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**ADVANCES IN PERIODIC STRUCTURES AND
MANUFACTURING TECHNOLOGIES FOR MM-WAVE
ANTENNAS**

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Advances in Periodic Structures and Manufacturing Technologies for mm-Wave Antennas

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SUMMARY

The arrival of 5G services demands the use of an increasing bandwidth, which leads to better use of the electromagnetic spectrum. 5G mass deployment requires solutions for working in upper frequency bands with low cost manufacturing. Printed substrate technologies, as stripline or microstrip, exhibit high transmission losses at high frequencies due to the presence of dielectric. At high frequencies, printed technologies have been replaced by waveguide. This technology is based on the propagation of electromagnetic waves in a hollow metallic cavity, which supposes very low transmission losses. However, materials and manufacturing mechanisms are expensive and the resulting structures are heavy and bulky. This makes waveguides incompatible with the development and mass production of small and low-cost technology.

In recent years, various technologies have been emerged, such as gap waveguide (GW), higher symmetries (HS) or additive manufacturing (AM), that present a solution for new millimetre-wave devices. These new technologies are becoming globally important because they offer good performance in terms of low weight, low profile and low losses, and a low-cost fabrication on a massive scale compared to other PCB technologies or classical waveguides.

Gap waveguide technology is based on fully metal pieces with several rows of periodic pins that prevent leakage in a certain frequency band when a metal plate is placed above them a distance below a quarter wavelength. A groove or a ridge can be inserted between those pins to propagate fields in the same way that classic waveguides, with the advantage that GW do not need good electrical contact between plates.

Higher symmetries consist of periodical structures with symmetry operations such as screw symmetry (a periodical rotation), or glide symmetry (a mirroring and displacement). Especially this last type of symmetry is of great interest because it can provide EGB (Electromagnetic Bandgap) properties or increase the refractive index in a multitude of different transmission lines.

The fabrication of the resulting structures in a complex design based on GW or HS can be costly using traditional CNC machining methods. In millimetre-wave bands, GW pins tend to be small and with a reduced separation between them, which increases the price. For that reason, it is interesting to consider additive manufacturing or 3D-printing for this type of structures, given that the time and cost of manufacturing do not depend on the complexity of the part, but on its size.

The increasing introduction of new materials with anisotropic properties that can be modified by applying a voltage has also been documented. Among these materials, liquid crystals (LC) stand out, which were studied in optical applications. LC can have a great importance in the creation of new electrically reconfigurable antennas or RF devices.

The objective of this thesis is to study in depth these novel technologies and materials and apply them in real prototypes of antennas and RF devices in millimetre-wave bands. Specifically, in chapter I a detailed introduction of each of the technologies used during the development of

the thesis is carried out. In Chapter II, a comparison between available technologies such as rectangular waveguide, substrate integrated waveguide, microstrip and gap waveguide technologies is carried out in terms of transmission losses between 10 GHz to 100 GHz. Chapter III shows a study of the properties of 3D printing and the effect of metallization of the resulting plastic parts at Ka-band. Different prototypes implemented in GW were designed, manufactured and experimentally validated. Chapter IV focuses on a more complex GW design for a Radial Line Slot Array (RLSA) antenna fed with a 3D-printed Butler matrix for a monopulse radar at 94 GHz. 3D printing technology, together with a liquid crystal mixture developed specifically for microwave frequencies, were used for the design and manufacturing of an electrically tunable phase shifter in chapter V. Chapter VI contains the validation of several prototypes based on higher symmetries that seek to enhance different aspects: stop-band and refractive index modifications and attenuation and phase shift mechanical reconfigurability. These higher symmetries are also applied in Chapter VII for enhanced gap waveguide EGB or increase of the beam steering variation with frequency in leaky-wave antennas for automotive applications. Finally, conclusion and future work are drawn in Chapter VIII.

KEY WORDS:

Millimeter-wave, antenna, gap waveguide, higher symmetries, glide symmetry, additive manufacturing, 3D-printing, liquid crystals, Ka-band, W-band, leaky wave antenna, monopulse, radar, RLSA.

RESUMEN

La llegada de los servicios 5G exige el uso de un ancho de banda cada vez mayor, lo que conduce a un mejor uso del espectro electromagnético. El despliegue masivo de 5G requiere soluciones para trabajar en bandas de frecuencias altas con un bajo coste de fabricación. Las tecnologías de sustratos impresos, como stripline o microstrip, presentan altas pérdidas de transmisión a altas frecuencias debido a la presencia de dieléctricos. A altas frecuencias, las tecnologías impresas han sido reemplazadas por guías de onda. Esta tecnología se basa en la propagación de ondas electromagnéticas en una cavidad metálica hueca lo que supone unas pérdidas de transmisión muy bajas. Sin embargo, los materiales y los métodos de fabricación son caros y las estructuras resultantes son pesadas y voluminosas. Esto hace que las guías de onda sean incompatibles con el desarrollo y la producción en masa de tecnología de pequeño tamaño y bajo coste.

En los últimos años han surgido diversas tecnologías, como *gap waveguide* (GW), simetrías superiores (HS) o la fabricación aditiva (AM), que presentan una solución para nuevos dispositivos de ondas milimétricas. Estas nuevas tecnologías están adquiriendo importancia mundial porque ofrecen un buen rendimiento en términos de bajo peso, bajo perfil y bajas pérdidas, y un bajo coste de fabricación a gran escala en comparación con otras tecnologías de circuitos impresos o guías de onda clásicas.

La tecnología *gap waveguide* se basa en piezas totalmente metálicas con varias filas de pines periódicos que evitan las fugas en una banda de frecuencia determinada cuando se coloca una placa de metal por encima de ellos a una distancia inferior a un cuarto de longitud de onda. Se puede insertar una ranura (*groove*) o una cresta (*ridge*) entre esos pines para propagar los campos de forma similar a las guías de onda clásicas, con la ventaja de que GW no necesita un buen contacto eléctrico entre las placas.

Las simetrías superiores consisten en estructuras periódicas con operaciones de simetría tales como simetría de tornillo (una rotación periódica), o simetría de deslizamiento (un reflejo y desplazamiento). Especialmente este último tipo de simetría es de gran interés debido a puede proporcionar propiedades EGB (*Electromagnetic Bandgap*), o aumentar el índice de refracción en una multitud de líneas de transmisión diferentes de forma sencilla.

La fabricación de las estructuras resultantes en un diseño complejo basado en GW o HS puede ser costosa utilizando métodos tradicionales de mecanizado CNC. En las bandas de ondas milimétricas, los pines GW tienden a ser pequeños y con una separación reducida entre ellos, lo que aumenta el precio. Por ello, es interesante considerar la fabricación aditiva o impresión 3D para este tipo de estructuras, ya que el tiempo y el coste de fabricación no dependen de la complejidad de la pieza, sino de su tamaño.

También se ha documentado la creciente introducción de nuevos materiales con propiedades anisotrópicas que pueden modificarse aplicando una tensión. Entre estos materiales destacan los cristales líquidos (LC), que fueron estudiados en profundidad en aplicaciones ópticas. Los cristales líquidos pueden tener una gran importancia en la creación de nuevas antenas o dispositivos de RF reconfigurables eléctricamente.

El objetivo de esta tesis es estudiar en profundidad estas nuevas tecnologías y materiales y aplicarlos en prototipos reales de antenas y dispositivos de RF en bandas de ondas milimétricas. Concretamente, en el capítulo I se realiza una introducción detallada de cada una de las tecnologías utilizadas durante el desarrollo de la tesis. En el capítulo II se efectúa una comparación entre las tecnologías disponibles, como la guía de ondas rectangular, la guía de ondas integrada en el sustrato, la microstrip y la guía de ondas gap waveguide, en lo que respecta a las pérdidas de transmisión entre 10 GHz y 100 GHz. El capítulo III muestra un estudio de las propiedades de la impresión en 3D y el efecto de la metalización de las piezas de plástico resultantes en la banda Ka. Se diseñaron, fabricaron y validaron experimentalmente diferentes prototipos implementados en GW. El capítulo IV se centra en un diseño basado en GW más complejo para una antena RLSA (*Radial Line Slot Array*) alimentada con una matriz de Butler impresa en 3D para un radar monopolso a 94 GHz. La impresión 3D, junto con un cristal líquido desarrollado específicamente para microondas, son empleados para el diseño y la fabricación de un desfasador eléctricamente sintonizable en el capítulo V. El capítulo VI contiene la validación de varios prototipos basados en simetrías superiores que buscan mejorar diferentes aspectos: modificaciones de la *stop-band* y del índice de refracción y reconfigurabilidad mecánica de atenuación y desfase. Estas simetrías superiores también se aplican en el capítulo VII para EBG mejorado en *gap waveguide*, o aumento de la variación de la dirección del haz con la frecuencia en antenas *leaky-wave* para aplicaciones de automoción. Finalmente, las conclusiones y trabajo futuro se describen en el capítulo VIII.

PALABRAS CLAVE:

Ondas milimétricas, antena, *gap waveguide*, simetrías superiores, simetría de deslizamiento, fabricación aditiva, impresión 3D, cristales líquidos, banda Ka, banda W, antena *leaky-wave*, monopolso, radar, RLSA.

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Chapter I

Introduction and State of art

1.1 Introduction

The rapid development of new generation services and the paradigm change that the introduction of 5G communications will entail requires the availability of large bandwidths. However, the spectrum is widely occupied by other services up to K-band frequencies. Therefore, it is inevitable to use technologies that make the use of spectrum be feasible at Ka-band frequencies and above.

One of the parts that must adapt to this technological developing is the antenna system. This system constitutes one of the most critical parts at any radiocommunication platform because it must provide the best matching between the radio system and the external environment in order to send information. Planar antennas are an adequate candidate for many radiocommunication systems applications due to ease to manufacture, low cost and the possibility of an easy integration because of their planar structure and reduced size. Printed substrate technologies, such as stripline or microstrip technologies, are usually used in distribution networks and patches for planar antennas.

Nevertheless, common printed substrate technologies exhibit problems at high frequencies such as a restricted bandwidth, the appearance of additional propagating modes through the surface of the antenna that reduce the radiating efficiency and high transmission losses due to the presence of dielectric. Because of that, at high frequencies printed technologies have been replaced by waveguide. This technology is based on electromagnetic waves propagation in a rectangular, cylindrical or elliptical metal cavity. That structure supposes a reduction of radiation losses because the excited modes propagate confined inside the guide, those modes are easily controlled and dielectric losses are very low due to it is usually air. However, materials and manufacturing mechanisms are expensive and the resulting structures are heavy and bulky. This makes waveguides incompatible with the development and mass production of small and low-cost technology. A technology that could substitute waveguides by keeping their positive properties while provide a reduced size and an easier and cheaper manufacturing process will suppose an impulse for the 5G systems developing.

1.2 Frequency Bands under Study for 5G

In WRC 2015, additional spectrum allocations have been considered to the mobile service for future 5G. WRC 2015 mainly focuses on frequency bands below 6 GHz. However, higher-frequency bands up to 100 GHz were identified as good candidate spectrum for 5G because there is abundant spectrum available to support ultrahigh data rate transmission. The propagation characteristics of the candidate spectrum should be friendly for carrying mobile communication

services. Since the propagation characteristics of the mm-wave bands can be quite different compared to the frequency bands below 6 GHz, the candidate spectrum selection should consider multiple channel propagation issues, such as severe path loss, the influence of weather conditions and atmosphere, Doppler with even slow movement due to higher carrier frequency, and NLOS channels. The candidate frequency bands are represented in Figure 1-1 for frequencies below and above 6 GHz.

Figure 1-1: Candidate frequency bands for 5G research. Source [1]

As it is specified in [2], the standard nomenclature for the frequency bands between 400 MHz and 100 GHz is shown in the Table 1-I.

TABLE 1-I: STANDARD FREQUENCY LETTER BAND NOMENCLATURE

Band designation	Nominal frequency band
UHF	300-1000 MHz
L	1-2 GHz
S	2-4 GHz
C	4-8 GHz
X	8-12 GHz
Ku	12-18 GHz
K	18-27 GHz
Ka	27-40 GHz
V	40-75 GHz
W	75-110 GHz

Nowadays, spectrum in Ku band and below is widely exploited and the technology has a good level of maturity while K and Ka bands are in intensive development due to their use in satellite communications. However, the maturity of the technology in Ka band and above is still low, especially if it is necessary to reduce size, weight and cost for a massive production and development. Because of that, the first objective of this thesis is the study of different technologies that can provide these characteristics.

1.3 Motivation. Problems with Classic Technologies

The most common technologies used in radiofrequency circuits are microstrip and hollow rectangular waveguide. These two technologies have been studied and implemented during decades and they present very high level of maturity. However, their principal properties present complementary and exclusive advantages and disadvantages.

1.3.1 Rectangular Waveguide

The rectangular waveguide [3][4] consists of a single hollow conductor that can propagate electromagnetic fields over a specific cut-off frequency, f_c . Rectangular waveguides can only propagate TE and TM modes, but not TEM modes since only one conductor is present. Infinite modes can be propagated, but waveguides are typically used in a single mode configuration.

TEM modes are characterised by fields with a zero value component in the direction of propagation $E_z = 0$ and $H_z = 0$, while in TE modes only the electric field in the direction of propagation becomes zero $E_z = 0$ and the magnetic field for TM modes $H_z = 0$.

The structure of a waveguide and the parameters of designs are shown in the Figure 1-2.

Figure 1-2: Structure of a rectangular waveguide.

The height of the guide, b , is usually selected to be $a/2$ to achieve a 66% transmission bandwidth on single mode configuration. Cut-off frequency is related to the width of the waveguide and can be calculated easily using the expression (1-1) for the first mode TE_{10} in a waveguide filled with dielectric of relative dielectric constant ϵ_r and a relative magnetic permeability of $\mu_r = 1$:

$$f_c(GHz) = \frac{c}{\lambda_c} = \frac{c}{2a\sqrt{\epsilon_r}} = \frac{150}{a\sqrt{\epsilon_r}}; \text{ with } a \text{ in mm} \quad (1-1)$$

The general equation (1-2) for the cut-off frequency of the infinite modes allows calculating the frequency at which each of the modes start propagating by selecting the corresponding integer values of m and n .

$$f_{c_{mn}}(GHz) = \frac{150}{\pi\sqrt{\epsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2}; \text{ with } a \text{ and } b \text{ in mm and } \mu_r = 1 \quad (1-2)$$

In Figure 1-3 it is represented the E-field (a) and H-field (b) of the six first modes in a rectangular waveguide with $a=2b$.

Figure 1-3: Six first modes in a rectangular waveguide. E-fields in (a) and H-fields in (b).

The wavelength λ_g in a lossless guide filled with dielectric can be calculated with the formula (1-3) which relates the free space wavelength λ , the wavelength in the dielectric λ_d and the cut-off wavelength λ_c .

$$\frac{1}{\lambda_d^2} = \frac{\epsilon_r}{\lambda^2} = \frac{1}{\lambda_g^2} + \frac{1}{\lambda_c^2} \quad (1-3)$$

The advantages and disadvantages of waveguides will be presented in a comparison with the microstrip line.

1.3.2 Microstrip Line

Microstrip technology [4][5] consists of the transmission of the electromagnetic field between two conductors: a metallic strip over a dielectric substrate with a ground bottom metal plate. The structure of this technology is shown in Figure 1-4.

Figure 1-4: Structure of a microstrip line.

Microstrip technology is derived from strip technology, in which the strip is completely surrounded by dielectric and a TEM mode is propagated. However, by subtracting the upper half of the dielectric, the strip is exposed to air, which reduces the effective dielectric constant, the dielectric losses, allows an easier integration of active components and keeps a quasi-TEM mode for low frequencies.

The reason for not a pure TEM mode is propagated is that the phase in the surface of the structure must match between the fields in the dielectric and the fields in the air region, but it is not possible due to the difference in the relative dielectric constant. The real mode is a hybrid TE-TM mode, but considering a quasi-static condition for low frequencies, it approximates to a TEM mode, which is not true for high frequencies. In Figure 1-5, typical electric and magnetic fields in the structure are represented with continuous lines and dashed lines respectively.

Figure 1-5: Quasi-TEM mode in microstrip.

In the next section, it is presented a comparison between waveguides and microstrip line for high frequencies.

1.3.3 Pros and Cons between Waveguide and Microstrip

As it has been said, microstrip technology presents good performance in low frequencies, but these good properties deteriorate for high frequencies. It is because the field concentrates in the region between the strip and the ground plane, which increases the effective dielectric constant and then the attenuation due to the dielectric. Apart from that, at high frequencies it is not

propagating a TEM or quasi-TEM mode and ohmic losses increase due to the higher current flow within the microstrip. These problems are solved by using hollow waveguide technology, but it supposes other problems.

The main problems that are present in waveguide technology refer to the manufacturing process. Waveguides are expensive and difficult to produce for mass applications. In addition, they are bulky and heavy and require very good metallic contact in the junctions of the walls.

A summary of the pros and cons in microstrip and waveguide is shown in Table 1-II. Each one of them, waveguide or microstrip, solves the problems that causes the other technology. Nevertheless, it is desirable to find a technology that could face all these problems and offer a solution to a cheap, mass producible technology in high frequencies. Substrate integrated waveguide and gap waveguide technologies could be the answer.

TABLE 1-II: PROS AND CONS COMPARISON BETWEEN WAVEGUIDE AND MICROSTRIP

Problems with common technologies in high frequencies		
	Microstrip	Rectangular waveguide
Compact	✓	✗
Mass producible	✓	✗
No electrical contact	✓	✗
Low dielectric losses	✗	✓
Low ohmic losses	✗	✓
No unwanted surface waves	✗	✓
Cost effective	✓	✗

1.4 Recent Waveguide Technologies

1.4.1 Substrate Integrated Waveguide (SIW)

The SIW technology involves synthesizing a substrate integrated structure with a similar waveguide behaviour. To this end, it is used a dielectric substrate where top and bottom ground plates are connected through two periodic rows of metallized via holes that behave like the walls of the guide, so that it is obtained a path in which the wave propagates. This kind of structures were first investigated in 1998 [6]. There are four structure parameters in a SIW design, which are the thickness d of the dielectric substrate with relative dielectric constant ϵ_r ; the width W of the SIW measured as the separation between the centres of the vias of the two periodic rows; the diameter D of the pins, and the periodic separation s between them [7]. A scheme of a SIW is shown in Figure 1-6.

Figure 1-6: Structure and parameters of SIW.

The width W_{eff} of the equivalent rectangular waveguide filled with the same dielectric can be calculated with the expression (1-4) from [8].

$$W_{eff} = W - 1.08 \frac{D^2}{s} + 0.1 \frac{D^2}{W} \quad (1-4)$$

In order to obtain the value of the SIW width W from the equivalent rectangular waveguide W_{eff} it is necessary to select the values for s/D and D/W first and then solve the equation (1-5).

$$W = \frac{W_{eff}}{1 - 1.08 \frac{D/W}{s/D} + 0.1 \cdot (D/W)^2} \quad (1-5)$$

The accuracy of expressions (1-4) and (1-5) is very good for $s/D < 3$ and $D/W < 1/5$.

To obtain the width W_{air} of the equivalent hollow air-filled rectangular waveguide it is necessary to apply the relation (1-6).

$$W_{air} = W_{eff} \cdot \sqrt{\epsilon_r} \quad (1-6)$$

By selecting the adequate parameters, it is possible to minimize the losses and obtain a very close result to the equivalent dielectric-filled waveguide but with a reduced cost and manufacturing effort. A design procedure for SIW can be found at [8] and some tips to reduce the losses are exposed in [9]. There are three parameters that have an important impact on the losses, which are the thickness of the substrate d , the diameter of via holes D , and the periodic separation s between them. As it is explained in [9], the increase of d supposes a reduction of the conductor losses due to the reduction of the current density J on the metal plates, while the increase of D and s means an increment of the leakage losses, which are negligible with a good choice of D and s . However, dielectric losses have a major impact for the contribution of total losses and cannot be reduced by varying the structure parameters, so that a careful selection of the dielectric substrate is extremely important.

In SIW, conductor losses are lower than in microstrip technology due to the currents flow within wider metal surfaces. Regarding to the dielectric losses, at high frequencies losses between SIW and microstrip are comparable because in both cases almost all the energy propagates within the dielectric.

Compared to hollow rectangular waveguides, SIW presents higher losses due mainly to the dielectric substrate. However, SIW presents better properties in relation to the manufacturing process.

1.4.2 Gap Waveguide (GW)

The gap waveguide is a new technology recently introduced [10] that tries to solve the dielectric losses problems that occur at high frequencies in technologies like SIW due to the wave propagation within a dielectric substrate. This is because in gap waveguide the field is propagated through an air gap between parallel metal plates, so it supposes low dielectric losses. In addition, gap waveguide does not require good contact between the top and bottom metal plates because the lateral walls that exist in classic waveguide technology are substituted by a metamaterial layer that blocks the propagation in the lateral directions. An overview of gap waveguide, its current development and future applications in mm-wave systems can be found at [11]. For these reasons, gap waveguide is a good alternative to other substrate technologies and hollow waveguide.

The basis of gap waveguide technology is the electric conductor surface over magnetic conductor surface configuration. In a perfect electric conductor (PEC) surface, the electric field is characterized to be zero in the tangential plane to the surface, while in a perfect magnetic conductor (PMC) surface the tangential magnetic field becomes zero. Then, if these two surfaces are close together, less than a quarter wavelength, both electric and magnetic fields become zero and propagation is not allowed. This effect is represented in Figure 1-7.

Figure 1-7: Propagation forbidden between two parallel PEC-PMC surfaces close together.

The problem is that magnetic conductors are only a mathematical concept and do not exist in real world. Because of that, the main characteristic of this technology is the use of a metamaterial surface, which is implemented with a textured structure of periodic metallic pins or

via holes separated a certain distance. This creates a high impedance surface at certain frequency band that is called artificial magnetic conductor (AMC). These artificial surfaces are based on translating the impedance in a short circuit to an open circuit through a quarter wavelength periodically across the surface. The operating principle is the same used for corrugations in horns and it is represented in Figure 1-8.

Figure 1-8: AMC structure scheme.

