be close to the Earth temperature but in some scenarios this temperature could increase a lot till the Sun noise temperature. Nevertheless, it should be taken into account that the antenna is for terrestrial communications, thus, the noise requirements are much lower than the requirements for a satellite communication antenna.

4.6 Control and power systems

This section explains the necessary elements to control and power the antenna. It also gives a time estimation for switching among beams.

The antenna is controlled through an I2C bus. Hence, expanders are needed to convert the I2C bus into bits. Each RF chain requires 4 bit for the phase shifter plus 5 bit for the attenuator. Although each digital attenuator has 6 bits, the last one is set to 0 since an attenuation larger than 16 dB is not necessary for the antenna scenarios. Therefore $(4+5) \cdot 8 = 72$ bits are necessary for controlling the phase shifters and attenuators. Five additional bits are necessary; three for selecting the operation mode of the T/R modules (high linearity, low linearity, LNA and bypass) and two more for selecting the switch state (two ports loaded, tx to output port 1 and tx to output port 2). The bits for controlling the T/R module must have a voltage of 3.3 V and the rest should have a voltage close to 5 V. For this reason three expanders are used: two of them of 40 bits and with a feeding voltage of 5 V and one of 8 bits with a feeding voltage of 3.3 V.

The I2C bus supports a speed of 3.5 MB. All the RF components have a commutation time of less than 500 ns. To refresh all the expanders, three instructions are necessary. For the bigger expanders every instruction has a length of 56 bits. For the small one the instruction has a length of 24 bits. Hence, the time to change between pointing directions is close to $50 \mu s$. This time is enough for tracking aerial assets at distance of several kilometers.

The speed of the I2C bus could not be enough to commutate between transmission and reception in a WiMAX frame. If the antenna does not commutates between Tx and Rx fast enough, it will not be possible to establish a link using WiMAX. For this reason, the bits responsible of this commutation are taking out through a parallel bus. Therefore, the control of these bits could be done through the I2C expander or independently, directly from the base station. If the bits are controlled independently, the expander pads where they are connected can be set as inputs so that the expander does not interfere with the control of these bits.

The T/R modules include a power detector that gives a voltage in function of the PA output power. This power is sampled with a I2C controlled AD converter. Since the detector is not linear, the power-voltage curve has to be measured. Figure 4.16 shows the curve fitted from the measured data.

FIGURE 4.16: Measured behavior of the power detector.

4.7 Calibration system

This section depicts the antenna calibration. In a first step a static and semi-dynamic calibration is explained. This calibration is for the case in which the base station does not saturate the power amplifiers. A second calibration is also proposed for the case in which the base station gives more power than needed.

The phase shifters have a different attenuation depending on the phase state and the attenuators change the phase depending on the attenuation value. Moreover the amplifiers could have some drifts in the gain. Therefore, it is necessary to calibrate the antenna to have the best performance in every pointing direction. The variables for calibrating the antenna are the phase shifter and attenuator states ($n1$ and $n2$ in Equation 4.5). In a first step a static calibration is performed with the measured S-parameters of the phase shifters and attenuators. The objective of this static calibration is to obtain the finest attenuator and phase shifter states for every pointing direction without taking into account the amplifier drifts. The objective is not to compensate differences but to maximize a fitness function. Hence, in some occasions a phase or amplitude error would be desirable if the performance is better. This calibration gets close to the optimum point but the final optimal point is reached when the real antenna behavior is sampled. It is done only one time and stored in a look-up table.

$$Row\ Power_i = PhaseShifter_i(n1) \cdot Att_i(n2) \cdot G_i \quad (4.5)$$

Once the first calibration has been established, a semi-dynamical calibration is performed. Using the values from the look-up table, the antenna is steered to a certain direction. Then, the amplitude value in each row is sampled. Taking into account this value the optimization algorithm is re-run. The number of evaluations will be low since the weights will be close to the finest ones. Changing the phase shifter/amplifier states could lead to a different behavior of each amplifier (the gain can vary depending on the input power). Thus, the amplitude is re-sampling in every row and the algorithm is re-run. This process could be repeated several times until some convergence is achieved.

Once the optimal states are found, they do not change unless the antenna conditions are changed (temperature, humidity ...). Hence, the calibration could be done periodically starting with the previous values.

The fitness function is presented in Equation 4.6. Since the variables are discrete (the phase shifter and attenuator states), a genetic algorithm is selected to minimize the fitness function. The main genetic algorithm parameters are shown in Table 4.7. The initial individuals stem from an individual with the chromosomes that correspond to the phase shifter states for the selected pointing direction and for the attenuator states with minimum losses.

$$f_{fitness} = \frac{C_1}{Gain} + \frac{C_2}{SLL} \quad (4.6)$$

Pool size	100
Upper bound	$[16 \cdot \text{ones}(1,8), 32 \cdot \text{ones}(1,8)]$
Lower bound	$\text{ones}(1,16)$
Elite count	10

TABLE 4.7: Genetic algorithm parameters.

The fitness function takes into account the gain at the steering direction and the SLL. The proper value of C_1 and C_2 can be modified depending on the calibration objective. Thus, if the SLL are prioritized, C_2 should be greater than C_1 and vice versa.

Figure 4.17 shows the results for the static calibration (without taking into account the amplifier drifts). Three graphs are shown: the reference that corresponds to the value without calibrating, the calibrated result prioritizing the gain and the calibrated result prioritizing the SLL. The plotted gain is for a lineal array of eight elements without amplifiers. As it is shown the gain improvement is 0.4 dB in the best case. In the case of the SLL the improvement is higher, especially for the highest pointing directions. Thus, the calibration pays off when the SLL is prioritized.

FIGURE 4.17: Static calibration. $\theta = 0^\circ$.

The results after the semi-dynamical calibration are shown in Figure 4.18. In this case, the amplifier behaviors are included. The improvement in gain is more or less the same. As far as the SLL is concerned, they are lower than in the previous case because of the differences in the tapering caused by the amplifiers.

FIGURE 4.18: Calibration including PA behavior. $\theta = 0^\circ$.

Figure 4.19 shows the radiation patterns in the H plane when the antenna is steered to 35° . The cases without calibration and with calibration are compared in simulation (a) and in the measurements (b). As shown the SLL are reduced around 3 dB. The gain in both cases is 0.1 dB higher for the calibrated results.

The phase and amplitude along the row for the case of Figure 4.19 are presented in Figure 4.20. They are compared to the case without calibration. The measured amplifier

FIGURE 4.19: Measured calibration results for a pointing direction of 35° .

drifts are also plotted. As shown, some of the rows have a small phase difference that increments the signal amplitude in these rows since the losses in these states are lower.

FIGURE 4.20: Phases and modules in every row for a pointing direction of 35° .

The calibration results show that the radiation pattern can be improved allowing a small phase error in the phase shifters. This way, the losses in the phase shifter can be minimized with small phase errors. As far as the attenuators are concerned, in most of the cases the algorithm gives results with all the attenuators set to the minimum attenuation. Hence, it would not be necessary to have a digital attenuator in each row. To limit the power from the base station to the antenna, it would be only necessary one attenuator at the beginning of the distribution network.

However, if the base station gives more power than the maximum to not saturate the power amplifiers, the different losses in each row can be equalized. This way, the total radiated power would be the maximum possible. Figure 4.21 shows the radiated power

normalized to the maximum when the antenna receives more power than necessary. Two cases are presented. In the first case the genetic optimization has been applied. The second case corresponds to an antenna with only one attenuator at the beginning. As shown if the row power is equalized to the maximum value, the antenna radiates almost 0.4 dB more on average than the case without equalization. The values at broadside are equal since at broadside all the rows must have the same phase and there is nothing to be equalized. The graph for the optimized values has less peaks and it is more uniform, very similar to the patch radiation pattern.

Compared to the previous case, the only change that has to be done in the fitness function is to set a penalty factor to the individuals that saturate one of the amplifiers.

FIGURE 4.21: Normalized radiated power. $\phi = 0^\circ$.

4.8 Antenna prototype

Figure 4.22 shows the first antenna prototype. Figure 4.22 (a) shows the active and parasitic patches. These patches are not seen in the Figure 4.22 (c) since they are covered by the ABS radome. Figure 4.22 (b) shows a detail of the RF components. The antenna is mounted in a camera tripod so it can be pointed easily.

Figure 4.23 shows the measured radiation patterns in the E and H plane. The diagrams in the E-plane correspond to the vertical direction that implements a diagram that points to broadside with a smooth gain decay in positive θ and with an abrupt decay in negative θ . The gain difference between the Tx and Rx is just the gain difference of the PA and LNA. The diagrams in the H-plane correspond to the horizontal plane where the antenna is steered. As it is shown, the gain differences when pointing are less than 2 dB.

Figure 4.24 (a) shows the measured matching in transmission. Two matchings are plotted: the matching when the attenuators are at maximum level and the matching

FIGURE 4.22: Prototype.

with the attenuators at minimum level and with the high linearity mode activated. The matching has moved a bit to lower frequencies. The main causes of this displacement are the digital attenuators and the Wilkinson dividers. The upper frequency is in the limit of the working band of the digital attenuators, therefore the attenuator matching is worse in the upper frequencies. Although the matching of the attenuators is not better than -10 dB, the antenna matching is better than -15 dB at the central frequencies. This is caused by the Wilkinson dividers that are used to distribute the power among rows. These Wilkinson dividers absorb part of the reflected power in the attenuators.

Figure 4.24 (b) depicts the Tx and Rx measured gain. The peak in the Tx gain is mainly

FIGURE 4.23: Measured radiation patterns.

caused by the power amplifiers that have a ripple of 3.5 dB in the band. Apart from this peak, the rest of the ripple is caused by two main reasons: the worse or better matching and the different losses of each monolithic with the frequency.

FIGURE 4.24: Measured gain and matching.

4.9 Conclusions and future lines

The design and manufacturing of an antenna for large range UAV communications has been presented. The antenna uses a commercial WiMAX station as a transmitter/receiver. The usual range is improved increasing the antenna directivity. Since the antenna directivity is increased, the beam width is reduced. Thus, to connect with mobile assets,

the antenna must have electronic steering. A mechanical steering has been discarded since it does not allow to connect with several assets with the necessary speed.

The design and measurement show a good agreement except for the upper design band. The main cause of the malfunctioning in the upper band is that the matching has moved to lower frequencies. The monolithic behavior is also better in the lower bands. Nevertheless, this is not a problem since the WiMAX station works in a wider bandwidth and can cover the bandwidth in which the antenna has a better performance.