Four design parameters are involved in the response of the structure, which are the diameter D of the pins, the periodic separation p between them, the height d of the pins or dielectric substrate in which via holes are drilled, and the relative dielectric constant ϵ_r of the material in which the metal pins are embedded. A scheme of the structure is represented in Figure 1-9.

Figure 1-9: Parameters of AMC metamaterial.

The desired PEC over PMC structure is obtained by placing a metal plate above the AMC surface at a distance h . This distance of separation or gap between surfaces is an important design parameter as it will be explained later, and it gives the name to gap waveguide technology. Finally, five parameters are involved in the design of the metamaterial and the frequency band gap in which propagation is not allowed will depend on them. The final metamaterial structure that will substitute the waveguide walls is shown in Figure 1-10.

Figure 1-10: Electric conductor over AMC metamaterial.

There are two types of main possible structures for gap waveguide of which at least four configurations are available [12]. These structures are called groove gap waveguide and ridge gap waveguide, which can be converted to inverted microstrip gap waveguide and microstrip-ridge gap waveguide. They are explained below.

1.4.2.1 Ridge Gap Waveguide (RGW)

Ridge Gap Waveguide is based on ridge waveguides on which the sidewalls are replaced by the metamaterial surface. This surface prevents the transmission of power in the transverse direction and confines the wave above the central ridge [13][14].

The structure of the RGW is represented in Figure 1-11 and a transversal view is compared with the equivalent single ridge waveguide in Figure 1-12.

Figure 1-11: Structure of ridge gap waveguide.

Figure 1-12: RGW (left) and equivalent ridge waveguide (right).

The width W_{eff} of the equivalent ridge waveguide mainly depends on the form of the pins (cylinders, parallelepipeds...) and their orientation. Ridge waveguides present infinite propagating modes as well as rectangular waveguides, whose cut-off frequency can be calculated theoretically. The cut-off frequencies for TE_{10} , TE_{20} and TE_{30} modes in ridge waveguides can be obtained from the curves in [15] and a more detailed analysis for TE_{10} mode in [17]. The cut-off frequency of the fundamental mode of the equivalent ridge waveguide must be lower than the band gap generated by the metamaterial in order to that mode can be propagated through the ridge when the metamaterial blocks its propagation in the lateral directions.

The ridge can be implemented using a microstrip line within a substrate. Two different structures are possible depending on the metamaterial configuration. These structures are called inverted microstrip gap waveguide and microstrip-ridge gap waveguide.

1.4.2.2 Inverted Microstrip Gap Waveguide (IMGW)

In inverted microstrip gap waveguide, the ridge is substituted by a microstrip line printed in a dielectric substrate that is placed over a bed of nails. This avoids the propagation in the longitudinal directions and in the space below the microstrip, concentrating the fields in the air gap between the strip conductor and the upper metal plate. Figure 1-13 represents the structure of this configuration.

Figure 1-13: Structure of inverted microstrip gap waveguide (top) and transversal view (bottom).

1.4.2.3 Microstrip-Ridge Gap Waveguide (MRGW)

Microstrip-ridge gap waveguide consists of the integration of the metamaterial in the dielectric substrate. The bed of nails is replaced by a periodic structure of via holes within the substrate that surrounds the microstrip. The microstrip is also drilled and connected with the lower conductor through small metallized vias to block the propagation of fields under the metallic strip. The schematic structure for this configuration is shown in Figure 1-14.

In this configuration, the size of the via holes must be a quarter of the wavelength in the dielectric substrate, reducing the thickness of the structure. Via holes in dielectric substrate are usually implemented with a mushroom patch that must be taken in account in the designs due to this form modifies the band gap behaviour.

Figure 1-14: Structure of MRGW (top) and transversal view (bottom).

1.4.2.4 Groove Gap Waveguide (GGW)

In groove gap waveguide the central ridge is deleted, having an emulated air-filled rectangular waveguide with the benefit of having contactless metal plates due to lateral metamaterial walls. In addition, groove gap waveguide has lower losses than ridge gap waveguide because current density is lower due to removing the ridge increments the volume for the propagating fields and then ohmic losses are reduced. The structure of GGW is shown in Figure 1-15 and a transversal section is compared with the equivalent rectangular waveguide in Figure 1-16.

Figure 1-15: Structure of groove gap waveguide.

Figure 1-16: GGW (left) and equivalent rectangular waveguide (right).

As well as for RGW, the cut-off frequency of the equivalent rectangular waveguide must be lower than the band gap provided by the metamaterial walls. This cut-off frequency depends on the width W_{eff} of the rectangular waveguide in Figure 1-16 that in turn depends on the form of the pins. An exhaustive analysis of groove gap waveguide was carried out in [16].

1.4.3 Benefits of SIW and Gap Waveguide over Microstrip and Waveguide

SIW and gap waveguide technologies present several advantages compared to traditional microstrip and rectangular waveguide technologies. These advantages are focused on the reduction of losses and low cost and ease of manufacturing process. SIW solves the problems with weight and volume of rectangular waveguides and the high ohmic losses and surfaces propagating modes in microstrip, although presents problems with dielectric losses. Gap Waveguide deals with this problem by propagating fields through an air gap between contactless metal surfaces. However, the need for at least two or three rows of periodic pins or via holes for the lateral metamaterial supposes an increment of the surface needed in gap technology, in special for GGW. The main advantages and disadvantages between technologies are summarized in Table 1-III.

TABLE 1-III: COMPARISON BETWEEN TECHNOLOGIES UNDER STUDY AND CLASSICAL TECHNOLOGIES

Pros and cons in microstrip, rectangular waveguide, SIW, RGW and GGW					
	MS	RW	SIW	RGW	GGW
Compact	✓	✗	✓	✓	✗
Mass producible	✓	✗	✓	✓	✓
No electrical contact	✓	✗	✗	✓	✓
Low dielectric losses	✗	✓	✗	✓	✓
Low ohmic losses	✗	✓	✓	✗	✓
No unwanted surface waves	✗	✓	✓	✓	✓
Cost effective	✓	✗	✓	✓ ✗	✓ ✗

1.5 Additive Manufacturing

3D printing technologies, also known as additive manufacturing (AM), have evolved rapidly in recent years and are currently used in a wide range of engineering applications. In particular, the use of the new manufacturing techniques is leading to the development of very promising 3D electromagnetic structures for millimetre-wave devices [18]. Thanks to the very competitive performance of 3D printing in electromagnetic structures, both the industrial and research communities are paying special attention to this manufacturing technique.

3D printing is a novel fabrication process for polymer, metal, and ceramic materials. This technique is an attractive option for innovative and complex part fabrication, providing improvements in terms of customization, part compactness, rapid prototyping, reduced weight, cost and mass production lead time. In the literature, there are a variety of recent publications on waveguides [19]-[23], antennas [24]-[30], filter components [31] and other passive devices [32]-[36] over a wide range of frequencies with different 3D fabrication techniques. Most common 3D-printing techniques use plastics or ceramics, such as stereolithography apparatus (SLA) [19],[21],[22],[24],[31], fused deposition modelling (FDM) [19], selective laser sintering (SLS) [29],[30] and Polyjet [32]-[35].

The FDM technique is the best known and most economical, but the resulting pieces have a finish that is not acceptable for use in millimetre wave bands due to the appearance of very marked grooves produced by the deposition of the plastic filament. The other techniques based on plastic or ceramics generally provide a good finish and a high level of detail in the printed structures. SLA is based on curing photoresins with UV light by means of a laser that sweeps the desired surface layer by layer. A low-cost variant of SLA is called Digital Light Processing (DLP), which instead of using a laser beam that sweeps the desired surface, uses an LED panel that simultaneously illuminates the entire surface of the corresponding layer. In the case of SLS, a laser is also used to heat and bond small plastic particles so that a rigid structure can be sintered. The Polyjet technique uses the traditional concept of inkjet printing, in which the ink is replaced by a polymer that is cured with UV light when it meets the desired surface. These processes then need to be metal coated for 3D electromagnetic structure parts. A typical material used for metal-

coating is copper due to its easier application process compared with other metals and good electrical conductivity.

Alternatively, direct 3D metal printing can be used, with processes such as binder jetting (BJ) [27],[28],[36], metal selective laser melting (SLM) [23],[28] and direct metal laser sintering (DMLS) [22],[25],[26] in stainless steel, aluminium, titanium, nickel or copper. The BJ technique works in a similar way to Polyjet, dispensing an adhesive substance to bind metal particles, therefore it does not require high temperature conditions. In the cases of SLM and DMLS, high temperatures are required during the building of the parts for the union of metallic particles. The difference between both technologies lies in the temperature applied. For DMLS, a laser beam scans the desired surface to bond each layer of the object. This process is performed in a chamber that is maintained at a temperature just below the melting temperature so that the passage of the laser raises the temperature enough to bond the metal particles. For SLM, the temperature must completely melt the particles to result in a robust and dense metal structure without porosity. Some studies analysing the structural differences between SLM and DMLS can be found in [37][38].

Each of the above technologies, based on both plastics and ceramics as well as metals, have advantages and disadvantages that have different impacts depending on the application. All solutions based on metal printing provide great robustness but have a rough finish due to the size of the metal particles, which has a very negative impact on losses in high frequency devices. However, it is possible to achieve a smoother surface finish by applying expensive post-processing. Techniques in plastic or ceramics provide a high resolution except FDM, the low resolution of which is not recommended for high frequency applications. Polyjet generates surfaces with an appreciable roughness due to the accumulation of small resin droplets produced by the jetting process. In SLS there is also a high level of roughness due to the size of the sintered plastic particles. Finally, SLA provides the best result in both resolution and smoothness of surfaces, although shrinkage and bending problems can occur that can be compensated for during the manufacturing process.

Figure 1-17: 3D-printing processes: FDM, SLS, DMLS, SLA, DLP, Polyjet. Images: www.3dhubs.com

Figure 1-17 shows a scheme of the 3D-printing process using the most common techniques. The rise of 3D printing and continuous development of the necessary technology has boosted the manufacturing of various professional designs for both prototyping and mass production. Stimulation of this technology has had a major impact on much sectors of industry, including the field of radiocommunications and antenna design. Accompanying 3D printing, it has developed the necessary technology for metallizing models printed in various plastic materials. This provides great advantages over existing metal machining technology, which are: ease of manufacture, reduced costs, reduced lead times, reduced weight, increase in RF design flexibility and plastic provides better thermal insulation than metal.

1.6 Liquid Crystals (LC)

One of the most promising materials for electronically reconfigurable millimetre-wave devices is liquid crystal. This material is widely used in photonics for LCD (Liquid Crystal Display) because its interaction with light can be controlled by applying certain voltages. LC are used under the same operating principle in polarizers. These liquid crystals can also change their properties as a function of temperature, so they are used in various temperature sensors. Apart from this ability to modify their properties through anisotropy, there are other applications of liquid crystals in industry (Kevlar manufacturing) and even in biology (phospholipids, a type of LC, are the main component of the cell membrane). Some examples of liquid crystal devices and applications are shown in Figure 1-18.

Figure 1-18: Common devices based on liquid crystal: (a-b) LCD display, (c) cell membrane, (d) temperature sensor.

Variation of the orientation of liquid crystal molecules in the presence of an electric field can modify the effective dielectric constant of the material. Some work is already done to exploit this characteristic of liquid crystal to achieve millimetre-wave antennas with reconfigurable beam steering [39].

In recent years, liquid crystal, whose properties and uses are already widespread in the photonics field, has been introduced and developed for use at microwave frequencies. This development is focused on improving the properties at these much lower frequencies. The main ones are the increase of the dielectric anisotropy and the reduction of the loss tangent [40]. The ability to modify the dielectric properties of the material is due to molecular reorganization in the presence of a quasi-static electric field. As shown in Figure 1-19, the main axis of the molecules tends to be oriented along the electric field lines, which modifies their properties with respect to the non-excited state. To characterize this behaviour, two dielectric constants are defined: $\epsilon_{//}$ and $\epsilon_{\perp}$. The parallel dielectric constant $\epsilon_{//}$ is considered when the electric field of the propagating mode inside the structure is parallel to the molecules of the liquid crystal. In the case of the perpendicular dielectric constant $\epsilon_{\perp}$, it occurs when this electric field is perpendicular to the molecules of the medium. An analytical method for obtaining the mean tilt angle of the molecules and the effective dielectric constant as a function of the voltage is developed in [41].

Figure 1-19: Orientation of the liquid crystal molecules depending on the voltage. Tilting of the molecules from left to right when the voltage increases from 0 V to $V_{\max}$.

Due to the anisotropic property of liquid crystal, its dielectric constant must be defined as a tensor. This tensor is calculated generically as the expression (1-7) indicates, where $\Delta\epsilon$ is the anisotropy of the dielectric constant ($\epsilon_{\perp}-\epsilon_{//}$) and θ_m is the mean angle of the liquid crystal molecules, which depends on the applied voltage.

$$\bar{\epsilon} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} + \Delta\epsilon \cos^2(\theta_m) & \Delta\epsilon \sin(\theta_m) \cos(\theta_m) \\ 0 & \Delta\epsilon \sin(\theta_m) \cos(\theta_m) & \epsilon_{\perp} + \Delta\epsilon \sin^2(\theta_m) \end{pmatrix} \quad (1-7)$$

When the applied voltage is 0, the molecules tend to align parallel to the parallel plates containing the liquid crystal. Assuming a very small mean angle, the tensor can approximate the expression (1-8). Similarly, when the applied voltage is high, the molecules tend to align in the direction of the electric field generated between the parallel plates. In saturation conditions we can approximate the tensor to the expression (1-9) assuming θ_m close to 90° . As the propagation modes used are typically TEM or TE_{10} modes, the electric field is oriented perpendicular to the parallel plates, the effective dielectric constant will be $\epsilon_{\perp}$ for $V=0$ and $\epsilon_{//}$ for $V_{\max}$.

$$\bar{\epsilon}_{V=0} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} + \Delta\epsilon & 0 \\ 0 & 0 & \epsilon_{\perp} \end{pmatrix} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{//} & 0 \\ 0 & 0 & \epsilon_{\perp} \end{pmatrix} \quad (1-8)$$

$$\bar{\epsilon}_{V \gg} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} & 0 \\ 0 & 0 & \epsilon_{\perp} + \Delta\epsilon \end{pmatrix} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} & 0 \\ 0 & 0 & \epsilon_{//} \end{pmatrix} \quad (1-9)$$

1.7 Higher symmetries

The recent study of higher symmetries [42]-[46] has been driven by a growing interest in the use of periodic structures to improve the electromagnetic properties of antennas and microwave devices. These symmetries were first investigated in the 1960s and 1970s for one-dimensional periodic structures [47]-[49]. More recently, two-dimensional glide symmetries have been proposed and studied, which are a particular case of higher symmetries, demonstrating great potential for modifying the dispersion properties of periodic structures. Periodic glide-symmetric structures are obtained by reflecting and translating a semi-period the unit cell. Figure 1-20 shows an example of how this glide symmetry works.

Figure 1-20: Glide symmetry operation: (a) basic unit cell, (b) mirroring of the unit cell and (c) translation of the structure half a period.

Glide symmetries have been successfully used to reduce the dispersion of periodic structures [50]-[53], to increase the equivalent refractive index [54]-[57], or to increase the width and attenuation of electromagnetic bandgaps [58]-[61]. For example, glide symmetry has been proposed to produce lens antennas for 5G communications [62][63] taking advantage of their ability to produce a higher refractive index, less dispersion and more isotropy [64].

1.8 Objectives

The main objective of this thesis is to evaluate different novel design and manufacturing technologies applied to real prototypes in millimetre-wave bands. The developed proposals have in common the simplification of the manufacturing and assembly process, while trying to

conserve a reduced manufacturing cost. It is also intended to seek alternatives for electrical or mechanical tuning of the phase, applicable to variable beam antennas. The specific purposes of this work are enumerated in the following points:

- To verify the suitability of additive manufacturing applied to prototypes for millimetre-wave antennas. This validation will be carried out on Ka-band and W-band prototypes. The adequacy of this manufacturing method lies in the simplification of the design, manufacturing and assembly processes, as well as in the reduction in cost. 3D-printing will be applied mainly to prototypes based on gap waveguide, in which periodic metal pins represent a high cost with traditional manufacturing methods.
- Integration of gap waveguide technology and additive manufacturing for the prototyping of a monopulse antenna with simultaneous sum and difference diagrams at 94 GHz. Specifically, the design of a Butler matrix with output ports coupled to a Radial Line Slot Array (RLSA) antenna for the generation of the two desired diagrams.
- Design, manufacturing and measurement of an electrically tuneable phase shifter based on liquid crystal at W band. For this purpose, a theoretical analysis of the liquid crystal will be carried out to estimate its phase shift capacity as a function of the applied voltage.
- Investigation of the capacities of reconfiguration of the electrical properties in structures with glide symmetry. Two different designs are proposed, one on a dielectric substrate and the other on a fully metallic waveguide. The design and manufacturing processes will be oriented to a mechanical tuning of the attenuation and phase shift based on the longitudinal displacement of a part of the prototype.
- Reduce the problem of space occupied by periodic pins in gap waveguide to minimize the separation between parallel waveguides. The design of the prototypes will be based on thin metal stacked layers with ellipses in a glide-symmetric configuration to reduce the space occupied in one dimension. This will allow small separations between parallel waveguides in resonant arrays, which will avoid problems of grating lobes.
- Increase the range of steering angles on leaky wave antennas using protrusions with glide-symmetric configuration. This implies the need of a much smaller bandwidth to cover the same range of pointing angles. The final prototypes will be based on gap waveguide with elements with glide symmetry around 80 GHz and will be printed in 3D. The results are of interest to the automotive sector, with radars that require less bandwidth in order to scan the entire angle of view.

Chapter II

Study and Comparison of Waveguide Technologies and Transmission Lines

2.1 Introduction

This chapter presents a study of recent waveguide technologies such as substrate integrated waveguide and gap waveguide with design suggestions and examples. It also includes a comparison in terms of transmission losses between these two technologies and classic technologies such as rectangular and ridge waveguide and microstrip. The results of this comparison will be useful to select the most suitable technology for designs made in later chapters.

2.2 Analysis of Substrate Integrated Waveguide (SIW)

SIW presents a simple design process based on the use of an empirical formula that relates the width W of the SIW, the diameter D of the via holes and the periodic separation s between them with the equivalent rectangular waveguide. In this section a SIW has been designed using a substrate of relative dielectric constant $\epsilon_r = 3$ and a cut-off frequency $f_c = 59.015$ GHz. This is the cut-off frequency of the standard WR-10, which covers an operational bandwidth from 75 GHz to 110 GHz. The obtained results will be compared with the response of the equivalent dielectric-filled rectangular waveguide. Figure 2-1 represents the phase constant β of the first two modes propagating in a WR-10 waveguide and the equivalent dielectric-filled waveguide with a dielectric constant $\epsilon_r = 3$. The physical dimensions of the dielectric-filled waveguide are obtained by scaling the WR-10 by a factor $1/\sqrt{\epsilon_r}$. The cut-off frequencies are the same in both waveguides, but the value of the phase constant is scaled by a factor $\sqrt{\epsilon_r}$.

Figure 2-1: Phase constant of the rectangular waveguides WR-10 and the equivalent waveguide filled with a substrate with dielectric constant $\epsilon_r = 3$.

The design procedure available in [8] has been used taking into account the considerations for reducing losses in [9]. The calculations in each step are shown in Table 2-I.

TABLE 2-I: PROCEDURE FOR SIW DESIGN

Step	Description	Initial value	Final value (fit)
Step 1	Select $f_c = 59.015$ GHz	$f_c = 59.616$ GHz	$f_c = 59.015$ GHz
Step 2	Select dielectric substrate (dielectric constant and thickness $d \leq W_{eff}/2$)	$\epsilon_r = 3$; $d = 0.733$ mm	$\epsilon_r = 3$; $d = 0.733$ mm
Step 3	Select $D/W \leq 1/8$	$D/W = 1/8 = 0.125$	$D/W = 0.126$
Step 4	Select $s/D \leq 2.5$	$s/D = 2$	$s/D = 2$
Step 5	Calculate W_{eff} by solving (1-1) with $a = W_{eff}$	$W_{eff} = 1.466$ mm	$W_{eff} = 1.466$ mm
Step 6	Calculate W by solving (1-5)	$W = 1.57$ mm	$W = 1.586$ mm
Step 7	Calculate D and s from D/W and s/D	$D = 0.196$ mm; $s = 0.392$	$D = 0.2$ mm; $s = 0.4$ mm

The SIW has been modelled in CST Microwave Studio using the design parameters in Table 2-I. The dispersion diagram has been calculated and the phase constant for the first four modes is represented in Figure 2-2, compared with the first four modes of the equivalent dielectric-filled rectangular waveguide. This initial design results in the propagation of modes whose phase constants do not exactly coincide with those of the equivalent waveguide presented in the figure (WR $\epsilon_r = 3$). Therefore, the initial parameters have been adjusted to achieve a response as equivalent as possible to that of the dielectric-filled rectangular waveguide.

Figure 2-2: Phase constant for the first four modes in the initial SIW design and the equivalent rectangular waveguide (WR).

Figure 2-3 depicts a zoom of the fundamental mode (Figure 2-3(a)) and the following two modes (Figure 2-3(b)) of the initial and final design of the SIW and the equivalent waveguide. The cut-off frequency of the fundamental mode in the initial design is slightly displaced up in frequency, as well as the third mode. However, the cut-off frequency of the second mode in SIW matches well with the second mode in the equivalent rectangular waveguide. This is because the

first and third modes are TE_{10} and TE_{20} modes, whose cut-off frequencies depend on the width of the structure, which is not exactly defined in SIW. However, the second mode is TE_{01} , whose cut-off frequency depends on the height of the waveguide and it is the same than for the rectangular waveguide. The cut-off frequency error for TE_{10} and TE_{20} can be reduced by enlarging the width of the SIW by a factor resulting from divide the cut-off frequency obtained and the cut-off frequency desired. The cut-off frequency of the TE_{10} mode for this design is $f_{c-error} = 59.616$ GHz, while the desired cut-off frequency is $f_c = 59.015$ GHz. If a correction factor is applied, given by $f_{c-error}/f_c = 1.0102$, it is obtained a corrected value for the SIW width, $W_{corr} = W \cdot 1.0102 = 1.586$ mm, instead of 1.57 mm. After the correction, the phase constant of the TE_{10} mode of the final SIW design is adjusted precisely to that of the equivalent rectangular waveguide. The TE_{01} mode of both the initial and the final design hardly varies its behaviour, as it mainly depends on the height of the SIW which remains the same. Furthermore, they match well with the same mode of the equivalent rectangular waveguide. However, the TE_{20} mode does not approximate as precisely to that of the equivalent waveguide. Nevertheless, the cut-off frequency for the final SIW design is above the cut-off frequency of the TE_{01} mode, so the operating band is not diminished.

Figure 2-3: Phase constant comparison for the first mode (a) and for the second and third modes (b).

The field lines of the generated modes are slightly distorted in SIW compared to those generated in a classical rectangular waveguide. This distortion is due to the fact that the walls are not continuous but are replaced by periodic metallic paths. This effect can be seen in Figure 2-4 and Figure 2-5 for the first six modes. However, the simulated field lines are similar to the theoretical field lines for the first six modes of the equivalent rectangular waveguide of Figure 1-3 in section 1.3.1.

Figure 2-4: E-field lines in six first modes at SIW.

Figure 2-5: H-field lines in six first modes at SIW.

2.3 Gap Waveguide Analysis

The design process for RGW or GGW is more laborious than in SIW because it is necessary to define the bandgap of the gap waveguide walls in first place and then include in the design the ridge structure or the groove structure. Because of that, an analysis of the metal plate over AMC surfaces has been conducted in order to determine the impact of the design parameters of gap waveguide in the resulting bandgap.

2.3.1 PEC over AMC Analysis

As it was explained in Chapter I, gap waveguide presents five main parameters of design, which are the height of the metal pins d , their diameter D , the air gap h between AMC and PEC surfaces, the periodic separation p between pins and the relative dielectric constant ϵ_r of the substrate in which pins are embedded. The substrate is usually air, but for IMGW or MRGW, the relative dielectric constant of the material is important.

The simulated model is shown in Figure 2-6, where only a unit cell of the periodic structure is represented. The selected parameters of the unit cell are in Table 2-II. CST Microwave Studio allows simulating periodic structures in a very simple way, which consists of modelling a single unit cell and expanding it periodically using periodic boundaries. In this configuration, it is possible to obtain dispersion diagrams and the cut-off frequencies of the propagating modes, as well as the phase constant as a function of the frequency. These modes are calculated with the Eigenmode solver. The bandgap in a periodic unit cell corresponds to the frequency bands where no mode propagates. To correctly study these modes to ensure that there is no propagation on a certain frequency band, the first Brillouin zone is defined. The first Brillouin zone is calculated as a Fourier transformation of the real unit cell into a reciprocal unit cell. Taking the primitive lattice vectors $\{\mathbf{a}_1, \mathbf{a}_2, \mathbf{a}_3\}$ for a real unit cell, the reciprocal vector set $\{\mathbf{b}_1, \mathbf{b}_2, \mathbf{b}_3\}$ is calculated as (2-1).

$$\mathbf{b}_1 = \frac{2\pi\mathbf{a}_2 \times \mathbf{a}_3}{\mathbf{a}_1 \cdot (\mathbf{a}_2 \times \mathbf{a}_3)}, \quad \mathbf{b}_2 = \frac{2\pi\mathbf{a}_3 \times \mathbf{a}_1}{\mathbf{a}_1 \cdot (\mathbf{a}_2 \times \mathbf{a}_3)}, \quad \mathbf{b}_3 = \frac{2\pi\mathbf{a}_1 \times \mathbf{a}_2}{\mathbf{a}_1 \cdot (\mathbf{a}_2 \times \mathbf{a}_3)} \quad (2-1)$$

As an example, for a unit cell with periodicity p in two dimensions (typically x -direction and y -direction) the primitive lattice vectors can be defined as $\mathbf{a}_1 = p\hat{\mathbf{x}}$, $\mathbf{a}_2 = p\hat{\mathbf{y}}$ and $\mathbf{a}_3 = \hat{\mathbf{z}}$. The corresponding reciprocal vectors $\{\mathbf{b}_1, \mathbf{b}_2, \mathbf{b}_3\}$ inside the first Brillouin zone are $\mathbf{b}_1 = (2\pi/p)\hat{\mathbf{x}}$, $\mathbf{b}_2 = (2\pi/p)\hat{\mathbf{y}}$ and $\mathbf{b}_3 = 2\pi\hat{\mathbf{z}}$ in the reciprocal space. Furthermore, when a periodic electromagnetic structure has a rotation, mirror reflection or inversion symmetry, we need not consider all propagating modes in the Brillouin zone. The smallest region within the Brillouin zone for which propagating modes are not related by symmetry is called the irreducible Brillouin zone [65]. The

complete Brillouin zone is made up of redundant copies of the irreducible zone. In most of the cases of interest, the minimum and maximum of the bandgap occurs in the edges of the irreducible Brillouin zone and sometimes in the corners. For that reason, typically only the reciprocal vectors contained in these edges are calculated.

Figure 2-6: Periodic unit cell.

TABLE 2-II: UNIT CELL PARAMETERS

Unit cell parameters	
Initial bandgap frequency f_{c1}	$f_{c1} = 65$ GHz
Final bandgap frequency f_{c2}	$f_{c2} = 115$ GHz
Dielectric constant ϵ_r	$\epsilon_r = 1$
$\lambda_{c1}/5 \leq d \leq \lambda_{c1}/4$	$d = \lambda_{c1}/5 \approx 0.9$ mm
$0.45 \cdot \lambda_{c2} \leq d + h \leq 0.5 \cdot \lambda_{c2}$	$h = 0.45 \cdot \lambda_{c2} - d \approx 0.3$ mm
$1 \leq d/D \leq 3$	$d/D = 1.8$
$1.5 \leq p/D \leq 3$	$p/D = 2.4$
D	$D = 0.5$ mm
p	$p = 1.2$ mm

Before presenting the analysis of the dispersion diagram of the unit cell with the parameters indicated in Table 2-II, we show an example of the main types of boundary conditions available in CST Microwave Studio, as well as the definition of the irreducible zone of Brillouin in a periodic unit cell. The unit cell with the parameters from Table 2-II is used for the example. Figure 2-7 shows the unit cell model with periodic boundary conditions in x -direction and y -direction and electrical boundary conditions in the z -direction. These periodic conditions are represented on each face of the parallelepiped containing the 3D model with a different shape, color and letter. Figure 2-8 shows the four most commonly used boundary conditions in the development of this document. The first of them (Figure 2-8(a)) is the periodic condition, represented in yellow with a P. This boundary condition states that the electric and magnetic field conditions at the boundary are identical to the conditions at the opposite boundary, except for the phase, which can be modified by a parameter. In this way, it is possible to define unit cells with periodic conditions whose phase difference between boundaries in a given direction varies. This allows the computation of the modes that can be excited in the structure with the given phase difference and the calculation of the frequency at which they propagate. This is very useful for extracting the phase constant of the modes propagating in a waveguide or in a transmission line in general. In addition, as will be seen below, it allows scanning the edges of the irreducible Brillouin zone of a periodic cell to calculate in which frequency band no mode propagates and to estimate the bandgap of the structure.

The electrical boundary condition is represented in Figure 2-8(b) by a green symbol and a letter E. This boundary condition is equivalent to placing a PEC material sheet in the corresponding boundary since it sets a condition for the electric field tangential to the surface $E_t=0$. The same applies to the magnetic boundary condition, which is shown in blue in Figure 2-8(c) with the letter M. In this case, it is equivalent to replacing this boundary condition with a PMC material surface, forcing the magnetic field tangential to the surface to be zero, $H_t=0$.

Finally, the open boundary illustrated in Figure 2-8(d) in pink with a letter O defines a perfect absorbent material that allows waves to pass through with minimal reflection. This is equivalent to virtually extending to infinity all the structures and materials existing in the plane in which the open boundary is defined. If all the material in contact with the open boundary is air (or vacuum), it can be considered as a free space condition (Open (add space)) widely used in antenna models to calculate the radiated fields and obtain the radiation patterns.

Figure 2-7: Periodic unit cell based on the parameters of Table 2-II. (a) Perspective view with boundary conditions and (b) top view.

Figure 2-8: Representations of boundary conditions in CST Microwave Studio. (a) Periodic boundary, (b) electric boundary, (c) magnetic boundary and (d) open boundary.

As an example for the definition of the irreducible Brillouin zone, we take the unit cell with the pin of Figure 2-7 and a unit cell of the same size and with the same boundary conditions filled with air only. In Figure 2-9, the empty unit cell (Figure 2-9(a)) is shown next to the representations of the reciprocal space of this unit cell (Figure 2-9(b)) and the unit of interest cell with the metal pin (Figure 2-9(c)). In the reciprocal space, the representations of the irreducible Brillouin zones are included. Due to the rotation and mirror-reflection symmetry of both unit cells, the irreducible Brillouin zone is contained in one eighth of the total area occupied in the XY plane. Therefore, the irreducible Brillouin zone is defined by three straight sections connecting the Γ , X and M points. The point Γ represents the center of the reciprocal unit cell ($x = 0 \text{ rad/m}$, $y = 0 \text{ rad/m}$), the point X is the central point of the lateral side of the cell ($x = \pi/p \text{ rad/m}$, $y = 0 \text{ rad/m}$) and the point M is the corner of the reciprocal unit cell ($x = \pi/p \text{ rad/m}$, $y = \pi/p \text{ rad/m}$). The Eigenmode solver in CST Microwave Studio allows travelling along the edge of the irreducible Brillouin zone by calculating the modes that propagate in the structure for various phase difference conditions between opposite periodic boundaries of the unit cell. Applying the operation $x' = px$ and $y' = py$, the Γ , X and M points are equivalent to $(x' = 0 \text{ rad}$, $y' = 0 \text{ rad})$, $(x' = \pi \text{ rad}$, $y' = 0 \text{ rad})$ and $(x' = \pi \text{ rad}$, $y' = \pi \text{ rad})$, respectively. In this way, we distinguish three sections: in the first section, from Γ to X, we calculate the modes that propagate by varying the phase difference x' between opposite periodic boundaries in the x -direction from 0 to π radians, or 180 degrees, keeping the phase difference between boundaries in the y -direction $y' = 0$. In the second section, from X to M, we keep the phase difference in the x -direction $x' = 180^\circ$ and vary the phase difference in the y -direction from 0° to 180° . Finally, in the third section, from M to Γ , the phase differences in the x -direction and in the y -direction vary simultaneously from 180° to 0° . For clarity in representing the results, we normalize the phase values with respect to π and represent the three sections as the

normalized values $\beta p/\pi$, where β is the phase constant, from 0 to 1 for the first section ($\beta p_1/\pi$), from 1 to 2 for the second section ($\beta p_2/\pi$) and from 2 to 3 for the third section ($\beta p_3/\pi$). The values p_1 , p_2 and p_3 correspond to the periodicity in the direction of each section of the irreducible Brillouin zone. In this case, $p_1 = p_2 = p$ and $p_3 = p\sqrt{2}$. To visualize the effect of the variation of the phase difference between opposite periodic boundaries, the amplitude of the electric field of the first mode that propagates in the empty unit cell at the three central points in each section of the irreducible Brillouin zone is represented in Figure 2-10. The phase of the electric field in the z component is illustrated in Figure 2-11 for the same three points. The first point (Figure 2-10(a) and Figure 2-11(a)) corresponds to a phase difference condition ($x' = 90^\circ$, $y' = 0^\circ$), the second point (Figure 2-10(b) and Figure 2-11(b)) to the conditions ($x' = 180^\circ$, $y' = 90^\circ$) and the third point (Figure 2-10(c) and Figure 2-11(c)) to ($x' = 90^\circ$, $y' = 90^\circ$). For the first point, a variation of the amplitude and phase of a plane wave in the x -direction is clearly visible. The phase difference between opposite boundaries in this direction is 90° . At the second point, there is an amplitude and phase variation of 90° in the y -direction. However, the wave is not plane because a phase difference of 180° is imposed between opposite boundaries in the x -direction. For the third point, there is again a plane wave in the direction of the diagonal of the unit cell with a phase difference between the lower left and upper right corners of 90° . The plane wave is recovered in the direction of the diagonal when the phase differences in x and y -directions take the same value.

Figure 2-9: (a) Empty unit cell with periodic boundary conditions in x and y -directions and electric conditions in z -direction. XY planes of the reciprocal unit cell representations of the empty unit cell (b) and of the gap waveguide pin (c) with the irreducible Brillouin zone.

Figure 2-10: E-field amplitude for the central point in each section of the irreducible Brillouin zone for the empty unit cell. (a) $\beta p/\pi = 0.5$ ($x' = 90^\circ$, $y' = 0^\circ$), (b) $\beta p/\pi = 1.5$ ($x' = 180^\circ$, $y' = 90^\circ$), (c) $\beta p/\pi = 2.5$ ($x' = 90^\circ$, $y' = 90^\circ$).

Figure 2-11: Phase of the z -component of the E-field for the central point in each section of the irreducible Brillouin zone for the empty unit cell. (a) $\beta p/\pi = 0.5$ ($x' = 90^\circ$, $y' = 0^\circ$), (a) $\beta p/\pi = 1.5$ ($x' = 180^\circ$, $y' = 90^\circ$), (a) $\beta p/\pi = 2.5$ ($x' = 90^\circ$, $y' = 90^\circ$). The color map varies from 0° to 90° .

First three propagating modes in the structure of Figure 2-7 are represented in Figure 2-12. There is a frequency gap between the mode 1 and modes 2 and 3 where no mode propagates. This stop band starts at 63.64 GHz and ends at 116.7 GHz, close to the desired values 65 GHz and 115 GHz, respectively. The line of light is included in the dispersion diagram to compare the mode propagation of the gap waveguide cell with an empty unit cell. These start and end frequencies depend on the structure. Therefore, an analysis has been conducted for the five main parameters in Table 2-II, which are d , h , d/D , p/D and ϵ_r . The analysis consists of representing the frequency variations in the bandgap when a sweep of each of the parameters is simulated, keeping the other four parameters constant. This analysis is shown in sections from 2.3.1.1 to 2.3.1.5. A similar study is found at [66] that also compares the behaviour of the bandgap when using pins with different forms.

Figure 2-12: Dispersion diagram of the gap waveguide unit cell. The modes propagating in the structure are compared with the line of light.

2.3.1.1 Height of Pins d Sweep

The height of the pins is usually defined as a quarter wavelength at the starting frequency of the bandgap generating an AMC surface at the top of the pins. The end frequency is strongly affected by the sum of the height of the pins and the air gap between them and the upper metal plate $d + h$. This value is close to half the wavelength at the end frequency of the bandgap. To check this behaviour, a sweep in parameter d is carried out from $d = 0.7$ mm to $d = 5$ mm, keeping the rest of the parameters constant $h = 0.3$ mm, $d/D = 1.8$, $p/D = 2.4$ and $\epsilon_r = 1$. An example of how the structure changes with the sweep is shown in Figure 2-13. The start and end frequencies obtained in the simulations for each step in the sweep are shown in Figure 2-14 together with the relation between the values d and $d+h$ with a quarter and a half wavelength respectively.

Figure 2-13: Representation of the d sweep in the structure.

Figure 2-14: Bandgap depending on height of pins d .

The bandgap represented in Figure 2-14 shifts up or down in frequency when the parameter d becomes smaller or larger respectively. This behaviour is expected because the start frequency occurs when the quarter wavelength is approximately the value d . The value of the end frequency approximately corresponds to a half wavelength equal to the total thickness of the unit cell $d+h$. Those wavelengths have been calculated considering the metal pins embedded in a material with a dielectric constant $\epsilon_r=1$ (air). For higher values of the dielectric constant ϵ_r , we consider an effective relative dielectric constant equal to ϵ_r , which is valid approximation when the air gap is small ($h \ll d$). It can be observed that the value of d approximates to a quarter wavelength and the value $d+h$ approximates to a half wavelength, but they slightly differ for small values of d . Furthermore, the effects of the AMC behaviour start when the quarter wavelength is close to the

value d , then, it is expected that the bandgap starts at frequencies slightly below the approximated values.

2.3.1.2 Height of Air Gap h Sweep

The variation of the height of the air gap h should affect mainly to the end frequency of the bandgap. This is verified performing a sweep in parameter h from $h = 0.01$ mm to $h = 1.4$ mm, keeping the rest of the parameters constant $d = 0.9$ mm, $d/D = 1.8$, $p/D = 2.4$ and $\varepsilon_r = 1$. The effect of the variation of this parameter on the structure is shown in Figure 2-15. The start and end frequencies obtained in the simulations for each value in the sweep are shown in Figure 2-16 together with the relation between the values d and $d+h$ with a quarter and a half wavelength, respectively.

Figure 2-15: Representation of the h sweep in the structure.

Figure 2-16: Bandgap depending on air gap h .

As represented in Figure 2-16, the smaller the values of h , the wider the bandgap. This is because the second mode starts propagating like a TE_{01} mode in a parallel plate waveguide. Then, when the total thickness $d+h$ is smaller, the cut-off frequency increases. Thicknesses over wavelengths relations keep almost constant at 0.2 for the start frequency, and close to 0.5 for the end frequency. The start frequency falls to zero when h tends to zero because the AMC surface is replaced by a PEC surface. The result is a parallel plate waveguide with via holes uniformly distributed.

2.3.1.3 Height of Pins vs. diameter Relation d/D Sweep

The variation of the relation between the height of the pins and their diameter affect to the shape of the pins and to the periodicity between pins when considering a fixed value of d , since p/D is also constant. This will have an effect over the coupling between adjacent pins and the effect on the fields of the amount of metal forming the pins. This is verified performing a sweep in parameter d/D from $d/D = 0.1$ to $d/D = 5$, keeping the rest of the parameters constant $d = 0.9$ mm, $h = 0.3$ mm, $p/D = 2.4$ and $\epsilon_r = 1$. The structure changes with the variation of the parameter as shown in Figure 2-17.

Figure 2-17: Representation of the d/D sweep in the structure.

In Figure 2-18, the end frequency increases linearly until $d/D=1.5$, and for higher values it keeps approximately constant with the corresponding half wavelength almost equal to the value $d+h$. The discontinuity of the first order derivative that is appreciated in at $d/D=1.5$ transition is due to the propagation of different modes that start at different frequencies and their variation is independent. For values lower than $d/D=1.5$ the second propagating mode is an undesired mode whose cut-off frequency is lower than the cut-off frequency for the mode of interest, which is the dominant in the typical cases. About the start frequency, it falls to zero when d/D tends to zero. This is the expected behaviour because when d/D tends to zero with a constant value of d , then D and the periodicity p tend to infinity. In addition, when the d/D relation takes values above 3, the bandgap stabilises. This means that for thin pins a small variation of the diameter does not significantly affect the behaviour of the bandgap.

Figure 2-18: Bandgap depending on d/D relation.

2.3.1.4 Periodicity vs. Diameter Relation p/D Sweep

The variation of the relation between the periodic separation of pins and their diameter p/D , considering a constant value of D , affects to the coupling between adjacent cells and to the amount of field propagating in the space between pins. This is verified performing a sweep in parameter p/D from $p/D = 1.05$ to $p/D = 7.5$, keeping the rest of the parameters constant $d = 0.9$ mm, $d/D = 1.8$, $h = 0.3$ mm and $\epsilon_r = 1$. An example of the effect of this variation in the unit cell is shown in Figure 2-19.

Figure 2-19: Representation of the p/D sweep in the structure.

Figure 2-20: Bandgap depending on p/D relation.

As in the section 2.3.1.3, the cut-off frequencies of two different modes cross in Figure 2-20 at around $p/D=3$ for the end frequency of the bandgap. The half wavelength corresponding to the cut-off frequency of the mode of interest is close to d when the new mode starts degrading the available bandwidth for the bandgap with values of $p/D > 3$. For values of p/D close to 1, the available bandgap becomes narrower and finally for $p/D=1$ the bandgap disappears because the pins are touching each other and a quasi-TEM mode propagates in the air gap as a parallel plate waveguide.

2.3.1.5 Relative Dielectric Constant ϵ_r Sweep

The variation of the relative dielectric constant ϵ_r affects to the start and end frequencies of the stop band by shifting it to higher or lower frequencies if the value of ϵ_r is lower or higher, respectively. This is verified performing a sweep in parameter ϵ_r from $\epsilon_r = 1$ to $\epsilon_r = 10$, keeping the rest of the parameters constant $d = 0.9$ mm, $d/D = 1.8$, $p/D = 2.4$ and $h = 0.3$ mm. The effect of the variation of the relative dielectric constant is represented as a change in the colour of the dielectric material in Figure 2-21.

Figure 2-21: Representation of the ϵ_r sweep in the structure.

Figure 2-22: Bandgap depending on relative dielectric constant ϵ_r .

As expected, the bandgap shifts up and down in frequency when the relative dielectric constant decreases or increases, respectively. The values of the relations between thicknesses and wavelengths are closer to 0.5 and 0.25 in each case when ϵ_r increases due to the effective dielectric constant is closer to the relative dielectric constant of the substrate, since the fields tend to concentrate more in the dielectric substrate.

2.3.2 Ridge Waveguide and Ridge Gap Waveguide Analysis

Designing an RWG structure in a desired bandwidth presents two fundamental steps: the design of the metal pin unit cell for gap waveguide, whose behaviour in function of the different parameters has been analysed in section 2.3.1, and the design of the equivalent ridge waveguide. The structure and parameters for a ridge waveguide are shown in Figure 2-23. The figure represents a single-ridge waveguide and a double-ridge waveguide. The cut-off frequency analysis and dispersion behaviour for single and double-ridge waveguide are identical since double-ridge waveguide can be considered as a single-ridge waveguide with an electrical

symmetry at half the height of the waveguide. Then, both of the structures illustrated in Figure 2-23 are equivalent in terms of phase constant and dispersion behaviour. Regarding transmission losses and power handling, the double-ridge waveguide provides better results, since the field density is lower because the space available for the propagation is larger.

Figure 2-23: Single-ridge waveguide (a) and double-ridge waveguide (b). They are equivalent across the symmetry axis represented by a dashed line.