The first future line is to include an additional LNA after the T/R module. The LNA of the T/R module has a gain that only compensates the phase shifter, commutator and attenuator losses. Thus, the noise figure can be improved if a second LNA is used. This way, the losses after the two LNAs would not be significant for the noise figure calculation. Nevertheless, the noise in the antenna for terrestrial communications is not as important as in satellite communications.

The antenna uses phase shifters to steer the antenna but a multi-beam network could be used (see [63]). The losses in this network would be even lower than the losses in the phase shifters. Moreover all the beams would be available all the time. If only one transmitter/receiver is used the beams could be selected using switches. Therefore the commutation time would be much faster. Moreover, several transmitters/receivers can be used, allowing to have several simultaneous beams. Another advantage of the use of this network is that the losses to each row for some pointing direction would be more or less the same, hence, the calibration process is not that necessary. The main disadvantage is that the network would take up more space and a more complicated stack-up would be needed.

Chapter 5

Slotted waveguide array in SIW technology at Ku band

5.1 Introduction

This chapter presents the design and manufacturing of a small size Substrate Integrated Waveguide (SIW) array antenna at Ku band with circular polarization. The antenna consists in a eight standing-wave slotted array and a polarizer to obtain circular polarization. There is some tapering among slots to reduce the SLL. To distribute the power to each slotted waveguide, a parallel distribution network is used. This network is done in a different layer than the radiating elements to reduce the antenna size. To couple the power between layers a coupling H-slot is used.

Substrate Integrate Waveguide technology consists of a drilled substrate that imitates a waveguide. This way, the conductor losses are much smaller than in traditional printed technologies such as microstrip. The main advantages are: the easy manufacturing process, the low cost compared to other technologies in high frequencies and the low profile. The main disadvantage is that the losses are generally higher than in regular waveguide technology for the substrate dielectric losses. Another advantage is the technology versatility. It is very easy to integrate active components into the SIW technology. Therefore low noise amplifiers or phase shifters could be easily added at the beginning of each slotted array, improving the antenna behavior.

The most common radiating element used in SIW antennas is the slotted waveguide. There are two main kinds of slotted waveguide arrays. The first kind is the traveling wave array [65, 66]. The problem with this arrays is that they require a large number of radiating elements to radiate most of the power, for this reason, they are difficult to

use if the array size is not very large. The second kind of slotted waveguide array is the standing-wave array. In this case all the slots are at resonance and should be separated usually $\lambda_g/2$. The number of elements in the array can be smaller, making it possible to design small size arrays.

To obtain circular polarization in a slotted waveguide, 45° inclined slots separated $\lambda_g/4$ are usually used (see [67]). They are also placed in pairs to reduce the cross polarization (see [68]). These arrays are generally based on traveling-wave array, therefore the number of elements have to be high. Another solution is to use longitudinal slots that radiate with lineal polarization plus a polarizer above the radiating elements. Example of polarizers can be found in [69, 70]. The solution chosen in this chapter is a meander line polarizer because of its simplicity and polarization purity.

Section 5.2 explains the SIW design process and makes a comparison among suitable commercial materials that can be used for SIW, contrasting good performance and easy manufacturing process. Antenna structure is depicted in Section 5.3. The transition between the input SMA connector and the SIW is presented in Section 5.4 . The simulation and measurement of half of the power distribution network are presented in Section 5.5 . The H-slot used to couple power between two SIW in different layers is presented in Section 5.6 . The design method for the slotted waveguide is explained in Section 5.7 . In Section 5.8, some problems related to the manufacturing process are also depicted. Section 5.9 shows the design of the meander line polarizer. The final prototype and its measurements are presented in Section 5.10 .

5.2 SIW technology

In this section, the SIW is introduced. First subsection explains how to design and simulate SIW lines. In the second subsection, a comparison among suitable commercial substrates for SIW is established.

5.2.1 Simulation procedure

The first step when designing a SIW structure is to design an equivalent structure in waveguide technology. The structure should be optimized as much as possible since the simulation time for regular waveguide is much smaller than for a SIW structure.

Once the structure has been optimized the waveguide has to be converted into SIW. The main parameters for the SIW can be seen in Figure 5.1. They are: the via diameter

(D), the distance among vias (b) and the distance between drilled walls or SIW width (W).

FIGURE 5.1: SIW main parameters.

There is a trade off in via diameter between an easy manufacturing (bigger vias) and better performance (smaller vias). The distance between vias should be smaller than $\lambda_g/5$ to prevent leakage. Finally, it can be established an equivalence between the SIW width and the width of a regular waveguide with the same propagation constant than the SIW. This is usually called effective width (W_{eff}). There are multiple papers explaining how to calculate this width but the most used one is presented in Equation 5.1.

$$W_{eff} = W - \frac{D^2}{0.95 \cdot b} \quad (5.1)$$

This way, if a SIW width has been chosen, the regular waveguide simulations have to be done using the equivalence width. More details on how to calculate the propagation constant and more restrictions can be found in [23].

Once the structure has been optimized in regular technology, it has to be drawn using a CAD software with the new distance between walls. In most of the cases there is not any difference between the regular waveguide simulation and the SIW. But in some of the cases there are some differences and the SIW structure should be optimized a little bit. In these cases, the SIW structure can be imported from the CAD software into the full wave simulator. Some vias can be moved to improve the behavior of the structure.

To simulate the SIW structure with a FIT or FDTD simulator the mesh ratio has to be high, otherwise the number of cells would be huge. Another thing that has to be changed in some simulations is the stability factor to make the simulation stable. In most cases, a stability factor of 0.8 is more than enough. Sometimes the simulation could become instable in times far from the excitation, the best option in these cases is to stop the simulation some time after the signal has arrived to the output port. This time has to be calibrated but the simulations using this method agree in most of the cases with measurements as it is shown in this chapter. FIT/FDTD simulators requires the

structure to be identical three longitudinal mesh steps from the port to apply the port conformation. For this reason, SIW-SMA transitions have to be used to simulate the SIW structure since the SIW is not identical in the propagation direction. The followed simulation procedure is shown in Figure 5.2.

FIGURE 5.2: SIW simulation procedure.

5.2.2 Choice of dielectric material

There are several suitable substrates for the SIW technology. Substrates based on PTFE (Teflon) have the lower losses. The problem with the regular PTFE substrates is that they are very difficult to metallize. This sets a limit in how small the plated holes can be. Moreover, it is very difficult and expensive to make blind or buried plated holes in a PTFE substrate.

On the other hand, substrates based on hydrocarbon/ceramic are becoming more and more popular since they are very easy to metalized and cheap to manufacture. One example of these substrates is RO4350B. The disadvantages of these substrates are that the permittivity is a bit higher, reducing the radiation efficiency in antennas, and the dielectric losses are higher.

As a trade-off between regular PTFE substrates and hydrocarbon/ceramic substrates, a ceramic filled PTFE substrate can be used like RO3035. This substrate has lower losses than RO4350B and higher losses than PTFE substrates but with a similar manufacturing process than standard FR4 substrates. A comparison of the losses for a waveguide of 100 mm length filled with the materials commented before is shown in Figure 5.3. For the PTFE simulation, the waveguide width has been increased to support the TE₁₀ mode at the simulated frequencies. Since the manufacturing process is similar to FR4, more complex SIW structures can be achieved with the mentioned Rogers substrates than with PTFE substrates. The figure also shows the isolated copper losses in the Rogers substrates. It proves that the conductor losses are much lower than the dielectric losses.

FIGURE 5.3: Losses depending on the substrate.

Figure 5.4 shows the losses for a SIW filled with RO4350B changing the substrate height. As it is logical the losses increase when the height is decreased. For this reason, the maximum available substrate height is used, for RO4350B and RO3035 the maximum substrate height is 1.524 mm.

FIGURE 5.4: Losses depending on the substrate height.

To validate all the antenna parts, all the prototypes have been manufactured using RO4350B since the substrate price is lower than RO3035. For the final prototype, RO3035 is used. Both substrates have a similar dielectric constant, thus, minimal changes in the design have to be done when changing between both substrates.

The main parameters of the SIW are summarized in Table 5.1.

For the chosen width, the dominant mode (TE₀₁ mode) has a cut off frequency of 9.2 GHz. The cut off frequency for the second mode is 19.3 GHz. Therefore, the SIW is monomode at the center frequency of 12.5 GHz.

Substrates	Ro4350B and Ro3035
Permittivity	~ 3.6
Height	1.524 mm
Via diameter (D)	0.4 mm
Via distance (b)	0.55 mm
Waveguide width (W)	8.95 mm
Effective width (Weff)	8.65 mm

TABLE 5.1: Main parameters of the SIW.

5.3 Antenna structure

The array presented in this chapter consists of eight slotted waveguides fed by a parallel distribution network. Each slotted waveguide has five slots. The power distribution network is formed by T-dividers [29] with an inductive wall and a capacitive iris. The distribution network and the slotted waveguides are usually made in the same layer [69]. To reduce the antenna size the power distribution network can be placed in a different layer below the slotted waveguides. Thus, the antenna size is just the size of the slotted waveguides. To couple the two layers, a coupling slot is used. The most common way to couple power between two waveguides is an inclined slot (see [71]). H-slot is used as coupling slot because the achieved matching and bandwidth are better.

The antenna scheme is shown in Figure 5.5. The size is 12x10x1.6 cm .

FIGURE 5.5: Antenna scheme.

5.4 Transitions

The input/output of the antenna has to be a conventional SMA connector. There are two main ways to design a transition between SIW and SMA connector. The most common way is to use a small microstrip network with a SMA soldered in one of the edges and then a transition between microstrip and SIW. The problem with this transition is that the antenna input has to be in one of the edges, increasing the antenna size. Moreover the losses in the microstrip plus the microstrip-SIW transition could be high.

The second way is to design a vertical transition with a SMA launcher that penetrates into the SIW. The advantages of this transition are that there is no need of an additional microstrip line and the transition can be placed in the center.

Both ways have been manufactured and tested. Subsection 5.4.1 shows the transition from a microstrip network to a SIW and Subsection 5.4.2 shows the transition with the SMA launcher. Eventually the transition with the SMA launcher has been used because its better performance.

5.4.1 Microstrip to SIW transition

The transition from a waveguide to a microstrip is shown in Figure 5.6 (a). The discontinuity between the waveguide and microstrip has been widely studied (see [72, 73]). Usually this step is modeled like an extra waveguide with a different impedance placed between the microstrip line and the waveguide. To cancel the mismatch, a tapered microstrip is used. The parameters to be optimized are the length and width of the tapered microstrip line.