Ridge waveguides are structures used and deeply analysed in both theoretical and empirical ways in the 50s and 60s of last century. An analytical study of cut-off frequencies for the three first modes in single and double-ridge waveguides is found in [15] and a broader study of the TE_{10} mode in [67]. In the design of RGW, the main interest is to calculate the cut-off frequency of the TE_{10} mode and be sure that the second mode does not start the propagation inside the band of interest. Authors of both papers summarize the results in different curves depending on the cut-off wavelength of the TE_{10} mode in the ridge waveguide. In these studies, the waveguides are filled with air, although some of the RGW models that are considered in this section are inhomogeneously filled with dielectric. As the main part of the energy is concentrated over the ridge in the air gap, it is expected that the results will not differ very much from the air-filled case. A more recent study [68] proposes an analytical expression (2-2) of the cut-off frequency of the main mode for certain cases of interest, where c is the speed of light in vacuum.

In this section, we take as a reference for the following designs the double-ridge waveguide proposed in [69]. Figure 2-24 represents two versions of the double-ridge waveguide, Figure 2-24(a) with right-angled corners and Figure 2-24(b) with rounded corners. In the article [69], the authors choose the rounded version because it provides slightly more band and has lower average transmission losses in the band from 45 GHz to 110 GHz. The parameters selected for the design are shown in Table 2-III.

$$f_{c10} = \frac{c}{2(a-w) \sqrt{1 + \left(2.45 + 0.2 \frac{w}{a}\right) \frac{wb}{g(a-w)} + \frac{4b}{\pi(a-w)} \left(1 + 0.2 \sqrt{\frac{b}{a-w}}\right) \times \ln \csc\left(\frac{\pi g}{2b}\right)}} \quad (2-2)$$

Figure 2-24: Double-ridge waveguide with right-angle corners (a) and rounded corners (b).

TABLE 2-III: DESIGN PARAMETERS FOR RIDGE WAVEGUIDE

Parameters	Value (mm)
a	2.6
b	1.3
w	0.91
g	0.47
R	0.127

Figure 2-25 depicts a comparison of the dispersion diagrams for the first two modes propagating in the single-ridge waveguide and in the double-ridge waveguide (Figure 2-25(a)). For the double-ridge waveguide, the curves for the right-angled model and for the rounded model are illustrated in (Figure 2-25(b)). The phase constant of the modes in the single-ridge waveguide and the double-ridge waveguide with right-angled corners overlap perfectly, as they preserve an equivalence by symmetry. Compared to the phase constant of the modes in the rounded double-ridge waveguide, the cut-off frequency of the first two modes is shifted slightly upwards in frequency from the right-angled corner version. However, the fundamental mode only shifts by 0.5 GHz, while the second mode moves up by 3 GHz. Therefore, the single mode band increases.

Figure 2-25: Phase constant of the first and second mode in the RW and DRW. (a) Comparison between RW and DRW and (b) comparison between right-angled DRW and rounded DRW.

2.3.2.1 Example of Design for Ridge Gap Waveguide

For the design of a ridge gap waveguide, a gap waveguide unit cell need to be designed to generate a bandgap in the working frequency band. In this design we seek to match the response of the phase constant of the RGW with the ridge waveguide analyzed in the section 2.3.2. The bandgap needs to cover the band from 45 GHz to 110 GHz. For this example, a structure for RGW has been designed for a cut-off frequency of 39 GHz, as for the single-ridge waveguide in Figure 2-25(a). It is important to consider the cut-off frequencies of the higher order modes, which could be in a range of frequencies within the desired operating band (45 GHz to 110 GHz).

Once the unit cell for the gap waveguide has been defined, we replace the lateral metal walls of the single-ridge waveguide with three rows of periodic pins on each side, conserving as much as possible the dimensions of the ridge and the waveguide. Following the recommendations in [16], the number of rows of pins at the laterals is set to three as an optimum value for

minimizing the leakage of the fields with a reduced number of rows. Further parameter adjustment is necessary to match the cut-off frequency of the fundamental mode to that of the reference single-ridge waveguide and to maintain the cut-off frequency of the first high order mode at least 5% above the end of the working band.

Although in this chapter we have discussed a gap waveguide unit cell with circular metal pins, most of the work in the literature has applied gap waveguide technology with square metal pins. The reason for this is simpler electromagnetic modeling and mainly the reduction of manufacturing cost by CNC machining. By using evenly distributed square pins, the resulting structures are machined faster than using circular pins. However, this advantage is lost when using additive manufacturing. In this case, it is more advantageous to manufacture circular pins since right-angled corners are manufactured less precisely. However, this chapter includes designs of all gap waveguides (except MRGW) with square metal pins equivalent to circular pins for fair comparison.

The final unit cells of the RGW for the band from 45 GHz to 110 GHz are illustrated in Figure 2-26 with circular pins and in Figure 2-27 with square pins. The parameter values of the two designs are shown in the Table 2-IV. These unit cells are periodic only in the direction of propagation of the waveguide, so to calculate the dispersion diagram it is only necessary to vary the phase difference from 0° to 180° between boundaries in this direction. The resulting phase constant is shown in Figure 2-28. Figure 2-28(a) compares the modes of the RGW with circular pins (RGWc) with those obtained in the ideal single-ridge waveguide (RW). The cut-off frequency of the mode of interest is the same, although the value of the phase constant above 40 GHz differs slightly from the ideal RW. The mode limiting the single-mode band begins to propagate at 119.5 GHz, above the calculated high-order mode at RW. Figure 2-28(b) compares the results obtained for the circular pin model (RGWc) and the square pin model (RGWs). The phase constants of the TE_{10} modes of interest are practically the same in both cases, although the undesired modes drop slightly in frequency for the RGWs design. However, these unwanted modes are outside the working band with enough margin.

Figure 2-26: Structure and parameters of design for a wide-band ridge gap waveguide with circular pins.

Figure 2-27: Structure and parameters of design for a wide-band ridge gap waveguide with square pins.

TABLE 2-IV: DESIGN PARAMETERS FOR WIDE-BAND RIDGE GAP WAVEGUIDE

Parameters	Value for circular pins (mm)	Value for square pins (mm)
W	2.74	2.66
W_r	0.69	0.69
d	0.98	0.98
h	0.05	0.05
H_r	0.795	0.795
D	0.74	0.62
p	1.27	1.27

Figure 2-28: Phase constant of several modes in the RW and RGW. (a) Comparison between RW and RGWc and (b) comparison between RGWC and RGWs.

In addition to increasing the single-mode band by a factor of about two relative to a rectangular waveguide, ridge waveguides can also be used to reduce the volume occupied by the waveguide to propagate the TE_{10} mode. Two additional RGW designs with circular and square pins have been implemented to reduce the overall width of the waveguide by propagating a TE_{10} mode whose phase constant is equivalent to that of the TE_{10} mode of a WR-10 rectangular waveguide. This new design also tries to minimise losses, so that they are lower than for a design that simply consists of scaling the structure of Figure 2-26 and Figure 2-27 to match the cut-off frequency of the TE_{10} mode with that of the WR-10. As will be seen in the section 2.4.1, the losses are reduced by approximately 40%. The unit cells of these designs are represented in Figure 2-29 and Figure 2-30 with circular and square pins, respectively. The design parameters are shown in Table 2-V. The phase constants of the modes of these two designs are depicted in Figure 2-31, also compared to the TE_{10} mode and TE_{20} mode of the WR-10. Figure 2-31(a) shows the modes for the RGW with circular pins (RGWc) next to those of the WR-10. The TE_{10} mode for RGWc has a cutoff frequency very close to that of the TE_{10} mode for WR-10. Above 65 GHz the phase constants of the TE_{10} mode in both structures start to differ substantially. In Figure 2-31(b), the results obtained with the circular pin design (RGWc) and the square pin design (RGWs) are overlapped. The TE_{10} mode is practically identical in both cases. The undesired modes are outside the 75 GHz to 110 GHz working band, although in the RGWs case they occur at slightly lower frequencies.

Figure 2-29: Structure and designs parameters for a narrow-band ridge gap waveguide with circular pins.

Figure 2-30: Structure and designs parameters for a narrow-band ridge gap waveguide with square pins.

TABLE 2-V: DESIGN PARAMETERS FOR NARROW-BAND RIDGE GAP WAVEGUIDE

Parameters	Value for circular pins (mm)	Value for square pins (mm)
W	1.88	1.725
W_r	0.38	0.315
d	1	1
h	0.2	0.2
H_r	0.72	0.76
D	0.5	0.41
p	1.3	1.3

Figure 2-31: Phase constant of several modes in the WR-10 and RGW. (a) Comparison between WR-10 and RGWc and (b) comparison between RGWc and RGWs.

2.3.2.2 Example of Design for Microstrip Ridge Gap Waveguide

The microstrip ridge gap waveguide design process is similar to that of RGW, except that the metal pins are embedded in a dielectric substrate. Therefore, the relative dielectric constant of the material, as well as its thickness, must be considered when designing the unit cell. Typically,

the metal pins are implemented as via holes in the substrate and the ridge is implemented as a microstrip line connected to ground using via holes. In the example shown in Figure 2-32, a substrate with dielectric constant $\epsilon_r = 3$ and thickness $d = 0.585$ mm is used. The design has been made to have a single mode band between 75 GHz and 110 GHz, maintaining the corresponding security margins with the undesired modes. The design parameters used are shown in Table 2-VI. The dispersion diagram of the designed MRGW is depicted in Figure 2-33. The single mode band occupies a range from 65 GHz to 115 GHz. It is difficult to define the cutoff frequency of the mode of interest from Figure 2-33, as several crossings occur between different modes that are coupled together and their phase constant is distorted. It has been found that the mode whose cutoff frequency is around 53 GHz belongs to the mode of interest, so we can estimate that this is its cutoff frequency.

Figure 2-32: Structure and design parameters for a microstrip-ridge gap waveguide.

TABLE 2-VI: DESIGN PARAMETERS FOR MICROSTRIP-RIDGE GAP WAVEGUIDE

Parameters	Value (mm)
W	1.595
W_r	0.425
d	0.585
h	0.1214
h_m	0.02 (cte.)
p	0.78
D	$p/3 = 0.26$
D_c	$W_r - 0.15/\sqrt{\epsilon_r} = 0.338$
ϵ_r	3

Figure 2-33: Phase constant of several modes MRGW.

A representation of the field lines in a RGW structure is presented in Figure 2-34 and Figure 2-35. Most of the energy of the fields is concentrated in the air gap over the ridge, where the direction of the fields is almost perpendicular to the direction of propagation. Therefore, the cut-off frequency goes down in frequency when the ridge is introduced and the resulting mode TE_{10} is closer to a TEM mode. The intensity of the fields drops rapidly to below -30 dB relative to the maximum after the second pin on the side.

Figure 2-34: E-field lines for the mode of interest in RGW.

Figure 2-35: H-field lines for the mode of interest in RGW.

2.3.3 Groove Gap Waveguide Analysis

2.3.3.1 Design of Horizontal Groove Gap Waveguide (GGWH)

Designing a groove gap waveguide is simpler than designing a ridge gap waveguide, since the equivalent waveguide only depends on two parameters, width a and height b of the waveguide and typically a height corresponding to half the width $b=a/2$ is chosen, as shown in Figure 2-36. In this sense, it is only necessary to calculate a periodic unit cell with a metal pin whose bandgap fits the working band. After that, it is sufficient to define the width W of the groove gap waveguide. However, extra optimization is often needed to adjust the cut-off frequency of the TE_{10} mode and ensure that no unwanted modes propagate in the working band. In this case, we design a GGWH whose fundamental mode is equivalent to a WR-10 waveguide. This optimization is done using circular and square metal pins. The final designs are illustrated in Figure 2-37 (circular pins) and in Figure 2-38 (square pins). The values of the parameters used for the designs are shown in Table 2-VII.

Figure 2-36: Structure and parameters of design for a rectangular waveguide.

Figure 2-37: Structure and designs parameters for a horizontal groove gap waveguide with circular pins.

Figure 2-38: Structure and designs parameters for a horizontal groove gap waveguide with square pins..

TABLE 2-VII: DESIGN PARAMETERS FOR HORIZONTAL GROOVE GAP WAVEGUIDE

Parameters	Value for circular pins (mm)	Value for square pins (mm)
W	3.1	3.106
d	1	1
h	0.05	0.05
D	0.7	0.6
p	1.15	1.15

The dispersion diagrams of the two designed versions of GGWH are shown in Figure 2-39 together with the TE_{10} and TE_{20} modes of the WR-10 equivalent waveguide. The TE_{10} modes obtained in both the GGWH design with circular pins (GGWHc) and the design with square pins (GGWHs) present a phase constant practically identical and very similar to that of the TE_{10} mode in the WR-10 waveguide. The undesired modes in both GGWH designs are outside the working band, at frequencies below 40 GHz and from 115 GHz.

Figure 2-39: Phase constant of several modes in the WR-10 and GGWH. (a) Comparison between WR-10 and GGWHc and (b) comparison between GGWHc and GGWHs.

2.3.3.2 Design of Vertical Groove Gap Waveguide (GGWV)

For the design of a vertical groove gap waveguide, one more design parameter is added, which is the depth of the groove in the metal material. Due to the orientation of the waveguide, a design in which the metal side walls are completely replaced by the periodic metal pins is not possible. In this case, approximately half of this wall is replaced by the pins and the other half remains as a solid metal wall. The final designs, which are equivalent to the WR-10 waveguide, are represented in Figure 2-40 with circular pins and in Figure 2-41 with square pins. The parameters defining the periodic pins are different from those used in the GGWH design, as they have been optimized to ensure that all undesired modes are out of the working band and to reduce transmission losses. The values of the parameters used are shown in Table 2-VIII. The dispersion diagrams calculated for both designs are depicted in Figure 2-42. Figure 2-42(a) shows the modes that are propagated in the circular clamp design (GGWVc) together with the TE_{10} mode and TE_{20} mode of the WR-10 waveguide. The cut-off frequency of the GGWVc design is very close to that of the TE_{10} mode in WR-10, although from 70 GHz the phase constant is slightly separated. Figure 2-42(b) compares the phase constant of the modes in both circular pinned (GGWVc) and square pinned (GGWVs) designs. Both TE_{10} modes are very close to each other and the undesired modes are located outside the working band.

A representation of the field lines in the GGWH structure is presented in Figure 2-43 and Figure 2-44. The field lines are similar to a TE_{10} mode in a classical rectangular waveguide. In this case, the intensity of the fields across the rows of pins falls more slowly than in the RGW structure because the power is less concentrated than in the RGW with the presence of the ridge. So, the energy leakage through the side pins will be higher in GGW structures than in RGW structures with the same pin configuration.

Figure 2-40: Structure and designs parameters for a vertical groove gap waveguide with circular pins.

Figure 2-41: Structure and designs parameters for a vertical groove gap waveguide with square pins.

TABLE 2-VIII: DESIGN PARAMETERS FOR VERTICAL GROOVE GAP WAVEGUIDE

Parameters	Value for circular pins (mm)	Value for square pins (mm)
W	1.2	1.25
d	0.9	0.9
h	0.065	0.065
H_g	2.65	2.627
D	0.4	0.33
p	1	1

Figure 2-42: Phase constant of several modes in the WR-10 and GGWV. (a) Comparison between WR-10 and GGWVc and (b) comparison between GGWVc and GGWVs.

Figure 2-43: E-field lines for the mode of interest in GGW.

Figure 2-44: H-field lines for the mode of interest in GGW.

2.4 Transmission Losses Against Frequency

The main objective of this section is to compare the transmission losses between GGW, RGW, MRGW, SIW, rectangular waveguide and microstrip technologies at high frequencies. Different models of each technology have been designed and simulated to obtain their losses per meter as a function of the frequency. The surface losses produced by the metallic materials are calculated considering that all the metals in all the analyzed structures are aluminum with a conductivity $\sigma = 3.56 \cdot 10^7$ S/m. For designs requiring the use of a dielectric substrate, such as SIW, MRGW or microstrip, a material with relative dielectric constant $\epsilon_r = 3$ and loss tangent $\tan\delta = 0.002$ is used.

First, we simulate a WR-10 waveguide, which we use as a comparative element with the other structures in the W-band (75 GHz to 110 GHz). All the waveguides or transmission lines simulated in this section and in the section have been performed using a length greater than 10 wavelengths at the corresponding central frequency of the band. For W-band designs without a dielectric substrate, a length of approximately 50 mm has been chosen. The dimensions of the structures as well as their length have been appropriately scaled as a function of the central frequency of the band and as a function of the effective dielectric constant of the medium of transmission. From the obtained S-parameters and knowing the total length of the design L (mm), the losses per meter α in each structure have been calculated with expression (2-3).

$$\alpha = \frac{-1000}{L} \cdot 10 \log_{10} \left(\frac{|S_{21}|^2}{1 - |S_{11}|^2} \right) \text{ [dB/m]} \quad (2-3)$$

In Figure 2-45, the losses obtained for the WR-10 waveguide are depicted, together with the losses of an air-filled SIW. The air-filled SIW has been designed by scaling the design parameters indicated in Table 2-I of section 2.2 by a factor $\sqrt{3}$, since an epsilon $\epsilon_r = 3$ was used in the initial design. Losses obtained vary from 4.35 dB/m at 75 GHz to 3.15 dB/m at 110 GHz for the WR-10. The air-filled SIW has slightly higher losses, between 5.4 dB/m and 3.6 dB/m. This increase in losses is caused by a higher current density that must flow through the side metal posts compared to a rectangular waveguide with solid metal walls. As the surface available for current to flow in metal posts is smaller than in a full metal wall, the losses increase. Some of these additional losses may be due to radiation produced through the metal vias, but this effect must be negligible because the dimensions of the SIW parameters have been chosen to meet this condition.

Figure 2-45: Losses per meter in W-band for WR-10 and air-filled SIW.

Another classic waveguide to compare to some of the gap waveguide versions is the ridge waveguide. In section 2.3.2, two different designs were presented: single-ridge and double-ridge waveguide, called RW and DRW, respectively. As seen in Figure 2-25(a), the phase constant of the fundamental mode is equivalent in both models. In Figure 2-46, the results of transmission losses obtained by both right-angled corner designs are compared (Figure 2-46(a)) and a comparison between the DRW design with right angled corners and rounded corners (Figure 2-46(b)). The losses in the single-RW are between 3 dB and 4.5 dB higher than in DRW, as expected. Meanwhile, DRW has losses between 5 dB/m and 6 dB/m in the 45 GHz to 110 GHz band. The rounded corner DRW design has very similar losses to the right-angled corner design. The rounded design has slightly higher losses at the beginning of the band up to 60 GHz, but they are below from 60 GHz to the end of the band. On average, losses are slightly lower for the rounded-corner design.

Figure 2-46: Losses per meter from 45 GHz to 110 GHz for RW, DRW and rounded DRW.

The losses in the RGW designed in section 2.3.2.1, both wideband versions (45 to 110 GHz) and W-band versions (75 to 100 GHz) are compared in Figure 2-47. In Figure 2-47(a) the curves of losses per meter of the RGW with circular pins (RGWc) and with square pins (RGWs) are overlapped next to the single-ridge waveguide (RW). The three curves are very similar and take values between 8.2 dB/m and 10.5 dB/m, although it is observed that RGWc has lower losses than RGWs.

Figure 2-47: Losses per meter comparison between RW and wide and narrow-band RGW.

The W-band RGW designs are compared in Figure 2-47(b) along with the wideband RW. The former have lower losses in most of the W-band. The RGW model with circular pins (RGWc) has losses taking values from 7.3 dB/m to 9.5 dB/m. The design with square pins (RGWs) has more losses than the model with circular pins, with losses between 8.25 dB/m and 11 dB/m.

The last simulated structures without dielectric substrate are the GGW both in horizontal (H-plane) and vertical (E-plane) versions. The design data of these structures are detailed in section 2.3.3. The transmission losses of the GGW are depicted in Figure 2-48 together with the losses in a WR-10. The horizontal versions with circular (GGWHc) and square (GGWHs) pins are shown in Figure 2-48(a) and the vertical versions (GGWVc and GGWVs) are shown in Figure 2-48(b). Again, the GGWHc has lower losses than the square pin version, although its losses vary between 4 dB/m and 6 dB/m. These losses are about 2 dB to 2.5 dB higher than in a classic WR-10. Vertical GGWs have lower losses than the horizontal version, in a range between 3.5 dB/m and 5 dB/m. This difference is because the vertical GGW only replaces the ideal metal wall of the WR-10 at one side, while GGWH replaces both sides.

Figure 2-48: Losses per meter in W-band for WR-10 and GGW in horizontal and vertical.

Although the ideal simulations have shown higher losses for the gap waveguide versions than for a WR-10 waveguide, it is necessary to consider that in real prototypes these ideal conditions will not be achieved in any case. When manufacturing complex prototypes with rectangular waveguide, the pieces are usually separated into several parts to machine the waveguide paths. In the assembly process, good electrical contact must be achieved in all areas of the prototype, which can be very complex. Using prototypes with gap waveguide minimizes this problem, which is of critical importance especially at Ka-band frequencies and above. Further analysis with these aspects in mind is detailed in chapter III.

Figure 2-49 shows the transmission losses of the prototypes with a dielectric substrate. These designs are SIW, MRGW and microstrip (MS). As indicated at the beginning of this section, the curves represented in the three cases correspond to designs with a dielectric substrate with dielectric constant $\epsilon_r = 3$ and loss tangent $\tan\delta = 0.002$. The model that presents more losses is SIW since the whole field is confined within the dielectric. In the microstrip, part of the field is propagated by the air in the upper area of the microstrip, so the losses are lower. Nevertheless, SIW is considered a better option for high frequencies because the microstrip can cause spurious radiation problems in discontinuities such as bends or power dividers. The MRGW design has lower losses than the other two because most of the field is propagated in the air gap between the microstrip-ridge and the upper metal plate. However, the effect of the dielectric substrate on losses

is very negative in this frequency band for the three technologies, with losses between 30 dB/m and 50 dB/m, while all-metal designs reached a maximum of about 10 dB/m and typical losses between 4 dB/m and 5 dB/m. In addition, the loss tangent of microwave materials tends to increase with frequency, so many commonly used materials are likely to have a loss tangent greater than 0.002 in band W.

Figure 2-49: Losses per meter in W-band for SIW, MS and MRGW with $\epsilon_r = 3$ and $\tan\delta = 0.002$.

2.4.1 Analysis of Losses from 10 GHz to 110 GHz

To evaluate how the transmission losses of all the prototypes behave in a larger frequency band, all the prototypes analyzed in the section have been scaled to cover smaller frequency bands down to 10 GHz. In Figure 2-50, the scaling factors used for each prototype are represented. The prototypes analyzed in the 75 GHz to 110 GHz band are scaled by taking the values in Figure 2-50(a), considering the central frequency of the bands represented. These values are calculated as $scale = 92.5 \text{ GHz}/f_{\text{central}}$. Similarly, the prototypes analysed in the 45 GHz to 110 GHz band are scaled according to the values in Figure 2-50(b) for the central frequencies of the corresponding bands as $scale = 77.5 \text{ GHz}/f_{\text{central}}$. These central frequencies are 12.5 GHz, 18.5 GHz, 27.5 GHz, 41.5 GHz, 62.5 GHz and 92.5 GHz in Figure 2-50(a) and 15.5 GHz, 23 GHz, 34.5 GHz, 52 GHz and 77.5 GHz in Figure 2-50(b).

Figure 2-50: Scale values and frequency bands. (a) Scale for W-band (75 to 110 GHz) and (b) scale for band from 45 GHz to 110 GHz

Figure 2-51 compares the average transmission losses of single-ridge waveguide (RW) and double-ridge waveguide designs with right-angled corners (DRW) and with rounded corners (DRW rounded). The relationship between the loss values is maintained for all frequency bands evaluated. The average loss varies with frequency in proportion to the value $\sqrt{f}/scale$. Since $scale$ is inversely proportional to the frequency, we conclude that the losses are proportional to $f^{3/2}$. This factor can be extracted from the expressions (2-4) and (2-5) for conductive loss α_{c10} in the TE₁₀ mode of a rectangular waveguide [4]. In the expressions, R_s is the sheet resistance of the conductor, a and b are the dimensions of the waveguide, η is the intrinsic impedance of the medium, k is the wavelength number, σ is the conductivity of the metal and μ_0 is the magnetic permeability of vacuum. Considering that the values of $q = b/a$ and $r = f_c/f$ keep constant because the only difference between prototypes is a scale factor, and that R_s is proportional to $\sqrt{f}$, we can obtain the expression (2-6). Therefore, losses are reduced by a factor of 11.18 from prototypes at the central frequency of 77.5 GHz to prototypes with a central frequency of 15.5 GHz.

Figure 2-51: Mean losses per meter with respect to frequency for RW, DRW, and rounded DRW.

$$\alpha_{c10} = \frac{R_s(2b\pi^2 + a^3k^2)}{a^3b\beta k\eta} = \frac{R_s \left(1 + \frac{2b}{a}(f_c/f)^2\right)}{\eta b\sqrt{1 - (f_c/f)^2}} \quad (2-4)$$

$$k = \frac{2\pi}{\lambda} \quad R_s = \sqrt{\frac{\pi f \mu_0}{\sigma}} = m\sqrt{f} \quad (2-5)$$

$$\alpha_{c10} = \frac{m\sqrt{f}(1 + 2qr^2)}{\eta b\sqrt{1 - r^2}} = \frac{t\sqrt{f}}{b} = \frac{t\sqrt{f}}{scale \cdot b_1} = \frac{t'\sqrt{f}}{scale} = t''f^{3/2} \quad t = \frac{m(1 + 2qr^2)}{\eta\sqrt{1 - r^2}} = cte \quad (2-6)$$

, where b_1 is the height of the waveguide when the scale factor is 1, i.e. b_1 is constant.

In Figure 2-52, the average losses for eight types of structures are represented, which are: rectangular waveguide (WR), double-ridge waveguide with rounded corners (DRW rounded), groove gap waveguide in horizontal with circular pins (GGWHc), narrow-band (RGWc) and wide-band (RGWc wideband) ridge gap waveguide with circular pins, microstrip-ridge gap waveguide (MRGW), microstrip (MS) and substrate integrated waveguide (SIW).

Figure 2-52: Mean losses per meter with respect to frequency for various metal and dielectric structures.

Microstrip (MS), SIW and MRGW show higher losses due to the presence of the dielectric substrate. In this case, the total losses α are calculated as the sum of the losses produced by two different causes: the conductive losses α_c and the dielectric losses α_d . The dielectric losses can be calculated by applying the expression (2-7). This formula is valid if the loss tangent $\tan\delta \ll 1$, which is true in the vast majority of cases. In addition, this expression is valid for both TE and TEM modes, in which it is sufficient to substitute $\beta = k$ or $f_c = 0$. From the expression, it is deduced that the losses produced by the dielectric are proportional to the frequency. In Figure 2-52, it is clearly observed that in the curve for SIW, the losses in the dielectric dominate since the behavior is almost linear with the frequency. The losses produced by the conductors in TEM modes are calculated differently than shown in (2-4). In the case of the microstrip, the expression (2-8) is used, from which it is deduced that the conductive losses are also proportional to $\sqrt{f}/scale \propto f^{3/2}$, as well as for TE₁₀ modes. Z_0 is the characteristic impedance of the microstrip line $Z_0 = 50 \Omega$ and W is the width of the strip. The curve associated to microstrip (MS) in Figure 2-52 is also quasi-linear, so it is concluded that dielectric losses predominate over conductive losses. For the MRGW curve, the behavior is no longer linear or quasi-linear, so conductive losses dominate. This was expected because most of the electromagnetic fields propagate through the air gap over the strip.

$$\alpha_d = \frac{k^2 \tan\delta}{2\beta} = \frac{\pi f \sqrt{\mu\epsilon} \cdot \tan\delta}{\sqrt{1 - (f_c/f)^2}} = uf \quad u = \frac{\pi \sqrt{\mu\epsilon} \cdot \tan\delta}{\sqrt{1 - (f_c/f)^2}} = cte \quad (2-7)$$

$$\alpha_{cMS} = \frac{R_s}{Z_0 W} = \frac{m \sqrt{f}}{Z_0 W_1} = \frac{v \sqrt{f}}{scale} = v' f^{3/2} \quad v = \frac{m}{Z_0 W_1} = cte \quad (2-8)$$

, where W_1 is the width of the microstrip when the scale factor is 1, i.e. W_1 is constant.

As a comparison, Figure 2-52 includes some technologies without dielectrics that present lower losses by a factor between 3 and 16. In Figure 2-53, these technologies are represented in more detail and also include the air-filled SIW, the vertical GGW with circular pins (GGWvc) and the single-ridge waveguide (RW). The frequency behavior of all of them is proportional to $f^{3/2}$. The WR, air-filled SIW designs and the GGWvc and GGWHc designs have very similar losses since the designs are broadly equivalent. On the other hand, there are the RGWc (narrow

band) and rounded DRW, with losses close to double those obtained with the designs equivalent to rectangular waveguide. Finally, there are the wideband RW and RGWc waveguides with losses of approximately three times larger than those obtained with rectangular waveguide.

Figure 2-53: Detail of the mean losses per meter with respect to frequency for all-metal models.

2.5 Transmission Losses Against Dielectric Substrate

A study of the transmission losses has been conducted for the mode of interest for SIW and MRGW technologies designed to work in W-band using different dielectric constants and loss tangents. The data analyzed in this section are calculated as the average losses in the W-band. In addition, the same study has been carried out for the dielectric-filled rectangular waveguide and the microstrip to compare the results obtained. The values of the relative dielectric constant used take values from 1 to 10 and from 0 to 0.01 for the loss tangent, considered constant at all frequencies. To simplify the design process for each dielectric constant used, a scale factor has been applied for both SIW and dielectric-filled rectangular waveguide designs. Since the fields propagating in these structures are completely confined within the dielectric substrate, it is sufficient to scale all design parameters as a function of $\sqrt{\epsilon_r}$. The scale factor for each value of the dielectric constant is represented in Figure 2-54.

Figure 2-54: Scale factor as a function of the dielectric constant.

For microstrip it is necessary to calculate the width of the metal strip for each dielectric constant to achieve a characteristic impedance $Z_0 = 50 \Omega$. The MRGW case is more laborious, since for each dielectric constant it is necessary to redesign the entire structure so that the periodic pins generate a bandgap in the desired band and the microstrip-ridge propagates the mode of interest in that band.

Firstly, we carry out a theoretical analysis of the losses produced by the conductors and the dielectric substrate. On this occasion, we are interested in observing the behavior of the losses as a function of the dielectric constant, the loss tangent and the scale factor. We apply again the equations (2-4) and (2-7) to find the relationship between the losses and the three parameters of interest, considering the other parameters constant. As a result, we have that the conductive losses in TE₁₀ modes are inversely proportional to the value $\sqrt{\epsilon_r}/scale$, as shown in the expressions (2-9) and (2-10). Since the scale factor is calculated as $1/\sqrt{\epsilon_r}$, we conclude that the conductive losses are proportional to ϵ_r . The dielectric losses are proportional to the product $\sqrt{\epsilon_r} \cdot \tan\delta$, extracted from the expression (2-11). This result is valid for the four designs analyzed in this section. The conductive losses for a microstrip TEM mode are calculated as indicated by the expression (2-12), which shows that these losses are inversely proportional to the scale factor applied to the width of the microstrip.

$$\alpha_{c10} = \frac{R_s \left(1 + \frac{2b}{a} (f_c/f)^2\right)}{\eta b \sqrt{1 - (f_c/f)^2}} = \frac{m'(1 + 2qr^2)}{b \sqrt{\mu/\epsilon} \sqrt{1 - r^2}} = \frac{g \sqrt{\epsilon_r}}{b} = \frac{g \sqrt{\epsilon_r}}{scale \cdot b_1} = \frac{g' \sqrt{\epsilon_r}}{scale} = g' \epsilon_r \quad (2-9)$$

$$g = \frac{m'(1 + 2qr^2)}{\sqrt{\mu/\epsilon_0} \sqrt{1 - r^2}} = cte \quad (2-10)$$

, where b_1 is the height of the waveguide when the scale factor is 1, i.e. b_1 is constant.

$$\alpha_d = \frac{k^2 \tan\delta}{2\beta} = \frac{\pi f \sqrt{\mu\epsilon} \cdot \tan\delta}{\sqrt{1 - (f_c/f)^2}} = l \sqrt{\epsilon_r} \cdot \tan\delta \quad l = \frac{\pi f \sqrt{\mu\epsilon_0}}{\sqrt{1 - (f_c/f)^2}} = cte \quad (2-11)$$

$$\alpha_{cMS} = \frac{R_s}{Z_0 W} = \frac{m'}{Z_0 W_1 W_\epsilon} = \frac{v'}{W_\epsilon} \quad v' = \frac{m'}{Z_0 W_1} = cte \quad (2-12)$$

, where W_1 is the width of the microstrip when $\epsilon_r = 1$, i.e. W_1 is constant. W_ϵ is the multiplier factor of W_1 depending on ϵ_r .

2.5.1 Losses in Dielectric-Filled Rectangular Waveguide

The simulations for dielectric-filled rectangular waveguide have been performed by scaling the height and width of the waveguide by the scale factor shown in Figure 2-54. The results for the average losses obtained in W-band are shown in Figure 2-55. As shown in the expressions (2-9) and (2-11), the conductive losses are proportional to ϵ_r and the dielectric losses to $\sqrt{\epsilon_r} \cdot \tan\delta$. This is observed in the linear trend with ϵ_r of the curve for $\tan\delta = 0$, associated only with losses in the conductors. As the loss tangent increases, the trend with $\sqrt{\epsilon_r}$ becomes more evident. The linear trend with the value of the loss tangent is also observed. The graph shows loss values close to 400 dB/m for $\epsilon_r = 10$ and $\tan\delta = 0.01$. Obviously, such high losses are not useful in any real proposal, except as an absorbent, but they serve as a reference value in the comparison with other structures. Rectangular waveguides are air-filled in most of the cases, but the dielectric-filled waveguide analysis is of interest for a comparison with the results obtained with SIW in section 2.5.2.

Figure 2-55: Mean transmission losses in dielectric-filled rectangular waveguide at W-band.

2.5.2 Losses in Substrate Integrated Waveguide

For the simulated SIW models, all design parameters have been scaled according to the dielectric constant. Due to the manufacturing limitations of the metallic vias, some of these designs are not feasible, as the diameter is too small. However, the results are useful to get an overview of the losses introduced by this type of waveguide for different dielectric substrates. The average losses in W-band are depicted in Figure 2-56. It should be noted that the results are very similar to those obtained for the dielectric-filled rectangular waveguide. In SIW, slightly higher losses are observed due to the increase in current density in the lateral metallic vias. The trend is linear with ϵ_r for $\tan\delta = 0$, but quickly dominates the trend with $\sqrt{\epsilon_r}$ for higher loss tangents.

Figure 2-56: Mean transmission losses in SIW at W-band.

2.5.3 Losses in Microstrip

The design of the microstrip for each dielectric constant value is addressed differently, as the scale factor used for SIW and rectangular waveguide cannot be applied. Instead, we keep the height of the dielectric substrate constant to $h = 0.4$ mm and calculate the value of the width of the strip W to obtain an impedance of 50Ω . The values of W obtained for each value of relative dielectric constant are shown in Figure 2-57(a). In Figure 2-57(b), we depict an example of the phase constant as a function of frequency for the TEM mode propagating through the microstrip with dielectric constant $\epsilon_r = 3$. Some of the higher modes and surface modes that are excited in the microstrip are also represented. These modes are responsible for spurious radiation, field leakage and unwanted coupling in microstrip transmission networks for high frequencies.

Figure 2-57: Width of the microstrip with the dielectric constant (a) and phase constant of the Q-TEM mode in the microstrip and several high order modes (b).

The average losses obtained in W-band in microstrip are shown in Figure 2-58. The losses for the $\tan\delta = 0$ curve have a quasi-linear trend since they are inversely proportional to the factor of variation of the width of the microstrip W_e . In turn, this factor W_e has a tendency close to $1/\epsilon_r$, so it justifies that the curve of losses is approximately proportional to ϵ_r . For the other values of loss tangent dominates the tendency with $\sqrt{\epsilon_r}$.

Figure 2-58: Mean transmission losses in microstrip at W-band.

2.5.4 Losses in Microstrip-Ridge Gap Waveguide

For the study of MRGW, the parameters length d and periodicity p of the periodic pins and the height h of the air gap have been optimized to maintain the bandgap over the entire W-band for each dielectric constant used. At the same time, the values of width W_r of the microstrip-ridge and the separation W between pins closest to the strip have been chosen so that only the mode of interest is propagated in the band. The rest of the parameters maintain the original proportion. The optimised design values are shown in Figure 2-59. The average loss values obtained in the simulations are depicted in Figure 2-60.

Figure 2-59: Variation of the design parameters for MRGW depending on the dielectric constant.

Figure 2-60: Mean transmission losses in MRGW at W-band.

An approximately linear trend is observed with the relative dielectric constant, associated to the losses in the conductors. This is consistent with the propagation of the fields above the strip through the air gap. Despite this, there is a certain propagation through the dielectric substrate that supposes an increase of the losses in a factor of approximately 2.5 between the curve $\tan \delta = 0$ and $\tan \delta = 0.01$. Compared with the increase in a factor close to 10 produced in SIW, the effect

of the dielectric in MRGW is relatively small. On the other hand, the increase in conductive losses as the dielectric constant increases is very pronounced. The increase from $\epsilon_r = 1$ to $\epsilon_r = 10$ is by a factor of approximately 10, which is explained by the reduction of the metallic surface available for propagation and by the reduction of the air gap. All this increases notably the current density through the strip, with the consequent increase of losses.

2.5.5 Combined Conductive and Dielectric Loss Comparison

For a better comparison between technologies, the maximum ($\tan\delta = 0.01$) and minimum ($\tan\delta = 0$) losses in dielectric-filled rectangular waveguide (WR), SIW, microstrip (MS) and MRGW are presented in Figure 2-61 for every simulated value of relative dielectric constant. Figure 2-61(a) shows the total losses in these four technologies considering a dielectric substrate with a high loss tangent $\tan\delta = 0.01$. The technology with the highest losses is SIW, closely followed by the dielectric-filled rectangular waveguide. The third technology is the microstrip with losses between 25% and 35% lower because the effective dielectric constant is decreased by the propagation of part of the fields through the air located over the microstrip. Finally, MRGW is the technology based on dielectric substrate with the lowest losses, with a reduction between 35% and 80% compared to designs in SIW. This reduction is maintained between 40% and 50% even for low loss tangents, around 0.002, using substrates with a dielectric constant between 1 and 4. On the contrary, considering only the losses in the conductors (or using a dielectric substrate with a loss tangent much lower than 0.002), MRGW is the technology with the highest losses. This can be clearly seen in Figure 2-61(b), where the conductive losses ($\tan\delta = 0$) in WR, SIW and microstrip are similar, but the losses in MRGW are between 2 and 3 times higher.

Figure 2-61: Mean transmission losses comparison at W-band for dielectric-based structures. (a) Total losses per meter with 0.01 loss tangent and (b) conductive losses per meter (loss tangent equals zero).

2.6 Conclusion

In this chapter, an extensive analysis has been conducted about the design methods of the most commonly used waveguide and transmission line technologies in microwave and millimeter-wave devices. Waveguide technologies on metal have been studied, such as rectangular waveguide (WR), single (RW) and double-ridge waveguide (DRW), groove gap waveguide (GGW) and ridge gap waveguide (RGW). The metal used in all the designs is aluminium ($\sigma = 3.56 \cdot 10^7$ S/m). Other technologies implemented on dielectric substrate have also

been analysed, such as substrate integrated waveguide (SIW), dielectric-filled rectangular waveguide, microstrip and microstrip-ridge gap waveguide (MRGW). The designs made have been simulated for different frequency bands, from 10 GHz to 100 GHz. In addition, the technologies based on dielectric substrate have been analysed as a function of the dielectric constant of the medium used and its loss tangent.

Among the all-metal technologies, the low losses of GGW, very close to those of WR, are remarkable. However, the use of GGW instead of WR is preferable in many cases for devices in Ka-band and above, as it does not require good electrical contact between the parts that conform the power distribution network. WR have lower losses only in the ideal case. In a practical case, field leakage and unwanted resonant effects may occur due to bad electrical contact somewhere in the device. In chapters III, IV and V, the GGW technology is applied in multiple 3D printed prototypes with excellent results. Regarding designs based on dielectric substrate, dielectric-filled waveguide technology has a high implementation cost compared to SIW, providing similar results, but it is of interest as a reference for a complete comparison. Microstrip has been included as one of the most widely used technologies in microwave devices for its simplicity of design, ease of manufacture and low cost. However, its use in millimetre-wave frequency bands is not recommended due to its high dielectric losses and undesired effects such as spurious radiation, appearance of surface modes, undesired couplings, etc. These undesired effects are eliminated using SIW technology, although high dielectric losses are maintained. This last problem is solved in the MRGW implementation, in which most of the fields are propagated through the air gap over the microstrip-ridge. In this case, as the field is confined by the presence of periodic pins, there are no problems of unwanted coupling and spurious radiation. However, for dielectric substrates with very low losses, similar loss results are obtained for SIW and MRGW. In these cases, the reduction of dielectric losses in SIW compensates for the relatively high conductive losses in MRGW. In any case, in SIW, losses between 20 dB/m and 50 dB/m could be achieved using a substrate with a loss tangent of 0.001, which is difficult to find in the W-band and has very high prices. In that case, GGW technology with losses around 4 dB/m in W-band is the best option for most of the design cases developed in this doctoral thesis.

Chapter III

Additive Manufacturing Applied to Gap Waveguide at Ka Band

3.1 Introduction

One of the most critical aspects in planar antennas at frequencies from Ka band and above is the losses in the distribution network. This is because the greater the losses the bigger the number of radiating elements needed and then, the size and cost of the antenna. In order to achieve very low losses while keeping a good manufacturing performance in cost, weight and size, gap waveguide technology has been considered, which uses a metamaterial built with a periodic bed of nails that prevents the transmission of power in the transverse direction and confines the wave along the desired path. The main objective of this chapter is to propose a manufacturing technique for low losses distribution networks with a fraction of the weight and cost of conventional metallic solutions like CNC machining. Many designs of microwave and mm-wave devices and antennas based on gap waveguide can be found in the literature, such as [70]-[73]. However, gap waveguide also has some disadvantages. The need for several rows of periodic pins on each side of the waveguide significantly increases the size of this type of structure, which makes it very complicated to design arrays of radiating elements that are close enough to avoid grating lobes. Another disadvantage is the high cost of manufacturing this complex lattice of metal pins using classic CNC milling technologies in aluminium. Gap waveguide is therefore an ideal candidate for testing the potential of 3D printing and for reducing the cost of prototyping.

The designed and manufactured prototypes in this chapter are based on GGW technology. For the interests of this work, additive manufacturing is oriented to the design of full metal gap waveguide structures. Gap waveguide is the suitable technology for additive manufacturing, because it is not needed an electrical connection between upper and lower metallic plates and the manufacturing of the two parts can be made separately. The critical piece is the one that includes the periodic lateral pins, which presents a complicated structure that can be printed and metallized very easily. The design of groove gap waveguide models has been implemented in 3D-printed technology plated with a 5-10 μm copper layer. The printing technology used is called stereolithography (SLA) and the accuracy of the final printed piece is specified as 30 μm in X-Y plane and 100 μm in Z direction due to the stacking of plastic layers.

The main objective of this chapter is to analyse the constraints that metallized 3D-printing technology imposes over GGW designs for distribution networks in planar structures and compare the performance with common milling techniques in aluminium. The organization of the chapter is as follows: section 3.2 shows the designs and simulations of transitions to WR-28, bends and power dividers in the selected technology. Section 3.3 presents the properties of the metallized 3D-printing technology that has been used, together with a comparison with common milling aluminium techniques. A parametric study of the effect of different physical properties such as roughness or manufacturing imperfections is conducted in section 3.4. Section 3.5 exposes a comparison of simulations and measurements of the final manufactured prototypes and a study of

losses per unit length in metallized 3D-printed GGW. A study of the effect of different materials, resolutions and metallization processes is detailed in section 3.6. In section 3.7, a comparison between 3D printed rectangular waveguide sections and commercial sections is presented. Finally, conclusions are drawn in section 3.8.