The measurements are presented in Figure 5.6 (b). The matching has been moved to lower frequencies. Even though the matching is below -20 dB in a bandwidth of 0.6 GHz. The transitions can be re-optimized to move the matching to upper frequencies.

More complex transitions can be designed. For instance, the part where the SMA connector is soldered to the microstrip can be shielded with a custom metal piece or a gap can be done in the SIW to improve the matching.

5.4.2 SMA launcher transition

The scheme of the transition with a SMA launcher is shown in Figure 5.7. A similar transition can be found in [74].

FIGURE 5.6: Microstrip to SIW transition.

FIGURE 5.7: Transition scheme.

The inner conductor of the SMA penetrates in the SIW substrate and it can be soldered to the top of the SIW. The distance from the coaxial line to one of the SIW walls is close to $\lambda_g/4$, therefore the short circuit that represents the wall is transformed into an open circuit. If the other side of SIW is well matched the input coaxial line sees an open circuit in parallel with a matched load, thus, the input coaxial line sees only a matched load.

The best way to match the transition is to leave a specific length of the inner coaxial line inside the SIW as it is usually done in regular waveguide technology. The problem is that since the substrate height is low, the accuracy when cutting the inner SMA conductor is critical. Thus, it is more difficult to achieve a good matching.

Therefore, to avoid accuracy problems, the inner coaxial is prolonged and soldered to the SIW top layer. There are two ways to match the transition. The first one is a wider piece of SIW at the beginning of the transition (W_a in Figure 5.7). The second one is

a capacitive circular gap concentric with the inner conductor that is soldered to the top of the SIW.

Figure 5.8 shows a comparison between a simulation of a transition with and without gap on the top. The bandwidth of the transition with gap is a bit higher. The losses of the transition without gap on the top are larger because the width and length of the matching waveguide is larger and spurious modes can be generated.

FIGURE 5.8: Simulation of a SMA-SIW transition.

The measurements versus simulations of a SIW through with two transitions are presented in Figure 5.9. Although the simulations for a transition with gap are better than the simulations of the transition without gap, the measurements show that the behavior of the transition without gap is better. The main cause of the malfunctioning is the accuracy in the gap etching. The size of the gap is close to 0.1 mm, for this reason if the accuracy in the etching process is not that high it is better to use a transition without gap even though the losses are a bit higher.

In Figure 5.10 is presented the measured versus simulated losses for the transition without gap. There is a good agreement between the simulations and the measurements. The presented losses include the losses of the two SMA connectors, two transitions and a 40 mm SIW line.

5.5 Power distribution network

To divide the power among each row of radiating elements, T dividers are used. Several techniques can be used to math T dividers. The most basic one is a central pin (Figure 5.11 (a)). The most complex is the divider matched with an inductive wall and a capacitive iris at the beginning of the divider (Figure 5.11 (c)).

FIGURE 5.9: Measured matching of a SIW through.

FIGURE 5.10: Measured losses of a SIW through with transitions without gap.

The simulated results for the dividers of Figure 5.11 are shown in Figure 5.12. As it is shown, the matching for the divider with an inductive wall and an iris are much better than the other ones.

The size of the inductive wall and the position of the first plated holes are optimized in every T divider depending on the position of the divider in the distribution network. The measured matching of a 1 to 4 distribution network is shown in Figure 5.13 (b). The matching is below -12 dB from 12 to 13.5 GHz. The losses of the 1 to 4 distribution network are presented in Figure 5.13 (a). The losses are around 1.25 dB for the designing frequency (12.5 GHz). In the measured losses are also included the losses of two SMA-SIW transitions and two SMA launchers. The losses are very similar to the losses of a 40 mm through presented in Figure 5.10. This means that most of the losses are in the transitions.

FIGURE 5.11: T Power dividers.

FIGURE 5.12: T Power divider matching comparison.

5.6 Coupling slot

The distribution network and the radiating elements could be done in one substrate. However the size of the antenna can be reduced if the distribution network and the radiating elements are stacked. Thus, the area of the antenna is just the area of the radiating elements. To couple the power from the distribution network to the radiating elements an H-slot is used.

The point of using this slot is that it resonates at two frequencies, thus, the achieved bandwidth is higher. Figure 5.14 (b) shows the simulated matching of an H slot and the matching of the same H slot eliminating the transversal central arm. It shows that the longitudinal arms resonate at the higher frequency and when the central arm is added a new resonant frequency appears. The resonant lengths are depicted in Figure 5.14 (a), they are close to $\lambda_g/2$ at the resonant frequencies.

The measurements versus simulations of a through with an H-slot are show in Figure 5.15. The agreement between simulations and measurements is high.

FIGURE 5.13: Measurements of half of the power distribution network.

FIGURE 5.14: H slot resonances.

5.7 Radiating element

The selected radiating element is a longitudinal slot. These slots have been widely studied. The main reference for studying them can be found in [75]. The difference in this case is that the waveguide is filled with a dielectric of 3.6 permittivity .

The first thing to do is to characterize a single slot. Every slot can be characterized as a single parallel admittance, if the slot is at resonance this admittance will be real. Therefore, there are two variables in each slot, the length and the offset to the waveguide center. The offset changes the amount of power that is radiated by the slot. This power is represented by the real part of the admittance.

FIGURE 5.15: Through with an H slot.

To characterize these parameters the model of the Figure 5.16 is used. The distance from the slot center to each port is $\lambda_g/2$. To take into account the effect of the coupling with the slots in the next waveguides, periodic boundaries are used. Therefore, the input impedance seen in one of the ports will be the impedance of the other port (waveguide characteristic impedance) in parallel with the slot impedance since the distance between the slot and each port is $\lambda_g/2$. This is only true at the center frequency where $\lambda_g/2$ has been taken.

FIGURE 5.16: Model for characterize a slot.

Taking the S_{11} at the input port, the slot admittance can be isolated (Eq 5.2).

$$\frac{Y_s}{Y_0} = \frac{-2S_{11}}{1 + S_{11}} \quad (5.2)$$

The admittance depending on the offset is shown in Figure 5.17 (a). The slot length has been changed until resonance for every offset. The graph demonstrates that the admittance is higher when the offset increases, meaning that it radiates more power. Since the slot is at resonance, the admittance imaginary part is close to zero. This

is true for offsets below 1 mm, for offsets above this value, the imaginary part cannot be neglected. This is not a problem since the offset value will be always below 1 mm because the sum of all slot real admittances should be equal to one. The resonant length depending on the offset is shown in figure 5.17 (b).

FIGURE 5.17: Single slot characterization.

Usually the slots are placed every $\lambda_g/2$ in the waveguide with alternatively offset, this way, every slot radiates with the same phase at the central frequency. One of the problems in waveguide filled with dielectric is the coupling between slots. This coupling is usually represented with an extra reactance in the slot. In waveguides filled with air, this is solved changing the slot length to compensate this extra reactance. The problem with SIW is that now the dielectric has a permittivity of 3.66, thus, the slots are closer. Therefore, the extra reactance cannot be easily canceled by changing the slot length.

To solve this problem, half of the slots have been eliminated. In this way, the distance among slots is now λ_g . There is not any problem with grating lobes since λ_g is equal to $0.75\lambda_0$.

The slotted waveguide is finished with a wall at a distance of $\lambda_g/4$ from the last slot. Hence, the wall that represents a short circuit is transformed into an open circuit at a distance of $\lambda_g/4$. Since every slot is separated by a distance of λ_g , at the beginning of the waveguide the input admittance is the sum of all slot admittances (eq. 5.3). To reduce the side lobe level a tapering in the feeding of each slot can be done. This way the central slot has to radiate more power, having more offset, and the edge slots have to radiate less power, having less offset. If the feeding distribution has been established ($\sum_i a_i$ in eq. 5.3), there is only one variable to be found, the reference slot impedance (Y_x in eq. 5.3).

$$Y_0 = \sum_i Y_i = Y_x \sum_i a_i \quad (5.3)$$

The offset for the reference slot impedance can be determined in Figure 5.17 (a). Then, the resonant length can be obtained from Figure 5.17 (b). In most of the cases, the offsets will be below 0.5 mm, thus, the resonance length for every slot will be almost the same.

Figure 5.18 shows the matching of the final waveguide with five slots. The matching has been calculated with a full wave simulator and concatenating full wave simulations of one single slot. The results are very similar, showing that the coupling among slots in the longitudinal direction can be neglected. Therefore, the design can be done using the graphs from Figure 5.17.

FIGURE 5.18: Simulated matching of 5 slots.

The simulated radiation pattern for five slots with periodic boundaries is shown in Figure 5.19. The side lobe level of the radiation pattern with tapering is lower as it is logical. It is also included the radiation pattern of an array of 10 elements spaced $\lambda_g/2$ with the previous tapering among each pair of slots. This array has been matched changing a bit the length of the slots. The radiation pattern shows that there is not any advantage in using a 10 element array because the SLL is worse due to the coupling.

A comparison among the radiated signals in every slot can be calculated with a full wave simulation. To obtain this comparison, a voltage monitor must be placed between longitudinal centers in the slots. Since the electric field is almost constant in a slot cross section, the voltage difference is directly proportional to the electric field complex value (Equation 5.4). Thus, the voltage amplitude gives an idea of the radiated power in every slot. The voltage phase should be the same for every slot if they are in phase.

FIGURE 5.19: Radiation pattern of the designed slotted arrays.

The resonant frequency of the slot can also be determined by checking where is the maximum amplitude of the voltage.

$$V = \int \vec{E} d\vec{l} = E \cdot W_{slot} \quad (5.4)$$

Figure 5.20 (a) shows the voltage difference in every slot for the previous 5 slot array. As shown, the resonance of the fifth slot has moved to higher frequencies. Moreover, the amplitude difference between the fourth and second slot is too high. The main cause of this malfunction is that the short end is very close to the last slot. Figure 5.20 (b) shows the voltage difference in every slot for the same array presented before but increasing the distance to the short in λ_g . This way, the short is still represented as an open circuit at $5\lambda_g/4$ but there is more distance between the last slot and the short circuit. The voltage amplitudes shows that now all the slots are more or less at resonance at the central frequency. The tapering is also well done since the slots almost radiate symmetrically to the longitudinal array center.

The radiation patterns for the 5 slot arrays of Figure 5.20 are presented in Figure 5.21. The radiation pattern for the case with smaller distance is asymmetric and with higher SLL because the tapering is not symmetric and because all the slots are not at resonance at the center frequency. On the other hand, the radiation pattern for the array with larger distance between last slot and the short circuit is the expected radiation pattern for a five element array (symmetric and with lower SLL).