3.2 GGW Design for a Band from 28 GHz to 30 GHz

3.2.1 Base Groove Gap Waveguide Structure

The base structure includes a bed of nails made of three rows of pins for the lateral lattice in a copper plated surface that is covered by an aluminium plate. A 3D view of this structure is shown in Figure 3-1(a), a cross-section in Figure 3-1(b) and the design parameters are provided in Table 3-I.

Figure 3-1: Basic GGW structure used for the prototypes designed in this work. View of the design parameters (a): $h = 0.6$ mm, $D = 1$ mm, $W = 8$ mm, $p = 2.6$ mm, $d = 3$ mm. View of the paths where the propagation is allowed or forbidden (b).

TABLE 3-I: PARAMETERS OF GGW BASE STRUCTURE

Parameter	Length (mm)	Parameter	Length (mm)
h	0.6	p	2.6
D	1	d	3
W	8	Materials	Aluminium/Copper

Figure 3-2 represents the dispersion diagram of the structure. This diagram has been calculated using the Eigenmode Solver in CST considering a unit cell with periodic boundaries in the direction of propagation. The desired band for our application includes frequencies from 28 GHz to 30 GHz. The diagram depicts the behaviour of the propagation constant β of the excited modes in both the GGW and in the equivalent rectangular waveguide. It can be observed how the periodic pins block the propagation between 20 and 40 GHz approximately and there is a single mode band between 23 and 40 GHz in which the TE_{10} mode propagates as well as in the rectangular waveguide model.

Figure 3-2: Dispersion diagram of the GGW unit cell (solid lines) and the equivalent rectangular waveguide (dashed lines). The total height of both structures is 3.6 mm, while the width of the equivalent rectangular waveguide for the same cut-off frequency of the TE_{10} mode is 6.51 mm.

3.2.2 Design of a 90-Degree Bend in GGW

The design in GGW is very limited by the diameter of the pins (D) and the separation between them (p), which cannot be very small (at least 1 mm) to ensure a correct metallization of the plastic-printed model. Because of that, the only element that has been optimized is the distance of one single pin from the corner. This distance is measured between the centre of the pins in the laterals and the centre of the optimized pin. Figure 3-3 shows the structure and parameters for the GGW bend implemented in metallized 3D-printed technology.

Figure 3-3: Bend structure for GGW.

The results for the bend in GGW are presented in Figure 3-4. The design matches very well in the desired frequency band from 28 to 30 GHz, with a value for the return losses higher than 40 dB with a bandwidth at 20 dB wider than 8 GHz. The losses of the GGW bend are lower than 0.03 dB in the desired band.

Figure 3-4: Simulated S-parameters for the optimized GGW bend.

3.2.3 Design of a Two-Way Power Divider in GGW

We have modelled a two-way power divider design for the power splitting needed in the distribution network of the antenna. The design is based on the proposed method found at [74] for an H-plane power divider in rectangular waveguides. The design consists of a septum divider with a determined length from the end wall of the divider and an iris at the input of the divider that narrows the width of the GGW. These two structures have been implemented with pins. In [74], it is determined the optimum length and width for both septum and iris, but in the GGW design width is limited by the radius of the pins.

As can be seen in Figure 3-5, the septum has been implemented only with a pin at a certain distance from the pin in the wall. The iris consist of an elliptic pin with a semi-major axis that slightly penetrates in the guide. However, this is not enough to obtain a good matching. Then, the pins of the end wall of the divider have been shifted from their original position. The four internal columns of pins have been moved outward.

Figure 3-5: Power divider structure for GGW.

The results for the two-way power divider in GGW are presented in Figure 3-6. The obtained return losses in the desired frequency band from 28 to 30 GHz are almost better than 25 dB, with a value higher than 20 dB in a 4 GHz bandwidth centered in 29 GHz. The losses of the two-way power divider are almost lower than 0.05 dB in the desired band.

Figure 3-6: Simulated S-parameters for the optimized GGW power divider.

3.2.4 Design of a Transition from WR-28 to GGW

The design of the WR-28 to GGW transition is based on a structure similar to that found in [16]; with the input slot in the bottom plate of the GGW model. The final transition is exposed in Figure 3-7. The optimization is based on four parameters, that are: a shift of the lateral pins around the input of the WR-28 measured from the centre of the pins; a displacement of the end pins near the input of the WR-28 measured between the external long side of the input and the centre of the pins; an iris that narrows the WR-28 slot in the bottom metal input; and an offset distance between the centre of the first set of pins in the guide and the internal long side of the slot. We have also considered the screws for joining the flange with the GGW model in the structure from bottom to top metal plates.

Figure 3-7: WR-28 to GGW transition structure.

Figure 3-8: Simulated S-parameters for the optimized WR-28 to GW transition.

The results for the transition are depicted in Figure 3-8. Return losses are better than 20 dB in a 3 GHz bandwidth centred in 29 GHz. Losses are approximately 0.03 dB in the desired band. The transition is the limiting factor for a good matching in comparison to the results for bend and power divider.

3.3 Metallized Plastic versus Machined Aluminium

We have manufactured two prototypes in GW technology: one using milling aluminium technique and the other one with metallized 3D-printing to conduct a realistic comparison between both manufacturing methods. The results of a single straight section both in metallized plastic and in machined aluminium are presented in Figure 3-9. A comparison between the properties of materials and technologies can be found in Table 3-II.

Figure 3-9: Prototype in machined aluminium (a) and metallized 3D (b).

TABLE 3-II: PROPERTIES OF COPPER PLATING, PLASTIC AND ALUMINIUM

Parameter	Copper plating	Plastic	Aluminium
Material	99.9% pure copper	Accura® SL 5530	AL-6082-T6 95% Al min.
Mass density	8.96 g/cm ³	1.25 g/cm ³	2.71 g/cm ³
Electric cond.	5.8 · 10 ⁷ S/m	~	2.55 · 10 ⁷ S/m
Thermal cond.	401 W/mK	0.173 W/mK	172 W/mK
Coeff. thermal expansion	16.6 · 10 ⁻⁶ /K	84 · 10 ⁻⁶ /K (T < 122 °C) 159 · 10 ⁻⁶ /K (T > 122 °C)	23.1 · 10 ⁻⁶ /K
Fusion/deflect. temperature	1,084 °C	210 °C (0.455 N/mm ²) 115 °C (1.82 N/mm ²)	610 °C

The more relevant advantages of metallized 3D-printed plastic in comparison to common machined aluminium technique for the prototypes in Figure 3-9 are presented next. The plated plastic model achieves a weight reduction of 48% of the weight in machined model. The total cost of the plated 3D-printed prototype is 16% of the cost for the machined model. This cost depends on the manufacturing time, which is 30 minutes for the 3D-printed model. However, our design is optimized for 3D-printing and machining time may be reduced considering squared pins homogeneously distributed. This would reduce degrees of freedom in the design process, but it supposes a reduction of 20% in machining time and cost.

Apart from these remarkable advantages, both techniques present manufacturing drawbacks. On the one hand, the flat and wide structure of the plastic piece has forced to build the part at angle to avoid curved surfaces due to material shrink. This transfers the accuracy in vertical direction ($100\text{ }\mu\text{m}$) of the stacked layers along all surfaces and print lines become visible, as it can be observed in Figure 3-10(b). On the other hand, with an accuracy of $10\text{ }\mu\text{m}$, the aluminium prototype does not present this problem. Nevertheless, there are limitations at the corners in WR-28 transitions due to the 0.5 mm radius of the milling cutter (Figure 3-10 (c)).

Figure 3-10: Scanning Electron Microscope images of a pin in aluminium (a) and in copper plated plastic (b). Optical microscope images of the corner of the transition in aluminium (c) and in copper plated plastic (d). In (b), the stacked layers can be observed in the top of the pin. A worse manufacturing result of the transition is observed in (c) than in (d) due to the diameter of the milling drill.

A study of the surface profile in both prototypes has been performed to observe the limiting imperfections. In Figure 3-11, Scanning Electron Microscope (SEM) images of the roughness are

presented in both cases. For a better estimation of the roughness, we include measurements with profilometer in Figure 3-12, whose results are coherent with the visual aspect shown in Figure 3-11.

Figure 3-11: SEM micrographs of the surface roughness in aluminium with a x1500 amplification factor (a) and x7000 in copper plated plastic (b).

Figure 3-12: Measurements and normal distribution fitting of the surface roughness obtained with the profilometer Veeco DekTak 150. Measured roughness (DIN EN ISO 4287) in the copper surface (a) is $R_q = 0.3 \mu\text{m}$. The value obtained for the aluminium surface corresponds to $R_q = 3.5 \mu\text{m}$. The practical resolution in vertical direction is better than 10 nm.

An additional measurement of the waviness in the plastic model reveals the stacking of layers and the printing angle in Figure 3-13.

Figure 3-13: Waviness of the copper surfaces due to the stacked layers of the plastic during the 3D printing process (a). Detailed view of a single layer in (b) with a thickness of $100\text{ }\mu\text{m}$ and an elevation angle of 15° . Apart from the tilt in elevation, the layers are also disposed with an angle of 45° in azimuth with respect to the direction of propagation (see Figure 3-15(a)).

We take as an example the prototype in Figure 3-9(b) to visually explain the concepts of waviness and roughness mentioned above. Roughness and waviness can be extracted from the profile measurement. There is a higher level called form, which in this case is a horizontal line and is not represented in the graphs. In the example shown in Figure 3-14, the roughness is multiplied by a factor of seven in order to clearly see its effect on the curves.

Figure 3-14: Extraction of the waviness and the roughness of a measured profile (a). The measured profile is shown as the sum of waviness and roughness (b).

In Figure 3-15, the detailed structure of transitions for both technologies is shown for the prototypes in Figure 3-9. The stacked layers in the printed prototype can be appreciated in Figure 3-15(a). In the case of the aluminium prototype in Figure 3-15(b), the rounded corners are visible, as well as a certain roughness of the material due to a sanding process applied after fabrication. We have simulated both models considering the roundness and the stacking of layers, together with the roughness measured in Figure 3-12. The simulated and measured results of the models presented in Figure 3-9 are compared in Figure 3-16.

Figure 3-15: Detailed structure of metallized (a) and machined (b) models. A caption of the simulated structure is placed at the top-right corner of (a) and at the top-left corner of (b).

Figure 3-16: Comparison of measurement (MEAS) and simulation (SIM) for the metallized 3D-printed plastic (3D) and the machined (MC) models. Amplitude of S-parameters (a) and S_{21} phase difference between 3D and MC (b). Prototypes have been measured with an Agilent 8722ES Network Analyzer after a TRL calibration with a HP R11644A WR-28 Calibration Kit.

As it can be seen in Figure 3-16(a), the main difference is in the insertion losses, being significantly higher in the machined model due to the roughness level. In general terms, simulation and measurement are in very good agreement, especially for the 3D-printed model. Simulated reflection coefficient slightly differs from the measurement in the machined model due to manufacturing deviations. The average losses between 28 to 30 GHz are shown in Table 3-III for the simulations and measurements presented in Figure 3-16(a), together with the losses considering ideal properties of the materials. Ideal simulations do not consider roughness or manufacturing imperfections. The roughness found in the machined device stems from a sanding process applied after the milling of the material. This additional process removes the burrs and undesired edges in the piece. Regarding the phase delay due to roughness, phase difference between 3D printed model and machined model is depicted in Figure 3-16(b) with a phase difference of 3 degrees at 29 GHz approximately. Additional comparisons between simulations and measurements of several prototypes in metallized 3D-printed plastic are presented in section 3.5.

TABLE 3-III: LOSSES IN METALLIZED 3D-PRINTED PLASTIC AND MACHINING TECHNOLOGIES

Model	Losses (dB)	Model	Losses (dB)
3D Ideal	0.056	MC Ideal	0.073
3D Simulation	0.085	MC Simulation	0.326
3D Measurement	0.088	MC Measurement	0.315

3.4 Parametric Study. GGW vs Rectangular Waveguide Performance

In this section, several parametric studies have been conducted for different values of electrical and physical properties of the materials.

3.4.1 Modelling the Impact of Roughness and Conductivity

Due to the fact that the surface roughness supposes a greater impact over the losses than conductivity, losses and phase delay have been simulated for different values of roughness R_q in Figure 3-17 and conductivity σ in Figure 3-18. Roughness is modelled using the gradient model in CST MWS [75], considering the value of roughness R_q as the standard deviation of the surface profile. This effect increases surface impedance and introduces additional attenuation α (Figure 3-17(a)) and a phase delay (Figure 3-17(b)). Losses increase approximately 1 dB/m for each additional 0.5 μm of roughness at 29 GHz. The increment in phase delay is not as linear, being lower when increasing roughness. The reduction of the conductivity of the metal itself also increases the surface impedance, but to a lesser extent as it is shown in Figure 3-18(a) and Figure 3-18(b). Losses and phase delay are inversely proportional to the square root of the conductivity. Comparing the results with the ones in Figure 3-17, the main contribution of losses and phase delay come from surface roughness in the most typical cases.

Figure 3-17: Attenuation (a) and phase delay increment (b) per unit length in the GGW structure in function of the roughness R_q of the copper surface.

Figure 3-18: Attenuation (a) and phase delay increment (b) per unit length in the GGW structure in function of the conductivity σ of the surface.

3.4.2 Modelling the Impact of Copper Thickness

As it can be seen in Figure 3-19(a), skin depths from Ka to W band are in the order of the measured roughness so it is expected that its effect on losses at 60 GHz or 90 GHz will not be remarkable. As copper thickness is several multiples of the skin depth, practically no field escapes from the metallic layer (see Figure 3-19(b)). A thickness of $2.67 \mu\text{m}$, or seven multiples of the skin depth, is considered enough to avoid any important effect in the performance due to radiation through the copper plating. Furthermore, as it will be seen in Figure 3-24, where even $100 \mu\text{m}$ of stacked layers do not suppose a great deviation in the S-parameters, the extra copper layer of several micrometres in all the surfaces during the fabrication process is negligible and it is not necessary to take into account in the design process.

Figure 3-19: Skin depth as a function of frequency (a) and amplitude decay (b) for the skin depth at 30 GHz considering an ideally smooth conductor surface.

3.4.3 Modelling the Impact of Bad Contacts in GGW and RW

Apart from conductor properties, the main advantage of gap waveguide technology over classical RW resides in the contactless surfaces. Manufacturing a planar structure in RW is very complex in both machined and plated-printed technologies in a single piece. For a fair comparison, both technologies have been compared divided in two parts, the main structure in the lower copper plate and a covering upper aluminium plate. We have simulated both structures considering a bad electrical contact represented by a certain separation of the upper metal plate from its ideal position. This separation inserts a space of air of several micrometres. In Figure 3-20, the effect in the losses for WR-28 and GGW is depicted for various separations. For WR-28, losses increase about 1 dB/m for each additional micrometre of separation, while in GGW are practically constant. Figure 3-21 shows the leakage of fields in the WR-28, which does not occur in GGW. This behaviour makes the GGW a better candidate than RW for integrated planar networks if losses are critical, despite the increase of volume due to the periodic pins.

Figure 3-20: Attenuation of WR-28 in (a) and of our GGW structure in (b) due to a bad contact with a separation (sep) between upper and lower plates, which is zero for the ideal case.

Figure 3-21: Field leakage of GGW in (a) and WR-28 in (b). Cross-section of the amplitude of the electric field at the bad contact for a separation of 1 μm. (a) is represented in Figure 3-22(b) with an extended scale for a better visualization.

3.4.4 Modelling the Impact of Field Leakage through Pins

Since insertion losses measured in Figure 3-16 are below 0.1 dB, the contribution of radiation through the periodic pins may be important. In Figure 3-21(a), radiation losses are not visible due to the limited scale. Therefore, we include a study of radiation losses in function of the number of pin rows (Figure 3-22(a)) and the level of simulated fields beyond the pins (Figure 3-22(b)). The results conclude that the optimum number of rows of pins is three, with a radiated field of about -80 dB in relation to the amplitude of the field at the centre of the GGW. This means that the impact of radiation losses in measurements is negligible.

Figure 3-22: Attenuation due to radiation through the lateral pins in function of the number N_{pins} of periodic rows (a). Cross-section of the amplitude of electric field (b) for the three rows case (our design). The difference between N_{pins} 3 or 4 is negligible, with an amplitude lower than -75 dB outside the structure.

3.4.5 Modelling the Impact of Stacked Layers of Plastic

Another aspect of the manufacturing process that may lead into important deviations in the final response of the prototypes is the thickness of the stacked layers of plastic. A parametric

simulation has been conducted considering different values of thicknesses δ_{layer} shown in Figure 3-23 for a 90 mm length model from input to output.

Figure 3-23: Structure and configuration of the stacked layers. Layers are disposed with a tilt of 15° and an angle of 45° in respect of the direction of propagation due to our experience (see Figure 3-15(a)). Simulated thicknesses δ_{layer} are 100 μm (a), 200 μm (b) and 300 μm (c).

The results for reflection coefficient are shown in Figure 3-24(a) and for transmission in Figure 3-24(b).

Figure 3-24: Deviation in S-parameters due to the thickness δ_{layer} of plastic layers. Reflection coefficient in (a) and transmission coefficient in (b).

An important shift to lower frequencies is appreciated for larger thicknesses, but the difference between the ideal structure ($\delta_{layer} \rightarrow 0 \mu\text{m}$) and our manufactured structure ($\delta_{layer} = 100 \mu\text{m}$) is not of great relevance. However, this may be a problem when extending this technology at higher frequencies. Then, a better printing resolution will be required, but this does not represent a problem at least until 100 GHz because the manufacturing capacity can achieve a stacking of 50 μm layers or less.

3.4.6 Modelling the Impact of Mechanical Deformation with Temperature

For a complete analysis, a study of the effects over the electromagnetic (EM) response in function of temperature has been conducted. Thermal and mechanical properties of the materials are introduced in the simulation, which uses a combination of the EM solver and the Multiphysics solver in CST [76]. First, a temperature simulation of the model considering thermal losses coming from EM calculation has been performed at a certain ambient temperature T . The result is used as an input for a mechanical simulation with a reference temperature. The generated deformation mesh is simulated again with the EM solver in order to obtain the deviation due to temperature deformations. Figure 3-25 represents the mechanical deformation in the bottom plate of the models in solid copper (a) and plated plastic (b) at 100 °C and Figure 3-26 depicts the simulated variations in S-parameters of such model for -50 °C, 25 °C and 100 °C. Results present a good behaviour in a wide range of temperatures with minimal changes in the response. Simulations have not been evaluated at higher temperatures because the deflection temperature of this plastic is about 115 °C.

Figure 3-25: Material displacement of a solid copper model in (a) and of our copper plated plastic in (b) at an ambient temperature of 100 °C (Reference temperature = 25 °C). Deformation in plastic is more pronounced because its coefficient of thermal expansion is higher. Transitions to WR-28 are fixed to the system and do not experiment any displacement.

Figure 3-26: Variation in S_{11} parameter (a) and S_{21} parameter (b) in function of temperature.

3.5 Final Prototypes and Results

In this section, we present the final models that have been manufactured compared with the simulated results. The models included in the prototype are a straight section, a U-shaped section with two bends in its structure, a two-way power divider and a four-way power divider. All of them have been modelled with a total path length of 70 mm without considering the size of transitions. This allows us to check the differences between them in return and transmission losses. The total length from inputs to outputs is 90 mm. The printing time for the complete piece was 12 hours and 14 minutes. Three pictures are shown in Figure 3-27 and Figure 3-28.

Figure 3-27: 3D-printed plastic (right) and metallized (left) GW prototype.

Figure 3-28: Bottom view of the prototype for measurements (a). The 3D-printed prototype is covered with a drilled aluminium plate. Top view of the prototype (b): Measuring set up.

3.5.1 Straight GW Prototype

Figure 3-30 shows the results obtained in the simulations and measurements for the straight GW model of Figure 3-29. All the prototypes have been measured with an Agilent 8722ES Network Analyzer after a TRL calibration with a HP R11644A WR-28 Calibration Kit.

Figure 3-29: Straight section GW prototype: simulation model (up) and prototype (down)

Figure 3-30: Measured and simulated S-parameters of the straight GGW prototype.

The measured losses in this prototype are approximately 0.2 dB in the desired band, which are very close to the simulated results of 0.15 dB. Return losses are better than 15 dB between 28 and 30 GHz and above 20 dB in most of the band. This result validates the good performance of the manufacturing technology and of the structure for the transition between WR-28 and GGW. There is a slightly displacement in frequency that can be produced by manufacturing imperfections.

3.5.2 U-shaped GGW Prototype

A comparison between simulation and measurement for the U-shaped GGW model of Figure 3-31 is depicted in Figure 3-32.

Figure 3-31: U-shaped GGW prototype: simulation model (left) and prototype (right).

The measured losses in this case are almost the same of the straight prototype. The result in return losses is very good, with a value greater than 20 dB between 28 and 30 GHz and above 25 dB in most of the band. Good agreement between measurements and simulation is observed.

Figure 3-32: Measured and simulated S-parameters of the U-shaped GGW prototype.

3.5.3 Two-Way Power Divider Prototype

The simulated and measured structures for the two-way power divider GGW model are in Figure 3-33. The associated results are in Figure 3-34.

Figure 3-33: Two-way power divider GGW prototype: simulation model (left) and prototype (right).

The power dividers prototypes present three or five ports for the two-way and four-way power divider, respectively. Therefore, several WR-28 loads have been used for matching the additional outputs when the prototype is connected to the network analyser. These loads have been measured and present a reflection coefficient better than -40 dB. Then, for a convenient comparison between simulation and measurement, we have done two different simulations: one considering an ideal perfect matching, which corresponds to the desired result, and one more for the -40 dB reflection loads, which corresponds to the measured result. The reflection level at the outputs can be configured in CST defining an open boundary in the output plane, whose reflectivity can be modified. Both simulations are represented in Figure 3-34 in order to compare the effect of the presence of loads with a reflection level of -40 dB. The mismatch supposes the presence of a small ripple in the response of 0.5 dB, which is close to the level obtained in the simulation. The result in return loss is good, although manufacturing imperfections slightly reduce its level with a value greater than 15 dB between 28 and 30 GHz.

Figure 3-34: Measured and simulated S-parameters of the 2-way power divider GGW prototype (simulation with ideal and -40 dB reflection loads).

3.5.4 Four-Way Power Divider GGW Prototype

The results obtained in the simulations and measurements for the four-way power divider GGW model in Figure 3-35 are shown in Figure 3-36.

Figure 3-35: Four-way power divider GGW prototype: simulation model (left) and prototype (right).

We have conducted the simulations for the divider with four outputs in the same way as the case of two outputs. Three loads are necessary to measure this prototype in order to match the three outputs that are not connected to the network analyser. In Figure 3-36, it can be seen that measurements and simulation are in good agreement. A ripple in S_{21} parameter is caused by the mismatch in loads. Return losses are better than 15 dB in most of the band. The same 15 dB peak at the beginning of the desired band is also found in Figure 3-34, which indicates that it is produced by errors in manufacturing process. Curves “Loads” are simulations of the power dividers considering that there are imperfectly matched loads with a reflection level of -40 dB in the non-connected ports during measurements. Simulations considering perfectly matched loads are indicated with “Ideal”.

Figure 3-36: Measured and simulated S-parameters of the 4-way power divider GGW prototype (simulation with ideal and -40 dB reflection loads).

3.5.5 Bandwidth and Loss Comparison between Prototypes

In this section, we present a comparison between the measured results. In general, the obtained results validate the good performance of our manufacturing technology. Manufacturing imperfections may be responsible of the slightly displacement in frequency that can be observed. In summary, we present a comparison of average losses in the frequency band from 28 to 30 GHz and values for fractional bandwidth at -10 dB and -20 dB in Table 3-IV.

TABLE 3-IV: MEASURED BANDWIDTH AND LOSSES

Prototype. 90 mm length	Fractional bandwidth		Average losses in 28-30 GHz band
	-10 dB	-20 dB	
Straight section	18.14 %	8.52 %	0.16 dB
U-shaped section	16.55 %	7.41 %	0.17 dB
Two-way power divider	16.10 %	4.72 %	0.22 dB
Four-way power divider	15.97 %	3.27 %	0.34 dB

We observe that bandwidth is degraded for the measurements in the U-shaped section and power dividers compared with the straight section. The bandwidth is even more degraded for the power dividers at 20 dB return losses because of the relatively narrow band of the designed power divider shown in Figure 3-6. The reflection level of the dividing structure is close to the level for the transitions to WR-28, so the combined contribution increases the reflection and reduces the bandwidth. The bandwidth for the straight sections and the U-shaped section are very similar, with slightly deviations in the position of maximums of reflection due to the presence of the bends.

Losses are very close to the simulations and their levels are in line with the simulated losses for the single designed bend and divider.

3.5.6 Estimation of the Attenuation Constant α

The effect of the transitions between GGW and WR-28 can be important in the estimation of the properties of the technology itself. Because of that, we have designed a TRL calibration kit that allows measurements removing the effect of these transitions. The TRL kit consists of the structures in Figure 3-37, which include thru, reflect and two lines with different lengths.

Figure 3-37: TRL calibration kit for our GGW structure.

The result for the straight section in Figure 3-30 after calibrating with the TRL kit is shown in Figure 3-38. Measured insertion losses and return losses are 0.1 dB and 20 dB respectively for a length of 70.2 mm. This allows the estimation of the attenuation constant α in this technology. Table 3-V presents the result, together with ideal and realistic simulation of the GGW model and measured losses for a WR-28 section. The attenuation constant for WR-28 in Table 3-V is measured for a 100 mm WR-28 standard section R896D, included in the HP R11644A WR-28 Calibration Kit. It is important to note that the measured value of 2.29 dB/m differs considerably from the theoretical value of 0.635 dB/m for an ideal WR-28 built in copper. Furthermore, attenuation constant is expected to increase if rectangular waveguide is implemented as an integrated planar structure with imperfect joints as it is shown in Figure 3-20 and Figure 3-21. Therefore, an attenuation of 1.4 dB/m in our GGW prototypes is a very good result.

Figure 3-38: Straight section after TRL calibration. Comparison between calibrated simulations and measurements.

TABLE 3-V: AVERAGE ATTENUATION CONSTANT α

Model	α (dB/m)	Model	α (dB/m)
GGW Ideal Sim.	0.916	GGW Rough Sim.	1.357
WR-28 R896D Meas.	2.29	GGW Measurement	1.4

3.6 Study of Materials and Plating Techniques

After the validation of the prototypes carried out in section 3.5, it is of interest to determine the repeatability of the manufacturing process, as well as the effect of different materials and plastic metallization techniques. In this section, several aspects of the metal-coated SLA fabrication method with different polymers and ceramic materials and metal-coating techniques are presented. The study includes a comparison of various metal-coated 3D-printed waveguide devices using groove gap waveguide technology and classical rectangular waveguide such as WR-10 and WR-28. Simulation and prototype measurements of transmission and reflection coefficients are presented at Ka and W bands. Performances in terms of feasibility, materials selection, metal coating thickness, roughness, losses and weight are also discussed.

This section presents the GGW structures designed and manufactured. We have manufactured a total of six pieces with different companies, materials, resolutions and metallization processes to determine aspects to take into account and make a comparison of the measured results with the simulations. The design is represented in Figure 3-39, which contains a part with a straight GGW section and a 4-way power divider.

Figure 3-39: Representation model of the manufactured prototypes. It includes a straight GGW section and a 4-way power divider. (a) 3D model and (b) zoom of the transition to WR-28.

As the prototypes have all been manufactured by SLA and subsequent metallization, it is important to know the available materials and metallization processes used for plastic coating, shown in the section 3.6.1. A comparison of the most critical structural details of the prototypes is provided in section 3.6.2 and the comparison between measurements and simulation with different levels of effective roughness of the metal-coated surfaces in subsection 3.6.3.

3.6.1 Selecting Materials and Metal Coating

The materials chosen are light-sensitive resins that harden in the presence of UV light, since all the pieces of this work have been manufactured using an SLA process. The selection of the base material for 3D printed prototypes can be crucial and depends on the application. There are elastic, impact resistant, waterproof or high temperature resistant materials available. For our interest, it is enough that they are rigid and can be easily post-processed to apply a thin metal layer on their surface. The materials available vary depending on the company we work with, but we have chosen those recommended by the manufacturing companies. Regarding the metallization of plastic parts, it is recommended that the thickness of the metal be greater than 7 times the skin depth of the metal at the working frequency. For reference, the skin depth of copper at Ka-band is between 0.33 and 0.4 μm , so metallization should be at least 3 μm . This paper shows a comparison of six prototypes, shown in Figure 3-40, manufactured with different materials and metallization. All the material information, layer resolution and metal coating of the different prototypes, as well as the companies that manufactured them, are shown in Table 3-VI.

Figure 3-40: Metal-coated prototypes. The identification number refers to the ID in Table 3-VI and Table 3-VII. Each prototype has one straight waveguide section 90 mm in length and a 4-way power divider with an input to output length of 106.5 mm.

TABLE 3-VI: MATERIALS AND METAL COATINGS

ID	Material	Layer Resol.	Metal coating	Companies	Country
1	Nanotool©	100 μm	Electroplating Cu 50 μm	Protolabs USA Repliform Inc	USA
2	Accura© SL 5530	50 μm	Electroplating Cu 50 μm	Protolabs USA Repliform Inc	USA
3	Accura© SL 5530	100 μm	Electroplating Cu 50 μm	Protolabs USA Repliform Inc	USA
4	Accura© SL 5530	100 μm	Electroplating Cu 10 μm	Protolabs Europe	France/ Germany
5	Material “D”	100 μm	Electroless Cu 3-5 μm	Swissto12	Switzerland
6	VisiJet© SL Clear	100 μm	Vacuum coat. Al $\ll$ 1 μm	Balentec SL WAAM SL	Spain

Of all the materials selected, Nanotool© has good thermal resistance properties and can withstand temperatures of over 200°C. This is of interest for possible component designs for satellites that must withstand high temperatures due to the incidence of sunlight. That is why we have included this material, which has now been replaced by Perform© with enhanced properties. Another important factor is the metal-coating process. In this work, we try three different processes, which are electroplating, electroless and vacuum coating technology. The electroplating technique, as its name indicates, is the only one that uses electrical charges to deposit a metallic layer on the material. As in this case it is desired to metallize plastics, the surfaces are first treated so that the metal ions can be deposited on them, which are dissolved in a chemical solution and circulate from the anode to the cathode. Therefore, the part to be metallized must be connected to the cathode. As for the electroless technique, the process is simpler because, instead of electricity, the metal is deposited by an autocatalytic reaction by immersing the part in a solution with a reducing agent. In [77], a comparison between the electroplating and electroless processes and their results is explained. Finally, the vacuum coating technology consists of the evaporation of aluminium ions that are deposited on the surface of the

piece inside a vacuum chamber. The thickness of the metallic layer obtained with this technique is less than one micrometre. More information about vacuum coating can be found in [78]. Only prototype 6 was metallized using vacuum coating, as indicated in Table 3-VI. Since the thickness of this type of coating is on the order of several nanometres, which is well below the value of 7 times the skin depth, an anomalous performance is expected.

3.6.2 Structural Details

One of the key steps in any manufacturing technology is to reproduce the designed model as faithfully as possible. In this section we present a structural analysis of the critical parts of the prototypes. Figure 3-41 shows micro-photographs of the cylindrical pins and corners of the rectangular waveguide transitions, together with the metal surface. Some samples show slight disturbances in the circularity of the pins, with a diameter of 1 mm. In general, the deviations are minimal, on the order of a few tens of microns. It is interesting to note the level of precision with which the corners of the transitions have been obtained. This ability to reproduce corners at such right angles is far superior to that available with CNC. It should be noted that the metallization achieved is smooth and very homogeneous in all cases except that of vacuum coating technology. Figure 3-41 (d,h) shows how the metal accumulates most at the transition edges and at the circular edge of the pin. This is typical of this method of metallization. In these areas the thickness of metallization is well above the manufacturer's stated value, but rapidly decreases to thicknesses below 1 μm on flat surfaces. With regard to the metal surfaces presented in the last row of Figure 3-41, the appearance of samples (i) and (j) is similar, with significant granularity but bright surfaces. The aluminium sample (l) is lacking in metallization with dark, uncoated areas. Finally, sample (k) has a very low granularity (low waviness), but the matt appearance of the surfaces could indicate a greater roughness.

Figure 3-41: Microphotographs of the prototypes. First row (a-d) shows a single pin, second row (e-h) shown the corner of the transition to WR-28 and last row (i-l) shows the metal surface at higher detail. First column (a,e,i) is associated with the prototypes manufactured in USA (ID: 1, 2 and 3), the second (b,f,j) with France/Germany (ID 4), the third (c,g,k) with Switzerland (ID 5) and the last column with Spain (ID 6).

A view of one of the transitions of each prototype is shown in Figure 3-42. It presents the structural details of the 6 prototypes in Table 3-VI. All of them have very smooth surfaces and shiny metallization. In Figure 3-42(b) and (c) we can see some undulation in the surface that may be due to surface alignment imperfections at the time of the 3D printing process. The difference in finishes due to the metal coating processes applied is notable. These smooth and homogeneous surfaces have been achieved because the pieces have been printed horizontally. However, if the part to be printed has a larger area, it may be necessary to print the part at an angle to avoid severe shrinking and bending. This was necessary for the prototype analysed in section 3.5, as the dimensions of the part were considerably larger and the plastic layers would contract during the curing process. Figure 3-43 shows a detailed image of the surfaces when the piece is printed at an angle. Cylindrical supports have been included in the prototypes in Figure 3-42 to ensure that the plastic does not deform when the flanges are screwed together to measure the parts. The prototype in Figure 3-43 did not include these supports, which made the measurement process difficult.

Figure 3-42: Detailed view of the transition to rectangular waveguide WR28 in each of the six prototypes. Letters *a* to *f* correspond to the identifications 1 to 6 in Figure 3-40, respectively.

Figure 3-43: Detailed view of the resin layers in the prototype analysed in section 3.5.

Finally, the bending of the plastic layers can be very negative. This effect is produced by the progressive curing of the plastic layers and it has greater incidence in flat and larger pieces. Some of the manufactured prototypes in this study experienced considerable bending, which coincides with those with irregularities in their surfaces. Therefore, these two types of manufacturing deviations may be related and they need further investigation. In Figure 3-44 this bending effect is compared to other pieces that have not undergone it. Fortunately, in this case,

this bending can be corrected by screwing the part tightly to a rigid aluminium plate. The result depends very much on the characterization of the processes that the company has carried out to compensate for this bending and on the material chosen. As a recommendation at the design level, it is convenient that the relationship between the largest lateral side and the height of the piece to be printed does not exceed a value of 30. On the other hand, printing the piece at an angle helps to minimize bending in return for introducing a waviness on the surface produced by the stacked layers of plastic, as in Figure 3-43.

Figure 3-44: Lateral view of prototypes 1 (a) without bending and 2 (b) bended.

3.6.3 Measurements and Comparison Between Prototypes

This section shows the measurement results of the prototypes compared with simulations. Figure 3-45 depicts the results of the straight waveguide sections, and Figure 3-46 shows the results of the 4-way dividers. Figure 3-45(a) and Figure 3-46(a) show the reflection coefficient, Figure 3-45(b) and Figure 3-46(b) the transmission coefficient and Figure 3-45(c) and Figure 3-46(c) a zoomed view of the transmission coefficient of the copper-coated pieces. Finally, Figure 3-45(d) depicts the simulation of the straight section considering different values of roughness of the copper plating to compare with the measured levels in Figure 3-45(c).

Figure 3-45: Simulations (SIM) and measurements (ID 1-6) of the straight waveguide section. Reflection (a), transmission (b), detailed view of the measured transmission (c) and a detailed view of the simulated transmission with roughness RMS values from 0 to 1.8 μm (d).

Figure 3-46: Simulations (SIM) and measurements (ID 1-6) of the 4-way power divider. Reflection (a), transmission (b) and detailed view of the transmission (c). Only one output of each is displayed for easy viewing.

A large discrepancy can be seen in the measurement of the transmission coefficient from prototype 6 with respect to the others. This result is expected, as the thickness of the metal coating of this prototype is much lower than recommended to ensure a good metallic performance. The difference in loss of prototype 6 compared to prototypes 1 to 5 is between 6 to 9 dB in the centre of the band, as shown in Figure 3-45(b) and Figure 3-46(b). Since the reflection coefficient in Figure 3-45(a) and Figure 3-46(a) is low, we conclude that most of the power injected into the prototype is dissipated into the structure. We have ensured that the losses fit very well with the simulation when considering a metal coating with a very low conductivity of $\sigma = 4 \cdot 10^3 \text{ S/m}$. The results obtained with prototypes 1 to 5 are similar, with some discrepancies in terms of losses or small frequency shifts. There is a difference of 0.25 dB in loss between prototypes 1 to 5. This

difference is mainly due to the metallization process of each piece, as the resulting roughness has a high impact on losses. After discarding prototype 6 due to insufficient metallization, Figure 3-45(c) shows that the best results are obtained with prototypes 1, 2 and 3, and the worst result with prototype 5. This implies that the roughness and conductivity of the metallization in the first case is the best, while prototype 5 presents a higher roughness. A larger variation occurs in the 4-way divider in Figure 3-46(c). This can be caused by the effect of the load terminations used in the measurement, which do not match perfectly. Table 3-VII shows the results of the average losses measured in the band of interest (28 to 30 GHz), the weight of the pieces and whether we had to correct bended pieces.

TABLE 3-VII: RESULTS AND AVERAGE LOSSES COMPARISON. GAP WAVEGUIDE PROTOTYPES

ID	Material	Bending	Loss section	Loss divider	Weight
1	Nanotool©	NO	0.23 dB	0.54 dB	115 g
2	Accura©	YES	0.18 dB	0.32 dB	95 g
3	Accura©	YES	0.29 dB	0.50 dB	95 g
4	Accura©	NO	0.33 dB	0.37 dB	65 g
5	Material D	NO	0.40 dB	0.47 dB	90 g
6	VisiJet©	NO	7.30 dB	8.41 dB	55 g

Considering that the straight section has a length of 90 mm, the average losses obtained per unit length are in a range from 2 to 4.5 dB/m, without considering prototype 6.

3.7 Printed Rectangular Waveguide WR-28 and WR-10 Comparison

This last section presents a comparison between copper-coated 3D-printed and commercial sections of WR-28 and WR-10 waveguides as a complement to the measurements made for the prototypes based on GGW in section 3.6. The pieces have been manufactured using copper-coated SLA in Swissto12. Sections WR-28 and WR-10 printed and commercial are shown in Figure 3-47.

Figure 3-47: A 100 mm long WR28 section printed in 3D, next to a 100 mm commercial section: side view (a) and front view (c). A 31 mm long section of WR-10 printed in 3D, next to a 35.1 mm long commercial section: side view (b) and front view (d).

The measured results of the waveguides in Figure 3-47 are shown in Figure 3-48. Figure 3-48(a) depicts the response of the WR-28 waveguides and Figure 3-48(b) the response of the WR-10 waveguides. The results obtained are summarized in Table 3-VIII, which indicates the average losses per unit length in each of the waveguides and the weight in grams of each piece. The behaviour of the 3D-printed sections is good, although equivalent roughness differs between prototypes. In the case of WR-28, the 3D-printed prototype has lower losses than the commercial section. This difference fits with simulations of ideal WR-28 sections considering copper with a roughness of $0.4\ \mu\text{m}$ for the 3D-printed piece and aluminium with $1\ \mu\text{m}$ of roughness for the commercial section, as shown in Figure 3-48(a). For the case of the WR-10 sections in Figure 3-48(b), there is a good agreement between the measurement of the commercial section and the ideal simulation with gold and a roughness of $0.075\ \mu\text{m}$. The measurement of 3D-printed WR-10 section fits well with a slightly modified WR-10 section that differs from $2.54 \times 1.27\ \text{mm}$ to $2.42 \times 1.22\ \text{mm}$ and rounded corners with a radius of $0.15\ \text{mm}$. These values have been obtained considering the mean values of the measurements of the two flanges of the 3D-printed WR-10 section using a microscope. Figure 3-49 presents the microphotography of this straight section with the measured values. Simulated curves of the ideal and modified WR-10 sections are shown in Figure 3-48(b) with a roughness of $0.55\ \mu\text{m}$. Of particular interest is the weight reduction of plastic parts, which is an advantage for satellite-embedded equipment or other applications that require lightness.

Figure 3-48: Measurement comparison of the response of WR-28 (a) and WR-10 (b) pieces with simulations fitted with the material and roughness. The curve *Cu 0.55 μm modif.* corresponds to the WR-10 structure with modifications of width, height and rounding of corners in accordance with the measured dimensions in Figure 3-49.

Figure 3-49: Microphotographs of the input of the 3D-printed WR-10 section with the measured values of height, width, and rounded corners.

TABLE 3-VIII: RESULTS AND AVERAGE LOSSES COMPARISON. WR-28 AND WR-10 PROTOTYPES

Device	Av. Losses	Weight
<i>Commercial WR28</i>	2.33 dB/m	110 g
<i>3D-Printed WR28</i>	1.17 dB/m	70 g
<i>Commercial WR10</i>	3.48 dB/m	15 g
<i>3D-Printed WR10</i>	10.48 dB/m	3 g

3.8 Conclusion

This chapter shows that it is possible to obtain low cost designs for high frequencies distribution networks with metallized 3D-printing technology without sacrificing a good performance. Copper plated plastic combines the good electrical conductivity of the copper and the low mass density of the plastic material. This allows obtaining a manufactured piece with lower weight and better conductivity than in machined aluminium. In addition, the reduced manufacturing time using 3D printing suppose an advantage for prototyping. However, plastic presents a relatively low deflection temperature and a considerably different coefficient of thermal expansion compared to copper or aluminium. Because of that, this technology is well suited for terrestrial environments, but manufacturing with this kind of plastics is not recommended for spatial applications. Nevertheless, other plastic polymers with a high thermal stability may be used for that end. A deep analysis of the manufacturing technology has been conducted, compared with conventional milling techniques. The study of the effect of several physical properties highlights that roughness takes a fundamental role in losses at high frequencies. The importance of achieving good electrical contacts in rectangular waveguide makes the groove gap waveguide a suitable alternative. Several prototypes in groove gap waveguide technology have been designed and manufactured in metallized 3D-printed technology. The measured results strongly agree with simulations, presenting a bandwidth better than 15% at -10 dB centred at 29 GHz. The losses per unit length in this technology have been calculated using a TRL calibration kit and they are very close to the losses in a typical rectangular waveguide with an average value of 1.4 dB/m between 28 GHz and 30 GHz, but with a fraction of the weight and cost.

In addition, a comparison of various metal-coated 3D-printed waveguide devices using groove gap waveguide technology, WR10 and WR28 were conducted. The results obtained are very good in most of the cases, with low losses and a great similarity between measurements and simulation. We emphasize the importance of the metal-coating process of the plastic structures, as an insufficient thickness of metallization prevents the correct guidance of the electromagnetic fields. The roughness of this metal layer has a strong influence on total losses, so it is desirable to achieve glossy surfaces. Finally, the comparison between commercial and 3D printed WR-28 and WR-10 waveguides makes clear the great potential of additive manufacturing in producing millimetre-wave components of low cost, low weight, low loss and great design versatility.

Chapter IV

3D-Printed Monopulse RLSA Antenna for Space Debris Radar at 94 GHz

4.1 Introduction

New applications of communications systems and radar at millimetre-wave frequencies are appearing in multiple areas of telecommunications. These applications demand viable solutions for deploying antenna systems where low losses, reduced cost, ease of assembly and low weight are of fundamental interest. These new applications have seen the light due to the assignment of new spectral bands in millimetre wave frequencies for communications and radar systems, among others. In particular, the frequency bands 71-76 GHz, 81-86 GHz and 92-95 GHz have been proposed in the W-band for high-speed point-to-point wireless communications [79]. Thanks to the reduction in size, in these bands it is possible to obtain highly directional antennas, which makes it possible to engineer systems that are very close to each other without causing interference [80]-[82]. In addition, radar systems can also benefit greatly from this increased directivity in antenna systems to enhance resolution and improve detection of smaller targets at greater distances. A clear application of these capabilities is the detection of space debris, in which great research efforts are being made [83]-[87]. The aim is to address the problem of the accumulation of debris from disused satellites and small particles a few centimetres in diameter from old collisions between satellites travelling at thousands of kilometres per hour. The collision of some of these particles with active satellites or with space exploration stations could lead to their destruction. This topic is becoming a challenge due to the increasing number of elements in orbit around the earth, coming from old satellites. According to the European Space Agency [83], objects from 1 to 10 cm of diameter, are the most potentially harmful since they are too small to be detected individually. A consortium of research groups in Madrid (from Universidad Politécnica de Madrid, Universidad Carlos III and Universidad Autónoma de Madrid) joined their efforts to work on a research project with the purpose of the research on RF technologies for space debris radar applications at millimetre waves.