Another option to solve the problem of the last slot is to change the length of this slot. Hence, since the resonance of the last slot in Figure 5.20 (a) is in a higher frequency, the length must increase. The slot length is increased from 8.3 mm to 8.45 mm, obtaining

FIGURE 5.20: Voltage module in every slot for a distance between last slot and short circuit end of $\lambda_g/4$ (a) and $5\lambda_g/4$ (b).

FIGURE 5.21: 5 slot radiation pattern changing the distance from the last slot to the short circuit.

the results show in Figure 5.22. The results are almost the same as the previous ones but without the need of increasing the antenna size.

The diagrams presented in this section are for the H-plane. The diagrams for the E-plane are not presented since they are the typical radiation patterns for an array of eight elements in phase.

5.8 Possible manufacturing problems

Once the design is done, some verifications have to be done regarding the implementation with the SIW technology. The first one is the copper thickness. Usually commercial

FIGURE 5.22: Slot voltages (a) and radiation pattern (b) optimizing last slot length.

substrates have 35 or 17 μm (1 or 1/2 oz.) of copper thickness but it should be taken into account that because of the drill metalization process, this height is increased until 70 μm (2 oz.). In Figure 5.23 is shown the simulation of the slotted waveguide changing the copper thickness. As it is shown the resonant frequency is moved to higher frequencies if the new copper thickness is not taking into account. Another problem related with the SIW manufacturing process is the via diameter and the separation among vias. To make easier the manufacturing process the best option is to have the maximum possible via diameter and space. The problem is that when the via space is increased, the leakage is also increased. Figure 5.23 also shows the simulation of a SIW slotted waveguide changing the via diameter from 0.4 mm to 0.5 mm. The via space has also changed from 0.55 mm to 0.86 mm to have the same effective width. As shown, the behavior of the simulation with smaller vias is more similar to the expected result.

FIGURE 5.23: Matching comparison of possible manufacturing problems.

5.9 Polarizer

The polarization radiated by the slots presented in Section 5.7 is linear. To obtain a circular polarization, a four layer meander line polarizer is used. The polarizer has an inductive behavior if the wave has a linear polarization parallel to the meanders and a capacitive behavior if the wave polarization is perpendicular to the meanders. The capacitance is controlled with the distance among meanders whereas the inductor can be controlled with the width of the meander. Each polarizer layer is printed in a RO4350B substrate with a height of 0.254 mm. Foam sheets of 2 mm are used to separate each layer.

To optimize the polarizer, a FEM simulator with unit cell boundaries has been used. Two ports are added in Zmin and Zmax. Two modes are excited in zmin as shown in Figure 5.24. There are three goals for the optimization process. The first one is the reflected power of the two modes that should be as small as possible. The second one is the transferred power from one mode to the other one which also must be small. The results of these two goals after optimization are shown in figure 5.25 (a). The third and most important goal is the phase difference between modes in zmax which has to be 90° . This phase difference is shown in figure 5.25 (b).

FIGURE 5.24: Model for optimize the polarizer.

Usually these polarizers are meant to be placed at certain distant from the antenna. The distance between polarizer and antenna should be as smaller as possible to keep the low profile. For this reason, a simulation with a slot plus a polarizer is performed as shown in Figure 5.26. The slot is inclined 45° respect to the polarizer so that the transversal field can be decomposed in the two necessary components in the polarizer to obtain a circular polarization. FIT periodic boundaries cannot be used in the simulation since the structure is not homogeneous in the boundaries. Thus, open PML boundaries are set.

FIGURE 5.25: Simulation of a polarizer unit cell.

FIGURE 5.26: Model for the polarizer plus one slot.

The simulated axial ratio is shown in Figure 5.27. As it is observed, the axial ratio is better when the distance between antenna and polarizer is increased. However to keep the antenna low profile with an acceptable axial ratio, a distance of 4 mm has been chosen.

FIGURE 5.27: Axial ratio for the polarizer changing the distance between polarizer and antenna.

5.10 Final prototype

The final SIW prototype is manufactured in R03035. The polarizer is manufactured in RO4350B since the losses because of the dielectric in the polarizer are low. Figure 5.28 shows the two SIW layers plus the meander polarizer. The meander polarizer is screwed with nylon bolts so that they do not modify the radiation pattern. The screws used to press the two SIW layers are metal screws because they press stronger.

FIGURE 5.28: Final prototype.

The matching for the case with polarizer and for the case without polarizer are shown in Figure 5.29 (a). They are almost the same, showing that the polarizer does not reflect any power if the wave has a perpendicular incidence. The axial ratio and the gain are

shown in Figure 5.29 (b). The axial ratio is below 0.2 dB at the frequencies where the antenna is well matched. In other frequencies, the antenna does not radiate to broadside because of the serial feeding, therefore, the wave does not have the normal incidence to the polarizer and the behavior is not the optimal one. The gain is close to 18.5 dB at the designing frequencies. The simulated antenna directivity is 21.5 dBi, hence, the antenna losses are around 3 dB.

FIGURE 5.29: Final prototype measurements.

The measured radiation patterns are shown in Figure 5.30. The E-plane corresponds to $\phi = 0^\circ$. The SLL is around 12.5 dB. This SLL is higher than expected because the etching of the last slot is not well done since this slot is very close to the via wall that represents the short circuit. For this reason, the last slot length is smaller than expected and it does not radiate the power it should do. One solution would be to increase the distance from the last slot to the via wall to $5\lambda_g/4$.

The photograph of one of the last slots is shown in Figure 5.31. As it is shown, the slot end does not have the proper width and it is inclined. The main cause is the presence of the via wall that is very close to the slot.

5.11 Conclusions and future lines

A small size SIW antenna has been presented. The results show a good agreement with the simulations taking into account the available manufacturing process. The results could be improved if the antenna is manufactured with better media.

The first future line is related to the inclusion of active components such as low noise amplifiers. They can be located in each of the outputs of the distribution network before

FIGURE 5.30: Final prototype radiation patterns.

FIGURE 5.31: Etching of one of the last slots.

the H-slot. They can be placed in a small microstrip network with two transitions between SIW and microstrip.

The antenna bandwidth is fixed by the slotted waveguides. The other elements work in a wider bandwidth, thus to increase the bandwidth the slotted waveguides should be changed. Chapter 6 describes a SIW antenna in which the bandwidth has been increased by modifying the radiating elements and the feeding among them.

The SIW manufacturing technology can be improved. Instead of plated drill, plated slots can be used, making easier the metalization process. Moreover the entire SIW structure can be cut in a substrate and then the edges can be plated. This way, the possible leakage would be eliminated. It would be also easier to check the good metalization at the borders. To stack the two layers, the H-slot could be also metalized. Figure 5.32 shows how the process would be. A H-slot of certain height is milled from a substrate removing also the copper in the top and bottom of the substrate. Then, the H-slot is placed between the two SIW layers. The layers should be pressed and metalized. The etching of the H-slot in the two SIW should be done before the metalization process.

This way, the resulting structure would be more compact and the screws would not be necessary. Finally, the etching of the top and bottom layer would be done. This etching can be done at the end of the process, protecting the edges so they are not de-metalized. This etching can be also done before the metalization process. In this case the top and bottom layer should be protected during the metalization process so that the etching is not metalized.

FIGURE 5.32: New manufacturing process scheme.

Chapter 6

Low profile broadband antenna at ka band

6.1 Introduction

The purpose of this chapter is to describe the design and manufacturing issues of a low profile array antenna for the upper ka band (29-31 GHz). The antenna has a directivity close to 32 dBi and linear polarization. There are several examples of low profile antennas at ka band in the literature, one of them can be found in [76]. Generally most of the low profile and low cost arrays that can cover the ka band are made in microstrip technology. The problem at these frequencies is that besides the high conductor losses, microstrip technology also presents high radiating losses.

For this reason and as it has been established in Chapter 5, SIW technology is a good choice for cheap, low profile and relatively low loss antennas. The usual problem is the small achieved bandwidth in large array antennas. Although some designs (see [77, 78]) exhibit a large matching bandwidth, the pointing direction changes with the frequency, reducing the bandwidth to hundred of Megahertz. Since the waveguide width is usually close to $\lambda_g/2$ it is very difficult to feed all the elements with a parallel distribution. For this reason, a mixed serial-parallel network is used. In this way, a trade-off between easy design process (more elements in series) and effective bandwidth (less elements in series) has to be reached.

The use of a serial distribution makes easier the antenna design but reduces the bandwidth since the elements in the antenna only have the same phase at the central frequency. Usually, the more elements are placed in series, the less bandwidth is obtained. One solution to increase the bandwidth is to have the lowest number of serial elements.

If the serial elements are fed with a central feeding, the miss-pointing losses can be reduced. The central feeding technique is depicted in Section 6.2.

The usual radiating element used in waveguide antennas is a longitudinal slot. The bandwidth of a standard longitudinal slot does not cover the whole ka band. Thus, the radiating element has to be less resonant. The antenna uses patches that are coupled to the waveguide through an aperture in the wider waveguide face. These patches are shown in Section 6.3.

The distribution network proposed in Chapter 5 has the necessary bandwidth so that similar T-dividers and H-coupling slots are used for the parallel distribution network. The only difference is that the antenna has some tapering among subarrays so that the T-dividers are not symmetric.

The antenna input is done through an standard WR38 waveguide. Thus, a transition between the SIW and this waveguide is designed. This transition is designed as an E-plane divider and it is explained in Section 6.4

The substrate used in the first boards is Ro4350B. The final prototype is manufactured in R04003C because its lower dielectric losses.

The general antenna scheme is shown in Figure 6.1. It has two main parts: the subarrays and the power distribution network. The subarrays has 4x2 radiating elements that are fed in series. The distance among elements in one of the directions is fixed by λ_g whereas in the other direction the distance can be chosen independently. The subarrays are fed through an H-slot that works also as a first divider. The parallel distribution network is in the lower substrate and is composed by asymmetrical T-dividers. The arm ratio of these dividers is chosen to reduce the side lobe level of the antenna (see Section 6.6).

6.2 Central feeding technique

This section explains the radiation patterns that can be obtained feeding the serial elements from the center instead of one of the edges.

In a serial feeding, the distance among elements is usually λ_g so that all the elements are in phase. The distance could be also $\lambda_g/2$ if the slots are carved symmetrically to the waveguide center since this symmetry adds 180° phase shift between adjacent slots (because of the H longitudinal component). As it has been explained in the Introduction, the radiating element is not a regular slot because of its low bandwidth. Hence, the distance among radiating elements in the serial feeding is λ_g .

FIGURE 6.1: AntennaScheme.

The usual problem with the serial distributions is that all the elements are in phase only at the center frequency. In the extreme frequencies there is a constant phase shift among elements. This makes the antenna to point to a different direction. This effect could be use to steer the antenna changing the frequency but it is not very useful if the antenna has to point always to the same direction. In Figure 6.2 (a) is shown the different array factors for a 16 serial fed elements. As shown, at central frequency the antenna points to broadside, but in the upper and lower frequency the array factor has a null in broadside direction.

To fix this problem the elements can be fed through the center instead through one end. This way in the extreme frequencies half of the antenna will point to a positive θ and the other half to negative θ , the combination of both array factors does not have a null in the center as shown in Figure 6.2 (b).

If the number of elements in series is high, the radiation pattern of each half is very directive and the SLL increases. To reduce the directivity of each half the number of elements in series can be reduced, this way, there will be an array of serial elements with central feeding technique. Figure 6.3 shows the radiation pattern for 16 elements changing the number of elements in series with central feeding. As shown, when the number of elements in series is high the SLL increases and the directivity at broadside direction decreases. For this reason the chosen number of elements in series for the first prototype is two.

FIGURE 6.2: 16 element radiation pattern feeding the elements from the end (a) or from the center (b).

FIGURE 6.3: 16 element radiation pattern at lower frequency changing the number of elements in series.

The calculated array factor can be decomposed into two main terms (Eq. 6.1). The first term is the array factor of the elements in series with central feeding (subarrays). The second term is related to the combination of several of the previous elements.

$$AF = AF_{subarray} \cdot AF_{global} \quad (6.1)$$

The global array factor is shown in Equation 6.2 (where N denotes the number of elements and N_s denotes the number of serial elements in the subarray). The distance among elements in this array factor is $\frac{N}{2N_s}$ times the distance between single elements.

Therefore, there will be several grating lobes in the global array factor, these lobes will disappear or not in the final radiation pattern depending on the subarray radiation pattern.

$$AF_{global} = \sum_{i=0}^{\frac{N}{2N_s}-1} e^{j k_0 \frac{N}{2N_s} d \sin(\theta) i} \quad (6.2)$$

The subarray array factor at the lower frequency is expressed in Eq. 6.3. The first exponential term is related with the difference in the phase excitation caused by the serial feeding. To calculate this phase difference the phase constant of a rectangular waveguide is used as shown in the second part of Eq. 6.3 (taking into account that at the central frequency the elements must be in phase). This effect causes some nulls to disappear and to be replaced by local minimums.

$$AF_{subarray} = \sum_{i=0}^{N_s-1} e^{j 2\pi \frac{\beta_{low}}{\beta_{central}} i} \left(e^{j k_o \sin(\theta)(-d/2-d i)} + e^{j k_o \sin(\theta)(d/2+d i)} \right) \quad (6.3)$$

$$\beta = \frac{1}{j} \sqrt{-\left(\frac{2\pi f}{c_0}\right)^2 \epsilon_r + \left(\frac{\pi}{a}\right)^2}$$

Figure 6.4 shows the array factor for the case of two elements in series ($N_2 = 2$) at the lower frequency. There are two grating lobes in the global array factor. One of them is erased in the final array factor since there is a null at this position in the subarray array factor. The other one is not erased since there is not a null in the subarray array factor. For this reason there is a higher side lobe than expected.

FIGURE 6.4: Array factor decomposition at lower frequency for two serial elements.

The subarray array factor for $N_2 = 2$ can be re-written as Equation 6.4. It probes that there is only one null instead of the usually two nulls in a 4 element array factor. The A term would be equal to one if the elements were in phase. Depending on the phase difference, this A term could be real and a second null will appear although the position of this null would not be equal to the grating lobe position, hence, the SLL would not decrease.

$$\cos(x) (8A \cos^2(x) + 2 - 6A) \quad (6.4)$$

with $x = k_0 \sin(\theta) d/2$ and $A = e^{j 2\pi \frac{\beta_{low}}{\beta_{central}}}$

6.2.1 Central feeding technique in waveguide technology

In waveguide technology the radiated lineal polarization sense depends on the waveguide propagation sense. With the central feeding technique there are opposite propagation senses in each half, therefore each half will radiate with an opposite vector as shown in Figure 6.6 (a). To overcome this problem an extra length of $\lambda_g/2$ is added in one of the arms, changing the phase difference between arms in 180° . This way, all the elements will radiate with the same phase at central frequency as depicted in Figure 6.6 (b). If the elements had some offset to the waveguide center, this 180° phase shift could be obtained with an alternatively positive and negative offset. However the elements with offset (e.g. slots) do not cover the necessary bandwidth.

FIGURE 6.5: Waveguide with central feeding with elements out of phase (a) and in phase (b).

Since the distance among serial elements must be λ_g , the distance between the feeding point and the first element should be $\lambda_g/4$ in one side and $3\lambda_g/4$ in the other side. In the real design a distance of $\lambda_g/4$ between the coupling H-slot and the first radiating element is very small and can cause error in the amplitude distribution in the radiating element. For this reason, an extra length has to be added as shown in Figure 6.6 (a). The array factors varying this extra distance are shown in Figure 6.6 (b). The lobes in directions above 40° are not very disturbing considering that the radiating element has lower gain in these directions. For the longest distance the SLL is increased in around 8 dB. The optimum distance has to be determined with a full wave simulator and is a

trade-off between a low SLL because of the array factor and a good performance because of the electromagnetic behavior in the feeding.

FIGURE 6.6: Array factor changing the distance between central elements at lower frequency (29 GHz).

In the previous analysis they are only plotted the cases for the lower frequency since the upper frequency results are very similar.

6.3 Radiating element

The radiating element simulation model is shown in Figure 6.7. It consists in a circular hole made in the top of the waveguide that excites a double patch. Every path is responsible for one of the resonances. The substrate used for the SIW is RO4350B with a height of 1.524 mm. Both patches are printed in a 0.254 mm PTFE substrate since plated holes are not necessary in the patches. The patches and SIW are separated by 1 mm foam sheets. The antenna E-plane corresponds to the longitudinal plane of the waveguide whereas the antenna H-plane corresponds to the transversal plane of the waveguide.

To validate the radiating element a test board was manufactured (Figure 6.8). There are two prototypes for the radiating elements. The first one consists of two elements central fed by a transition with two output ports (see Section 6.4). The second one is a radiating element fed by a transition with one output port.

The measurements of the radiating elements are shown in Figure 6.9. The agreement between simulations and measurements is higher in the second case than in the first one. Nevertheless the results show that it is feasible to cover the upper ka band with these elements with a good matching.

FIGURE 6.7: Radiating element scheme.

FIGURE 6.8: First test board.

6.4 Regular waveguide-SIW transition

There are two main options to feed the antenna: a coaxial line with a standard connector or a standard waveguide.

SMA connectors do not cover the upper ka band. One options is to use K-type connectors. These connectors are very fragile and requires very precises custom mechanical pieces in order to have a good transition behavior.

On the other hand, standard waveguide technology are not that fragile and do not require custom mechanical pieces in the transition. For this reason the antenna is fed by a WR28 rectangular waveguide.

The transition from regular waveguide to SIW is taken from [79]. Figure 6.10 shows a scheme of a double though plus two transitions (the upper ground is hidden). The standard waveguide is perpendicular to the SIW and it fixed through 4 screws. The transition consists of an E-plane T junction that works as a transition and a first power divider. One of the characteristics of the E-plane T junction is that the outputs have a phase difference of 180° that must be taken into account. To improve the bandwidth a

FIGURE 6.9: Radiating element matching. One single element (a) and two elements with central feeding (b).

gradually step transition in the SIW is used. The waveguide and the SIW are communicated through a rectangular aperture in the SIW bottom ground. The main parameters to be optimized are depicted in Figure 6.10.

The transition can also be done with only one output shielding one of the arms at $\lambda_g/4$. Both transitions (with two outputs and one output) can be seen in Figure 6.8.

FIGURE 6.10: Waveguide to SIW transition.

The measurements of the double through are shown in Figure 6.11. The length is $6\lambda_g$. The matching agreement between simulation and measurement is high, covering the upper ka band with a matching lower than -17 dB. The losses are higher than expected, one of the explanations is that RO4350B $\tan\delta$ is specified at 10 GHz and maybe the value is higher in higher frequencies. Nevertheless, the losses would be lower if a substrate

based on reinforced PTFE or another low loss material was used as it is done in the final prototype. The results for the transition with only one output port are not presented since they are very similar.

FIGURE 6.11: Measurements of a double through with two waveguide to SIW transitions. Matching (a) and losses(b).

6.5 Subarray

The final subarray contains 2×4 elements as shown in Figure 6.12. The subarray is fed through an H-slot that also works as a first divider. The next power division is done through a regular T-divider with an inductive wall.

FIGURE 6.12: Subarray Scheme.

Two optimizations are done for the T-divider and the H-slot. The first one prioritizes the matching in the working bandwidth and the second one prioritizes the achievable

FIGURE 6.13: Subarray matching comparison between waveguide model and SIW model.

bandwidth with a matching under -10 dB. The first subarray prototype is shown in Figure 6.14. There is another SIW under the subarray. This SIW serves as a bridge between the WR28 waveguide and the subarray, thus, it has an H-slot on the top and a SIW-waveguide transition on the bottom. Nylon bolts are used to press the two SIWs and to center the patches. As shown, these bolts are bigger than the radiating elements. For this reason, the two SIW substrates are also stuck with conductive epoxi so that the number of bolts can be reduced to three (just for centering the patches).

FIGURE 6.14: Subarray Prototype.

Two different subarray prototypes are manufactured and measured. Figure 6.15 (a) shows the matching for the first one. It also compares the difference between sticking the SIWs (Meas 2) and using bolts (Meas 1). As shown, the use of conductive epoxi is a feasible alternative, especially for the final prototype where the number of bolts is higher. The second prototype (Figure 6.15 (b)) has a better matching.

FIGURE 6.15: Subarray matching measurement.

6.6 Array design

The antenna is composed of 4x8 subarrays. The power distribution network that feeds the subarrays has some tapering. This tapering is optimized so that the level of side lobes is as lower as possible.

An asymmetric waveguide T-divider is shown in Figure 6.16. Depending on the offset to the center of the inductive wall, the transmission to each arm is modified. One of the differences compared to other technologies is that when changing the offset not only the amplitude ratio is modified but also the phase difference between arms. This means that the phase has to be compensated in every T.

FIGURE 6.16: Asymmetric T-divider.

The general scheme of the power distribution network is depicted in Figure 6.17. As shown, there are five different asymmetric Ts (the first two Ts are symmetric). The power that receives each subarray can be calculated as a multiplication of the power ratio of each branch that takes to get to the subarray. Therefore, the power is the multiplication of the three power ratios corresponding to the asymmetric Ts and the

power ratio of the first two symmetric Ts (Equation 6.6). The power ratio in each T can be normalized (Equation 6.6), reducing the number of variables to five.

$$P_{sub} = \frac{1}{4} \prod_1^3 Tx_{arm}^2 \quad (6.5)$$

$$Tx_{arm2} = \sqrt{1 - (Tx_{arm1})^2} \quad (6.6)$$

FIGURE 6.17: Distribution network scheme.

Taking into account the power that receives each subarray and the simulated radiation pattern of each subarray, the overall radiation pattern can be calculated. To optimize this radiation pattern, the cost function of Equation 6.7 is defined.

$$f_{cost} = \frac{C1}{SLL_{Hplane}} + \frac{C2}{SLL_{Eplane}} \quad (6.7)$$

The optimization algorithm is a genetic algorithm. The chromosomes are the five power ratios commented before. The parameters that have been changed from the general MATLAB optimizer are shown in Table 6.1. The lower bound is taken from the real limitation in a T-divider. In this case, the maximum advisable ratio is -6 dB (0.5 in natural units) in one of the arms.

The weighting between the SLL in the H-plane and in the E-plane has to be done carefully. The SLL in the E-plane is going to be lower than in the H-plane since some elements are in series in this plane and the number of subarrays in this plane are lower. For this reason, the weight that corresponds to the E-plane should be lower than the

Initial chromosomes	$\text{ones}(1,5) \cdot \frac{1}{\sqrt{2}}$
Pool size	50
Upper bound	$\text{ones}(1,5)$
Lower bound	$\text{ones}(1,5) \cdot 0.5$

TABLE 6.1: Genetic algorithm parameters.

H-plane weight. After some optimizations, the best found weighting is C1 equal to one and C2 equal to 2.

Figure 6.18 shows the results before and after optimization. The improvement in the E plane is lower than in the H plane as it was commented before. In the case of the E plane there are two lobes that are hardly modified. These two lobes correspond to the lobes caused by the serial feeding (Section 6.2) and cannot be reduced with tapering. In the case of the H plane the SLL is improved by 11 dB. The antenna directivity is only reduced in 0.1 dB because of the tapering.

FIGURE 6.18: Normalized array factor after optimization.

The optimized feeding is presented in Figure 6.19.

6.7 Final prototype

The subarrays can be manufactured independently or in one single SIW. The advantage of manufacturing them independently is that if the metallization process is not well done in one of them, it can be replaced. On the other hand, if they are manufactured in one single SIW board it is easier to stick them to the power distribution network. Figure

FIGURE 6.19: Optimal feedings for reducing SLL.

6.20 shows the photo of the first prototype. The substrate used for the SIW structure is RO4003C. The patches are printed in a thin PTFE based substrate.

FIGURE 6.20: First prototype layer view.

The measured matching is presented in Figure 6.21. The matching is lower than -12 dB in the upper ka band. This matching is mostly influenced by the waveguide-SIW transition and the first T-dividers. Since the losses in the antenna are close to 5 dB, the reflected signal from the radiating elements is attenuated around 10 dB, therefore, it does not have a significant influence in the antenna matching. However, the subarrays and the patches have been measured independently, showing a good behavior. Moreover if they were not well matched the total losses would be larger than 5 dB.

The measured radiation patterns are shown in Figure 6.22. The E-plane corresponds to the plane in which the radiating elements do not have the same distance among them. For this reason there is a big lobe close to 20° . The expected level for this lobe were 15 dB and it was supposed to have the maximum level at 30 GHz. The measurements show that the maximum level has moved to 30.5 GHz and the lobe level for 30 and 29.5 GHz has decreased. One of the possible causes is a permittivity change respected

FIGURE 6.21: Measured antenna matching.

to the specified value. A permittivity change would not affect very much the matching of T-dividers and transitions but would affect the radiation since the serial feeding is optimized for the specified permittivity value. The H-plane diagrams are presented in Figure 6.22 (b). The expected side lobe levels at the central frequency for the H-plane were 24 dB. The measured side lobe level at 30.5 GHz is close to 20 dB, the levels for the other frequencies are higher.

FIGURE 6.22: Measured radiation patterns.

The measured gain is plotted in Figure 6.23. The simulated directivity is 32-33 dBi, thus, the losses in the structure are close to 5 dB in the frequencies where the gain is maximum. This losses could be reduced a lot if another material based on PTFE was used. For instance, an antenna made in reinforced PTFE substrate would have a losses lower than 3 dB. The bandwidth has been reduced, in the upper frequencies the

antenna has a gain close to 27 dB but there is a decrease in the lower frequencies. One of the causes is the increment of the lobes at 20° at lower frequencies. These lobes were supposed to have a lower level, but at 29.5 GHz they have almost the same level that the main level, thus, the antenna is dividing the power among three lobes, decreasing the directive gain. However, the results are not that bad taking into account the university manufacturing capabilities.

FIGURE 6.23: Measured gain.

6.8 Conclusions

The design of a broadband antenna in SIW technology at ka band has been introduced. The antenna uses a combination of serial and parallel feeding (serial distribution in subarrays and parallel distribution among subarrays). In this way, the design takes advantage of the small size of the serial distribution at the end of the array and the broadband behavior of the parallel distribution. In the serial distribution the SIW are central fed, this way, the antenna always points to broadside independently of the frequency. The main disadvantage of the proposed design is that it has a large side lobe because not all the elements are at the same distance. The first manufactured prototype has a lower bandwidth than in the simulations, although the result is not that bad taking into account the used substrate material and the manufacturing constraints.

The antenna has been manufactured using the university capabilities. For this reason, there is a lot of things that can be improved in future designs. Every layer is pressed through bolts but a multilayer structure could be used reducing the miss-alignment between boards and preventing the leakage. The antenna losses can be reduced if other substrates more suitable for these frequencies were used, e.g., reinforced PTFE.

Larger arrays antennas could be used. The only drawback would be the losses in the distribution network. To overcome these losses, a simple distribution network made in regular WR28 waveguide technology can be used. This network could feed four or eight SIW array antennas without increasing too much the losses. Moreover, the network in waveguide technology would be relatively easy to manufacture since it would only feed 4 or 8 ports separated more or less 150 mm.

The coupling among subarrays are not taken into account. The influence of these coupling could be obtained through full wave simulations, using E field monitors. This way, the couplings could be compensated by varying the patch radius in every element. However it would require a high computational cost.

Chapter 7

Gap Waveguide antenna

7.1 Introduction

This chapter describes the design of a small antenna using a novel technology named gap waveguide. The objective is to design an antenna in the upper ka band (from 28 to 31 GHz) using this technology and the standard printed circuit process.

Although SIW technology has an easy manufacturing process and lower losses compared to other printed technologies, it has the inconvenient of the dielectric losses. The dielectric losses can be reduced using more expensive materials but there is a limit as the frequency goes up, not only because of the loss increase but also because of the influence of the dielectric characterization.

The novel alternative in high frequencies is the gap waveguide technology. This technology consist of a ridge waveguide in which the lateral walls are substituted by a meta-material surface. This surface stops the propagation of power in the transversal direction, confining the wave around the central ridge. An scheme of how it works is shown in Figure 7.1.

One of the advantages of this technology is that it does not require a union between the top metal ground plane and the structure with the ridge and pins. The first attempts using this technology have been manufactured milling the pins and the ridge in an aluminum block (see [45]). In this thesis the pins and the ridge are done with plated holes in a dielectric substrate. Besides the lower price, the printed circuit technology has the advantage of higher accuracy. The main drawback could be the dielectric losses but since most of the wave is confined in the gap between the ridge and the top ground plate, the wave only sees air as dielectric. Figure 7.2 shows a transversal cut of the two

FIGURE 7.1: Scheme of the meta-material surface in Gap Waveguide.

manufacturing options. The metal walls in the second option are used just to separate the ridge from the top ground plane the appropriate distance.

FIGURE 7.2: Two manufacturing possibilities: milling a metal piece (a) and using printed circuit technology (b).

A comparison between a waveguide and a ridge waveguide is done in Section 7.2. The implementation of the metamaterial surface is explained in Section 7.3. This surface is implemented using mushroom type Electromagnetic Band Gap surface (EBGs). The antenna is fed through a regular waveguide, thus, a transition from regular waveguide to gap waveguide is needed. This transition is done in two steps, the first one is a transition between the regular waveguide and a small SIW line (similar to the one designed in the previous Chapter), the second one is a transition between the SIW and the gap waveguide. This transition is explained in Section 7.4. The design of the radiating

element is explained in Section 7.6. The design of the final array is presented in Section 7.7.

7.2 Ridge Waveguide

A ridge waveguide is a rectangular waveguide in which a small ridge has been introduced, generally at the center of the wider face as show in Figure 7.3. The big advantage of a ridge waveguide compared to a traditional rectangular waveguide is that it has a wider mono-mode bandwidth. Ridge waveguide has also a smaller size than rectangular waveguide for a specific cut-off frequency. The main drawbacks are the more complex manufacturing process and the lower power handling. Design equations and analytical models for ridge waveguides filled only with one dielectric can be found in [80–82].

FIGURE 7.3: Ridge waveguide cross-section.

The ridge waveguide that is used in this chapter is done using printed technologies and a metamaterial surface that simulates the walls. However, in a first step, the ridge waveguide with metal walls is studied. The simulated cross-section is shown in Figure 7.4. It has two different dielectrics: air and a RO4003C substrate. In the final prototype the ridge is implemented with plated vias from the bottom layer to a strip in the top substrate layer.

FIGURE 7.4: Ridge waveguide cross-section with two dielectrics.

Figure 7.5 shows the dispersion diagram for a ridge and a rectangular waveguide. Both waveguides are half filled with Ro4003C substrate. This means that the waveguide has two dielectrics: air and Rogers. The substrate height used in the simulations is 1.524 mm and the ridge width is 2 mm. As it is shown, the bandwidth for the ridge waveguide is around 30 % higher. The bandwidth could be extended if proper height and ridge width were chosen, however, the bandwidth is more than enough for a single band array antenna. More bandwidth could be useful if a ultra wide band or a multi-band antenna were used.

FIGURE 7.5: Dispersion diagram of the first three modes in a rectangular waveguide and a ridge waveguide filled with air and Rogers substrate.

The election of the ridge dimensions are based on minimizing the losses more than increasing the bandwidth since the bandwidth is much more than needed. In this way, the ridge height is determined by the available substrate heights whereas the only ridge width constraint is the minimum drill diameter. Figure 7.6 shows the losses per meter of a ridge waveguide changing the width and height of the ridge. The total height is maintained in the simulations, thus, if the ridge height is decreased, the waveguide will contain more air and less substrate. As shown the losses are lower when the height is lower. This is logical since most of the losses are in the dielectric substrate. However, the minimum height is more influenced by the minimum height to suppress the propagation in the meta-surface that simulates the walls as depicted in Section 7.3. As far as the ridge width is concerned, the losses are lower when the width is increased. A larger ridge width is also desirable for the manufacturing process since the vias could have a larger diameter.

7.3 Ridge Gap Waveguide

The vertical walls in the ridge waveguide can be substituted by a soft surface that cancels parallel-plate modes as shown in Figure 7.1. These structures are usually implemented with a periodic array of small patches printed in a dielectric substrate. The Electromagnetic band gap surfaces or EBG surfaces can be divided into patches with central via connected to the ground plane (mushroom type) or without vias (planar type). A comparison of both structures can be found in [83].

FIGURE 7.6: Losses per meter of a ridge waveguide changing the ridge dimensions ($f=29$ GHz).

The chosen EBG for a gap waveguide is the mushroom type because of its larger bandwidth. Figure 7.7 (a) shows the unitary EBG cell. As it is shown it consists of a via hole plus a small circular path on the top. The real via holes are not filled with metal (at least with the conventional manufacturing process). The hole is plated a certain thickness around the dielectric material but keeping the central core with air. However, having filled plated holes or not does not have a high influence in the electromagnetic behavior and the simulation time is reduced if filled vias are used. In Section 7.7 is shown the simulation of the final prototype simulated with non completely filled via holes, showing that there is not a big influence in not having the via completely filled.

The optimized EBG parameters are sketched in Table 7.1. Figure 7.7 (b) shows the 2D Brillouin diagram of the optimized EBG. The plot has two curves, the lower one corresponds to the first mode whereas the second one corresponds to the second mode. The 2D Brillouin diagram measures the propagation with the frequency for 0° , 45° and 90° incident angle. Each of these directions is represented in the lower graph with one color. This way, the forbidden band is calculated taking into account every possible propagation direction in 2D. As shown, there is a band gap between 24 and 32 GHz where there is no propagation for the first mode or the second mode. This means that the propagation is not allowed at these frequencies. More details in how to calculate the 2D Brillouin diagram can be found in [84].

To check the behavior of the EBG, a simulation of a EBG grid is performed as shown in Figure 7.8 (a). The simulation has open boundaries in all the direction except for the ground plane. Two ports are placed in the edges of the grid to check the transmitted power across the structure. The A configuration corresponds to the case that has been simulated with the Brillouin diagram (since it employs periodic boundaries). In the B

FIGURE 7.7: Dispersion diagram of the 2D EBG periodic structure, based on Brillouin Zone definition.

Substrate	Ro4003C
Permittivity	~ 3.6
Substrate height	1.524 mm
Gap height	1 mm
Grid width	1.5 mm
Grid length	1.75 mm
Via diameter	0.5 mm
Pad width	0.15 mm

TABLE 7.1: Main parameters of the EBG unit cell.

configuration half of the plated holes has been eliminated. As shown in Figure 7.8 (b), the power transmitted in the A configuration agrees with the band gap obtained with the Brillouin diagram (lower S21 from 24 to 32 GHz). The B configuration has a larger band gap with half of the vias. the only problem is that it only works if the number of rows is larger than 3, making it difficult to use in practice.

To check the behavior and set the minimum number of EBG rows a simple simulation is performed. Figure 7.9 (a) shows the model used for this simulation. As presented, it consist of a first section of a regular ridge waveguide, followed by a larger section of a gap waveguide and finally ended in a small section of ridge waveguide. The reason to employ a first and final section of regular ridge waveguide is because the FIT simulator requires the structure to be identical in three mesh steps. This simulation is just an

FIGURE 7.8: Transmission of two EBG grid configurations. Models (a) and S_{21} (b).

approximation since no optimization has been performed to match better the modes of the ridge waveguide and the gap waveguide. Nevertheless, it shows that the matching is acceptable in the area of interest. It also shows that the losses are very low (lower than 0.25 dB) in the frequencies of the band gap, meaning that the EBG surface stops the transversal propagation just with two rows.

FIGURE 7.9: Thru in gap waveguide technology. Model (a) and simulated S-Parameters (b).

7.4 Transitions

There are two kind of transition for gap waveguide technology in the literature. The first one is a stepped transition as shown in [45]. This transition cannot be used here because it would require the capability of having different ridge heights which can be very complicated in regular printed circuits boards.

The second transition in a gap waveguide consist of a small microstrip network that acts as a transition between a coaxial connector and the gap waveguide (see [85]). The problem with this transitions at these frequencies is that the amount of losses of this microstrip plus the coaxial connectors could be very high. If the frequency goes up (as it is expected for this technology), these losses could be unaffordable.

Since other kinds of transitions in SIW technology have been studied in previous chapters, a SIW- gap waveguide transition is designed as shown in figure 7.11. This way the SIW losses would be lower than microstrip technology (especially if a good dielectric material is used). The SIW feed can be done through a conventional waveguide technology with the transition designed in Chapter 6. The SIW can also be fed using a microstrip line as shown in Chapter 5 but with higher losses.

Figure 7.11 shows a thru with two transitions fed by standard waveguides. The transition is depicted in three steps. In the first picture the top ground plane is hidden, in the second the metal pieces that support this ground plane are also hidden and in the last picture the top layer of the substrate is also hidden. The metal piece at the beginning of the transition is used to prevent the generation of modes in the opposite direction. If this piece was not there, at the SIW end, part of the power would be coupled to the gap waveguide whereas part of the power would be transmitted in the opposite direction. As shown in the second picture there is a triangular cut in the SIW to couple the TE₀₁ mode of the SIW to the first mode into the gap waveguide (taking into account that the power in the SIW is propagated in the substrate and most of the power in the gap waveguide is propagated above the substrate (in the gap between the substrate and the top ground plane)). Because of the presence of this triangular cut, the metal piece to prevent the backward propagation has a square hole at a distance close to $\lambda/2$. This way, this metal piece acts as a short circuit for the backward propagation. Finally, a SIW to standard waveguide transition is used.

The simulation results for the thru are shown in Figure 7.11. As expected the bandwidth has been reduced compared to the EBG bandwidth because of the presence of the transitions. However it covers the desired frequencies in the upper ka band. The simulated losses are lower than 0.4 dB.

FIGURE 7.10: Thru with regular waveguide -SIW and SIW-Gap WG transitions.

FIGURE 7.11: Simulated losses and matching of a thru with waveguide-SIW and Gap WG-SIW transitions.

7.5 Dividers and bends

The layouts of the designed T-divider and bend are shown in Figure 7.12. As shown, the EBG plated holes do not follow a rectangular grid. They are placed following the central ridge path and some of them are moved to improve the behavior of the structure.

The T-divider has a 50 to 100 Ω adapter plus a triangular cut to improve the matching. It also has plated holes with a smaller radius to follow the ridge in the areas where its width is smaller.

FIGURE 7.12: Bend (a) and T-Divider (b) in gap waveguide technology.

The matching and losses are not presented since they are very similar to the thru of the previous section.

7.6 Radiating element

The radiating element is show in Figure 7.13. It consists of a rectangular aperture in the top ground plate plus a printed circular patch above this aperture (the substrate where the path is printed is hidden in the Figure). The polarization is linear and the phase changes with the propagation direction in the ridge (a difference of 180° with opposite propagation). Although the radiating element could be only the rectangular aperture, the path adds an extra resonance.

FIGURE 7.13: Radiating element. Inner part of the gap waveguide (a). Aperture in the ground plus the patch (b).

The matching is presented in Figure 7.14 (a). As shown, there are two resonance that increase the bandwidth to about 10%. Figure 7.14 (b) shows the 3D radiation pattern.

As shown it has a gain of 6.76 dB with a shape very similar to the diagram of a standard patch.

FIGURE 7.14: Radiating element matching (a) and 3D diagram (b).

The gain at broadside over frequency is shown in Figure 7.15. As shown, the differences in gain are less than 0.3 dB in the matched frequencies.

FIGURE 7.15: Gain over frequency at broadside direction for the radiating element.

7.7 4x1 Array

A four by one array is designed as shown in Figure 7.16. In this simulation, the plated holes are not filled. This means that they have a vacuum core. In this way, the simulation is more similar to the real manufacturing. The elements are spaced $0.85 \cdot \lambda_0$ (with λ_0 taken at 29 GHz). This distance is a trade-off between a minimum distance for having

the minimum number of EBG vias to stop the propagation in the transversal direction and the maximum distance for not having big grating lobes at broadside.

FIGURE 7.16: Model of the 4x1 array. Inner part (a) and ground plus patches (b).

The simulated matching is presented in Figure 7.17 (a). As shown the matching is not as good as the single element but it is good enough taking into account the constraints in the via placement. The 3D diagram at 29 GHz is shown in Figure 7.17 (b). The simulated efficiency is close to 90%.

FIGURE 7.17: Array matching (a) and 3D diagram (b).

The gain at broadside over frequency is shown in Figure 7.18. There is a gain decrease as the frequency goes up. There are two main causes for this decrease. The first one is the EBG surface. Although this surface has a stop band that ends in 31 GHz, the behavior is degraded as the stop band limit is reached. The second cause is the presence of grating lobes. the distance among elements has been chosen for not having grating lobes at 29 GHz but when the frequency goes up, the level of the grating lobes is increased.

FIGURE 7.18: Gain over frequency at broadside direction for the 4x1 array.

7.8 Test board

A first test board with the element sketched in the previous sections was manufactured. Figure 7.19 (a) shows the bottom layer where the feeding waveguide is screwed. Figure 7.19 (b) shows the gap waveguide structure with the inner metal piece that is used as a support for the top plate. Finally, Figure 7.19 (c) shows the top plate plus the patches.

FIGURE 7.19: Test board. Bottom view (a), top view without top plate (b) and top view (c).

The matchings and losses of the bend, the power divider and the radiation element are presented in Figure 7.20. As shown, the matchings are very similar to the simulations. As far as the losses are concerned they are a bit higher than in the simulations. This

is logical since the dielectric losses in the initial SIW section are always larger than the simulations because the substrate $\tan\delta$ is specified in lower frequencies.

FIGURE 7.20: Matching (a) and losses (b) of the test board.

It has not been possible to measure the radiation pattern of the 4x1 array by the time this thesis has been finished. This array is expected to be measured in the following months.

7.9 Conclusions

The novel gap waveguide technology has been studied. The traditional milling manufacturing has been substituted by printed board technology. In this way the manufacturing cost is greatly reduced. A small prototype has been built and measured. The main future line would be to extend the use of this technology to higher bands such as W band in which there is a need of directive antennas for novel radar applications. Another future line would be to design larger arrays antennas. The problem is that there is a minimum distance among elements in order to suppress transversal modes in the EBG surface. This distance is not small enough for not having grating lobes in an array antenna. There would be two approaches to solve this problem. The first one would be to make subarrays in a different technology such as microstrip or SIW and feed them through gap waveguide. This way the distance among subarrays would be large enough for the EBG structure and the distance among elements in the subarray would be small enough for not having grating lobes. Of course, this would increase the losses accordingly to the subarray size. The other option to solve this problem is to use new geometries that

allow having larger distance among elements without increasing the grating lobes. One option could be to run genetic optimizations of distance offsets among rows of elements.

Chapter 8

Conclusions

8.1 Framework

This thesis has been carried out in the Radiation Group of the Signal, Systems and Radiocommunication Department in the ETSIT de Telecomunicación of Technical University of Madrid from October 2010 to May 2015 under the supervision of Dr Manuel Sierra Pérez.

The contributions presented in this manuscript have been reflected in public research projects:

- *Project title*: Communication systems for emergency environments (SICOMORO) TEC2011-28789-C02-01. *Financial institution*: Ministerio de Educación y Ciencia. *Research Institution*: Universidad Politécnica de Madrid. *Research in chief*: Belén Galocha Iragüen.
- *Project title*: Caracterización de Canales Radio, Optimización y Calibrado de la Antena GEODA (CROCANTE) TEC2008-06736-C03-01. *Financial institution*: Ministerio de Educación y Ciencia. *Research Institutions*: Universidad Politécnica de Madrid and Universidad de Vigo. *Research in chief*: Leandro de Haro y Ariet.

8.2 Novel Contributions

The main novel contributions of this thesis are pointed out below.

- The design of a man pack antenna for satellite communications at X band. The antenna uses both circular polarizations simultaneously. The antenna has a modular design with two main parts: the subarrays and the low loss power distribution network. The subarrays implements several new techniques to cover the whole X band. They necessary hybrid circuit to have circular polarization uses a novel miniaturization. The power distribution network is done using shielded stripline. The novelty is not in the use of this technology but in the band where it is used. Moreover, a novel ad-hoc pluggable connector is used between subarrays and power distribution network.
- The design of a reflective phase shifter with a novel configuration. The reflective phase shifters that use varactors have at least three stages to cover the 360° . In the proposed design, this is obtained with one single stage using a series-parallel configuration.
- The design of a phased array for a novel application (large range communications with an UAV). The phased array combines the used of commercial WiMAX monolithics with a shaped beam designed to give the UAV the maximum coverage.
- The design of a slotted array in Substrate Integrated Waveguide at ku band. The design impements classical elements in this new technology. It also uses a meander line polarizer attached to the antenna at a short distance.
- The design of a broadband SIW antenna at ka band. This antenna uses some novel elements such a patch coupled to waveguide, H coupling slot and regular waveguide to SIW transition in the E plane. All the elements are designed to cover the necessary bandwith with a low profile structure.
- The design of a small antenna using gap-waveguide technology. This novel technology reduces greatly the losses, keeping an easy manufacturing process. The proposed antenna uses a novel manufacturing process in which the milling is replaced by printed board metallization. This reduces the price and allows to design more complex structures. The design includes also a novel SIW to Gap waveguide transition.

8.3 Future research lines

The future lines are explained at the end of every chapter. A small summary is included below.

- To build and test a large array at X band using shielded suspended stripline and ad-hoc connectors. The previous prototypes presented in Chapter 2 were made using expensive materials and connectors with high losses. To reduce the price and overcome the connector losses a new array was proposed but they were only tested small prototypes.
- To include the proposed phase shifters in Chapter 3 in an array antenna. These phase shifters were designed to be included in the antenna of Chapter 2. Although the first prototype showed a good behaviour, it would be necessary a industrialization of the manufacturing process if they want to be included in an antenna.
- To include a Butler network in the antenna of Chapter 4 instead of commercial phase shifters. This way the antenna price could be reduced and it could have several simulated beams if needed.
- To substitute the standard SIW technology used in Chapters 5 and 6 by novel structures such as multilayer SIW keeping the core with air. This way the dielectric losses could be reduced.
- To design an array antenna using gap waveguide technology in the W band. This antenna should use the printed board manufacturing process. It also should have a genetic optimization of the distance among elements to allow larger distance among elements without increasing so much the grating lobes.

8.4 Publications

The publications done during the course of this thesis are listed below.

8.4.1 Journal publications

- Garcia-Aguilar, A.; Inclan-Alonso, J.; Vigil-Herrero, L.; Fernandez-Gonzalez, J.; Sierra-Perez, M., “Low-Profile Dual Circularly Polarized Antenna Array for Satellite Communications in the X Band,” *Antennas and Propagation, IEEE Transactions on* , vol.60, no.5, pp.2276,2284, May 2012, doi: 10.1109/TAP.2012.2189729.
- Inclan Alonso, J.M.; Amazaes Calderon, G.; Sierra Perez, M., “SIW Antenna With Polarizer at ku-Band,” *Antennas and Propagation, IEEE Transactions on* , vol.63, no.6, pp.2782,2786, June 2015 doi: 10.1109/TAP.2015.2414471.

- Inclan Alonso, J.M.; Sierra Perez, M., “Phased Array for UAV Communications at 5.5 GHz,” *Antennas and Wireless Propagation Letters, IEEE* , vol.14, no., pp.771,774, 2015 doi: 10.1109/LAWP.2014.2379442.
- J. M. Inclan-Alonso, J.-M. Fernandez Gonzalez, and M. Sierra-Perez, “Low loss power distribution networks in stripline technology for planar array antennas,” *Progress In Electromagnetics Research*, Vol. 143, 369-384, 2013. doi:10.2528/PIER13061107 .

8.4.2 Conference contributions

8.4.2.1 International

- Garcia-Aguilar, A.; Inclan-Alonso, J.M.; Fernandez-Gonzalez, J.M.; Sierra-Perez, M., “Printed antenna for satellite communications,” *Antennas and Propagation (EuCAP), 2010 Proceedings of the Fourth European Conference on* , vol., no., pp.1,5, 12-16 April 2010.
- Garcia-Aguilar, A.; Inclan-Alonso, J.M.; Vigil-Herrero, L.; Fernandez-Gonzalez, J.M.; Sierra-Perez, M., “Printed antenna for satellite communications,” *Phased Array Systems and Technology (ARRAY), 2010 IEEE International Symposium on* , vol., no., pp.529,535, 12-15 Oct. 2010 doi: 10.1109/ARRAY.2010.5613315.
- Garcia-Aguilar, A.; Inclan-Alonso, J.M.; Vigil-Herrero, L.; Gonzalez, J.M.F.; Sierra-Perez, M., “Dual-polarized portable printed antenna for satellite communications at X-Band,” *Antennas and Propagation Conference (LAPC), 2010 Loughborough* , vol., no., pp.429,432, 8-9 Nov. 2010 doi: 10.1109/LAPC.2010.5666217.
- Inclan-Alonso, J.M.; Garcia-Aguilar, A.; Vigil-Herrero, L.; Fernandez-Gonzalez, J.M.; Sanmartin-Jara, J.; Sierra-Perez, M., “Portable low profile antenna at X-band,” *Antennas and Propagation (EUCAP), Proceedings of the 5th European Conference on* , vol., no., pp.1923,1927, 11-15 April 2011.
- Inclan-Alonso, J.M.; Garcia-Aguilar, A.; Vigil-Herrero, L.; Fernandez-Gonzalez, J.M.; Sanmartin-Jara, J.; Sierra-Perez, M., “Modular planar antenna at X-band for satellite communications,” *General Assembly and Scientific Symposium, 2011 XXXth URSI* , vol., no., pp.1,4, 13-20 Aug. 2011 doi: 10.1109/URSIGASS.2011.6050477.
- Inclan-Alonso, J.M.; Perez, M.S., “Man-pack antenna at X-band,” *Antennas and Propagation (EuCAP), 2013 7th European Conference on* , vol., no., pp.2222,2226, 8-12 April 2013.

- de Toca, A.N.S.; Inclan-Alonso, J.M.; Trujillo, J.G.; Gonzalez, J.M.F.; Perez, M.S., “GEODA-SARAS: Multi-phased array planar antenna for satellite communications,” *Antennas and Propagation (EuCAP), 2013 7th European Conference on* , vol., no., pp.252,253, 8-12 April 2013.
- Fernandez Gonzalez, J.M.; Inclan Alonso, J.M.; Garcia-Gasco Trujillo, J.; Sanchez de Toca, A.N.; Perez, M.S., “GEODA-SARAS: Multi-phased array planar antenna design and measurements,” *Antennas and Propagation (EuCAP), 2014 8th European Conference on* , vol., no., pp.423,425, 6-11 April 2014 doi: 10.1109/EuCAP.2014.6901782.

8.4.2.2 National

- J.M. Inclan-Alonso, A. Garcia-Aguilar, M. Sierra-Perez and J.M. Fernandez-Gonzalez “Antena plana com polarización circular dual para comunicaciones por satélite em banda X” *URSI 2010*. Bilabo. Spain. Sep. 2010.
- Lucía Vigil Herrero, Andrés García Aguilar, José Manuel Inclán Alonso, José Manuel Fernández González, Manuel Sierra Pérez “Antena plana para comunicaciones por satélite en banda X” *URSI 2011*. Leganés. Spain.Sep. 2011.
- J.M. Inclán-Alonso, J.M. Fernández-González and M. Sierra-Pérez “Low losses power distribution network at X band in stripline” *URSI 2012*. Elche. Spain. Sep. 2012.
- Gloria Amazares Calderón, José Manuel Inclán Alonso, Manuel Sierra Pérez “Antena Plana banda ku en tecnología SIW” *URSI 2013*. Santiago. Spain. Sep. 2013.
- José Manuel Inclán Alonso, Álvaro Noval Sánchez de la Toca, Javier García-Gasco Trujillo, José Manuel Fernández González y Manuel Sierra Pérez, “Radiating element of GEODA-SARAS” *URSI 2013* Santiago. Spain. Sep..

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