Monopulse radar systems are currently being developed at 94 GHz [88]-[92]. One of the typical problems of antennas at these frequencies is the high losses introduced by commonly used dielectric materials employed by substrates. Microstrip is a clear example of cheap technology, easy to access and simple to manufacture and design, but due to the properties of the available dielectric substrates, the efficiency of the antennas is greatly reduced by the high losses. Other substrate-based technologies tried to improve this efficiency. There are works that use low temperature cofired ceramics (LTCC) technology to reduce losses in antennas based on microstrip distribution networks [93]. This technology is characterized by using a ceramic substrate with a very high dielectric constant and low losses, making it interesting for high frequencies. LTCC allow the production of small pins or via holes using several stacked layers. However, different

power distribution levels complicate the design and still introduces losses due to the presence of dielectric. Other works based on LTCC in millimetre-wave bands can be found in [94], [95].

Another widely used substrate-based technology is substrate integrated waveguide (SIW), which consists in the implementation of a waveguide in a substrate replacing the lateral walls with metallic vias. These metallized vias must be separated up to a certain distance to avoid leaks. Some W-band designs use this technology [89], [96]. However, despite their low cost and ease of manufacture, the distribution networks implemented are often long and sometimes multi-layered, so the losses in the dielectric are high. The result is a very negative impact on the overall efficiency of the antenna, as shown by the examples [89], [96] with gains of 30 and 25 dBi, but efficiencies of 15% and 25%, respectively.

A solution to the problems presented by designs based on dielectric substrate is the use of all-metal antenna systems based on waveguide. This maximizes efficiency due to the low losses provided by these types of structures but makes the manufacturing and assembly process difficult. For complex flat networks based on rectangular waveguide, fabrication is normally done in two or more parts. Subsequent assembly must ensure very good electrical contact between the metal walls along the entire structure, something that is difficult and expensive to achieve for high frequencies due to manufacturing tolerances. Diffusion bonding is a monolithic metal device manufacturing technology that consists of stacking thin metal layers with the desired patterns. Specific pressure and temperature conditions are applied to these layers, allowing them to weld into a single metal part. In this way, it is possible to manufacture antennas with intricate feeding networks in waveguide with a very good electrical contact between layers [97]-[100]. Despite its good performance and design versatility, this technology is very expensive and the resulting parts are very heavy because they are made of solid metal, especially if they are made of copper.

As a common solution to all these problems, this chapter proposes a monopulse antenna designed and manufactured in two layers. The upper layer contains the radiating elements in a Radial Line Slot Array (RLSA) configuration. The lower layer consists of a power distribution network based on a Butler matrix implemented in gap waveguide technology. We considered to use CNC machining to build the Butler matrix, but the problem with the periodic structures when designs are made at higher frequencies is that they require a lot of precise milling time, so a gap waveguide design in W band requires thinner drills and a longer manufacturing time. In Chapter III, we proposed additive manufacturing as an alternative to classic CNC machining. Typically, high-performance 3D-printers use plastic resins or powder particles of plastic or metal that harden and bind together to form the desired part. Some works demonstrate that 3D-printing provides good results in millimetre band designs [19][22][24][31]. Most of them are based on circular or rectangular waveguides and are manufactured in one piece. This can be a problem when evacuating resin or metal/plastic powder residue from the interior of intricate structures and makes internal metallization of plastic devices difficult. Several gap waveguide designs at Ka band avoided these problems using additive manufacturing [101], [102], including the study conducted in Chapter III. However, printing tolerances may affect depending on the frequency band, especially when printing small pins. 3D-printing research applied to gap waveguide structures at W band has not been reported yet, so this work also aims to verify the behaviour of this 3D-printing technology in W-band gap waveguide designs.

This chapter is divided into five sections: section 4.2 presents the antenna scheme with the desired behaviour of the RLSA antenna and the Butler matrix with the basic components that are part of it. Section 4.3 details the design and experimental validation of the Butler matrix and all of its individual components. Section 4.4 presents the design and prototyping of the coupling from the Butler matrix to the RLSA antenna. The integration of feeding network and the RLSA antenna into the full monopulse antenna is discussed in section 4.5 and conclusions are drawn in section 4.6.

4.2 Antenna Scheme

Due to the complexity of the design, the development and validation of the monopulse antenna has been carried out in two parts: the design of the RLSA antenna and the interface with the Butler matrix with the design of the Butler matrix itself.

4.2.1 Monopulse RLSA Antenna Scheme

RLSA antennas consist of the radial propagation of fields between two parallel plates that are radiated through slots conveniently distributed over the surface of at least one of these plates. The distribution of these slots determines the performance of the antenna. It is possible to achieve an arrangement that cancels the reflections produced in the centre of the radial guide and thus minimizes the reflection coefficient of the antenna. With a structure of slots arranged in concentric rings it is possible to generate the sum (Σ) and difference (Δ) patterns required for monopulse operation. An azimuthally uniform phase mode (Figure 4-1(a)) within the RLSA results in a Δ beam and a rotary phase mode (progressive linear phase in Figure 4-1(b)) generates the Σ beam. The elevation angle (θ) is obtained by comparing the amplitude of the signal received by both beams, while the azimuth angle (ϕ) is obtained from the comparison of phases, since the phase is constant in azimuth for the Δ pattern and linear for the Σ pattern. Previous studies for monopulse antennas based on RLSA were presented in [103].

Due to the complexity of the problem, the antenna design has been divided into two parts: the radiating structure and the feed structure. A 10-ring RLSA antenna (45 mm in diameter) has been used to validate the design of the feed structure. The complete monopulse antenna is left hand circularly polarized with two waveguide ports, one for the Σ pattern and a second one for the Δ pattern, connected to the Butler matrix.

Figure 4-1: Uniform mode pattern (a) for difference beam configuration and rotary phase pattern (b) for sum beam configuration.

4.2.2 Butler Matrix Scheme

As will be seen in the section 4.4.1.2 dedicated to the cavity coupled to the RLSA, it is necessary to design a Butler matrix that generates in its outputs two different phase configurations to excite the Σ and Δ patterns of the antenna. Butler matrices have been mainly used to feed antenna arrays and get an electronic steering of the beam by modifying the shifting that each phase shifter introduces in its internal branches [104]. In this case, we need to combine hybrids 3dB-90°, 90° phase shifters and other components such as bends and crossovers to get a configuration with all the outputs in phase for the Δ pattern and another with the phases shifted 90° sequentially for the Σ beam. The in-phase configuration will generate a radially uniform field distribution within the cavity, while the sequentially 90-degree phase-shifted configuration will result in a rotary field distribution. A schematic of a Butler matrix meeting these conditions is shown in Figure 4-2. The two first 3dB-90° hybrids of the typical Butler matrix were substituted by power dividers because only two of the inputs are used. The corresponding 90° phase differences are rearranged along the scheme using 90° phase shifters. In this way, we avoid the use of loaded terminations in the two unused inputs. In [103], those loads were included in the microstrip design, but they were cheap and small compared to WR-10 loads.

As can be seen in Figure 4-2 (ignoring the phase shifters which are a perturbation of the waveguide), the Butler matrix has a horizontal symmetry axis, but not vertical. The path lengths between each input port and each output port are different, so it is necessary to introduce a phase adjustment in some sections to obtain the desired phases in the outputs. The design of a horizontally and vertically symmetrical Butler matrix in which all paths are the same length was discarded. The problem in this case is the size of the resulting Butler matrix, with the consequent increase in transmission losses and manufacturing cost. As the application is relatively narrow band (less than 0.5 GHz) we preferred to implement the smaller design.

Figure 4-2: Butler matrix scheme.

In Figure 4-2, the four outputs of the Butler matrix excite the circular cavity forming 90-degree angles to each other. The Butler matrix outputs are labelled as A, B, C and D in Figure 4-2. The phases obtained in these outputs for the Σ and Δ modes are shown in Table 4-I. Both Butler matrix and circular cavity will be implemented using gap-waveguide technology. Cross paths in Figure 4-2 have been implemented with crossovers based on hybrids 3dB-90°.

TABLE 4-I: PHASE AT THE OUTPUTS OF THE BUTLER MATRIX

Output phase	Outputs			
Input	A	B	C	D
Σ	0°	-90°	-180°	-270°
Δ	-90°	-90°	-90°	-90°

4.3 Design and Characterization of a Butler Matrix Prototype

This section presents the design, simulation and measurement of each of the components that are part of the Butler matrix, as well as their interconnection to build the final Butler matrix.

4.3.1 Design of the Butler Matrix

4.3.1.1 Base Groove Gap Waveguide Structure

To begin with the Butler matrix design, we have selected the basic unit cell from which all components will be built. The GGW unit cell has a periodicity $p=1.2$ mm and three rows of pins on the sides of the gap waveguide. Figure 4-3 (a) shows the dimensions of all the details of the structure. The separation $W_g=3.25$ mm between the inner pins of the waveguide has been chosen to minimize the losses and dispersion introduced by the structure taking into account that the cut-off frequency of the next propagated mode is at least 5% above the design frequency (94 GHz). In the dispersion diagram in Figure 4-3(b), a TE_{10} mode propagates from 64.5 GHz to 99 GHz in a single-mode band configuration. The number of rows of pins at the sides of the gap waveguide has been chosen according to the transmission losses and the coupling between parallel gap waveguides considering paths with 50 periodic pins in length. The results obtained in simulation with 1, 2, 3 and 4 rows of pins are represented in Figure 4-4.

Figure 4-3: GGW unit cell. Structure (a) and dispersion diagram (b).

In Figure 4-4(a), it is clear that the behaviour is not good when only one row is used, but it is very good in the rest of cases. This is because, using a single row, the periodicity effect in

the transverse direction is lost and the blockage of waveguide gap propagation is considerably reduced. In the level of coupling between parallel guides of Figure 4-4(b), greater differences are observed. Acceptable coupling levels (below -40 dB) are achieved from 3 rows of pins on each side. Therefore, the base structure has been designed with 3 rows of periodic pins.

Figure 4-4: Simulations of loss (a) and coupling (b) between parallel gap waveguide with 1, 2, 3 and 4 rows of pins at both sides.

An important quality to consider in the design is the roughness of the copper surfaces obtained in the manufacturing process. As noted in Chapter III, the roughness of the metallization process carried out by our supplier is around $0.3 \mu\text{m}$. Low roughness values reduce the equivalent conductivity of the metal used, which in this case is copper. Several simulations with different roughness values have been carried out to observe their effect on transmission losses and phase. Figure 4-5 represents these values per centimetre for four different copper roughness values from $0 \mu\text{m}$ (ideal) to $1 \mu\text{m}$. Figure 4-5(a) shows that a roughness of $1 \mu\text{m}$ can triple the losses per unit of length and introduce a phase-shift between 4 and 7 degrees per centimetre. This effect on the phase is of great importance, given that in the design of a Butler matrix we seek to achieve the minimum phase error between the outputs. Taking as reference the measurements in Chapter III, we took a roughness of $0.3 \mu\text{m}$ for the design of the Butler matrix components.

Figure 4-5: Simulations of loss (a) and phase shift (b) depending on the roughness RMS of the copper surfaces.

4.3.1.2 WR-10 to GGW Transition

The Butler matrix has two input ports and four output ports, which are connected to interfacing equipment (MS4647 VNA Anritsu) via 3743A Anritsu mmW Module up to 110 GHz. The input and output ports must be connected to WR-10 flanges in the measurement process. It is therefore necessary to design a transition from waveguide WR-10 to the GGW structure used in this design. The transition is designed considering that the WR-10 flange will be connected perpendicularly to the plane of the Butler matrix. This is intended to facilitate connection and prevent the feed elements from affecting the radiation pattern by placing them on the back of the antenna. The use of additive manufacturing provides greater versatility and more degrees of freedom at the design stage. Therefore, the design of the transition has focused on reducing as much as possible the reflection in the 91 to 97 GHz band using an arc-shaped structure. All design details and millimetre values are given in Figure 4-6 where cover plates have been removed to expose internal features. The manufacturing of this arc is feasible by means of 3D-printing, unlike traditional CNC machining. As a result, a reflection coefficient below -40 dB was obtained in the band studied (Figure 4-7).

Figure 4-6: Design details and values of the structural parameters of the transition to WR-10. All values are expressed in millimetres.

Figure 4-7: Simulation of the S parameters of the transition to WR-10.

4.3.1.3 Right-Angled Bend

The bend design is very simple. It is based on a circular rotation with a radius equal to the width of the GGW ($R=3.25$ mm). With the structure of Figure 4-8, a reflection below -40 dB in the band of interest is achieved without the need for an optimization process. S-parameters are represented in Figure 4-9.

Figure 4-8: Design details and values of the structural parameters of the right-angled bend. All values are expressed in millimetres.

Figure 4-9: Simulation of the S parameters of the right-angled bend.

As indicated in section 4.2.2, the ultimate goal of the design of this Butler matrix is its coupling to a circular cavity connected to an RLSA antenna. To achieve this connection without breaking the symmetry with respect to the horizontal axis it is also necessary to design 45° bends. To do this, we have modified the structure of the 90° bend in Figure 4-8 so that it has a mirror symmetry along the diagonal while keeping the unit cell structure of Figure 4-3(a) along this axis of symmetry. This makes it possible to connect this element to any other, since the junction point is always the unit cell. Due to limitations in the orientation of waveguide ports in CST, we simulated two consecutive 45° bends (Figure 4-10) whose S-parameters are similar to those presented in Figure 4-9.

Figure 4-10: Design details and values of the structural parameters of the 45° bend.

4.3.1.4 90-Degrees Phase Shifter

For the 90° phase shifter design, we have modified the bend in Figure 4-8 to introduce an additional phase shift of -90° . With this solution we avoid having to add an additional phase shifting element in the chain, which would mean an increase in length, but it is only enough to replace a normal bend by this new 90° phase shifting bend. In order to add this extra phase shift, the total length of the fields inside has been slightly increased. In Figure 4-11, the structures of both the original bend (Figure 4-11(a)) and the 90° phase shifter (Figure 4-11(b)) are shown. The results of the simulation are depicted in Figure 4-12: magnitude of the S parameters (Figure 4-12(a)) and the introduced phase shift (Figure 4-12(b)).

Figure 4-11: Design details and values of the structural parameters of the original bend (a) and the 90° phase shifter (b). All values are expressed in millimetres.

Figure 4-12: Simulation of the S parameters of the 90° phase shifter: Magnitude (a) and phase shift (b).

4.3.1.5 Power Divider

The power divider has been designed in a Y-shape as its outputs must be parallel. The width of this design is smaller than for a T-divider with two right angle bends for parallel outputs. The details of the design and the S-parameters obtained after an optimization process are shown in Figure 4-13 and Figure 4-14. We know from basic circuit theory [4] that in a three-port, reciprocal, lossless device such as this one, it is impossible to match all ports simultaneously. However, since ports 2 and 3 are excited simultaneously when operating in reception, a good combined matching is obtained.

Figure 4-13 Design details and values of the structural parameters of the power divider. All values are expressed in millimetres.

Figure 4-14: Simulation of the S parameters of the power divider: Input reflection (a) and transmission and output reflection (b).

4.3.1.6 Hybrid Coupler 3dB-90°

The 3dB-90° hybrid coupler design is based on existing design implemented in waveguide [105]. The main design parameters are width and length calculated from expressions in [106]. Considering a value of $n = 3$ in equations (4-1) and (4-2), where k is the wavenumber, the calculated width and length for the hybrid coupler are $w = 5.60$ mm and $l = 5.83$ mm. The position of the pins (Figure 4-15) has been adjusted to obtain greater bandwidth. Both the reflection and isolation between parallel ports presents a band greater than 5% (Figure 4-16(a)) with a coupling to the output ports crossing at the design frequency at a level of -3.1 dB (Figure 4-16(b)) due to losses in copper. The phase difference between outputs is practically constant throughout the 91 to 97 GHz band, varying between -89° and -90° (Figure 4-16(c)).

$$w = \frac{\pi}{k} \sqrt{\frac{4(3n+1)(n+1)}{4n+1}} \quad (4-1)$$

$$l = \frac{\pi}{k} \sqrt{\frac{(3n+1)(n+1)}{3}} \quad (4-2)$$

Figure 4-15: Design details and values of the structural parameters of the hybrid 3dB-90°. All values are expressed in millimetres.

Figure 4-16: Simulation of the S parameters of the hybrid 3dB-90°: Reflection and isolation (a) transmission (b) and phase difference at outputs (c).

4.3.1.7 Crossover

The design of a crossover is usually done by concatenating two 3dB-90° hybrids [106]. This structure allows two transmission lines that follow different trajectories to cross keeping the coupling between them at a minimum level. As can be seen in Figure 4-17, the crossover has been designed in a similar way to the hybrid, considering a value of $n=7$ in equations (4-3) and (4-4). The calculated width and length are $w = 5.77$ mm and $l = 12.45$ mm. To improve the adaptation and minimize the coupling, the position of the pins has been slightly optimized. The simulation results are shown in Figure 4-18. We added two small pins 0.2 mm high in the centre of the crossover. These pins have been introduced to eliminate a resonance appearing at 92.15 GHz (see Figure 4-18(b)). A similar bandwidth to that obtained with the hybrid is observed, with an isolation better than -40 dB in the direct output and a transmission to the diagonal output around -0.1 dB due to losses in the copper.

$$w = \frac{\pi}{k} \sqrt{\frac{(6n+1)(2n+3)}{8n}} \quad (4-3)$$

$$l = \frac{\pi}{k} \sqrt{\frac{(6n+1)(2n+3)}{12}} \quad (4-4)$$

Figure 4-17: Design details and values of the structural parameters of the crossover. All values are expressed in millimetres.

Figure 4-18: Simulation of the S parameters of the crossover: Reflection and isolation (a) and transmission (b).

4.3.1.8 Propagation Constant Adjustment for Tuning Phase

This subsection details the phase adjustment of the Butler matrix branches. This adjustment is necessary due to the difference in length of the branches of the Butler matrix, as already mentioned in section 4.2.2. Phase correction has been carried out by adjusting the phase constant of sections $S1$, $S2$, $S4$ and $S5$ of the complete Butler matrix in Figure 4-19(a). Sections $S3$ and $S6$ have been kept fixed. The remaining components of the Butler matrix are indicated in Figure 4-19(a) with the letters: T (WR-10 transition), D (power divider), B (bend 90°), P (90° phase shifter), C (crossover) and H (hybrid coupler). Several types of bend are shown in the design: B_s for a shorter bend and B_h for the half bend with 45° rotation. To facilitate matching line lengths, the bend B^* will be substituted by a 90° phase shifter P . The adjustment of each section consists of modifying the width of the gap waveguide. Figure 4-19 (b-c) shows the paths that differ in length. For a narrow-band application such as this, the necessary adjustment for paths X and Y on the one hand and Z and W on the other must meet that the wrapped phase introduced at 94 GHz is the same. Therefore, the expressions (4-5) and (4-6) must be met.

(a)

Figure 4-19: Complete design of the Butler matrix (a), detail of the X and Y paths (b) and of the Z and W paths (c) for phase adjustment.

$$Ph(X) - Ph(Y) = n \cdot 360 \quad (4-5)$$

$$Ph(W) - Ph(Z) = n \cdot 360 \quad (4-6)$$

, being n an integer number and:

$$Ph(X) = Ph(S1) + Ph(B) + Ph(Bs) + Ph(S2) \quad (4-7)$$

$$Ph(Y) = Ph(Bs) + Ph(B^*) + Ph(C) \quad (4-8)$$

$$Ph(Z) = Ph(C) + Ph(Bh) + Ph(S6) \quad (4-9)$$

$$Ph(W) = 2 \cdot Ph(B) + Ph(S3) + Ph(S4) + Ph(Bh) + Ph(S5) \quad (4-10)$$

The use of $n \neq 0$ results in a narrower band. Because one of the design choices was to maintain symmetry vertically but not horizontally, some paths must be longer than others. This prohibits using $n=0$, but as the application is narrow band (less than 0.5 GHz), we assume this condition. After unwrapping the phase of each one of the components of the analysed paths, the final phase introduced is obtained. The initial result *X-Y with Bends* and *W-Z* is represented in Figure 4-20. Keeping the widths of sections *S3* and *S6* fixed at a value of $W_g = 3.25$ mm, it is necessary to modify the width of sections *S1* and *S2* to a value of 2.79 mm and sections *S4* and *S5* to a value of 3.6 mm. It was observed that it is possible to minimize the width variation of sections *S1* and *S2* to 3 mm by replacing the B^* bends with 90° phase shifters (see curve *X-Y with P* in Figure 4-20). The final phase difference represented in Figure 4-20 with solid lines is 0 (+ $n \cdot 360$) degrees at 94 GHz.

Figure 4-20: Phase corrections between paths X and Y and paths W and Z.

TABLE 4-II: WIDTH AND LENGTH OF SECTIONS $S1$ TO $S6$

Section	Width	Length	Section	Width	Length
$S1$	3.00 mm	8.40 mm	$S4$	3.6 mm	28.98 mm
$S2$	3.00 mm	14.04 mm	$S5$	3.6 mm	7.94 mm
$S3$	3.25 mm	8.40 mm	$S6$	3.25 mm	10.93 mm

The final values of width and length of the section $S1$ to $S6$ are shown in Table 4-II for the phase correction of -81.79° between X and Y paths and 132.3° between W and Z paths. To verify that the phases obtained in the outputs of the complete Butler matrix are as expected, we have calculated the total phase shift introduced by all components. The results obtained for input port 1 (Figure 4-21(a)) show a sequential phase difference of 90° . For port 2 in Figure 4-21(b), the value of the phases at the outputs coincides. A maximum error of less than 2° has been found in the simulation so we validate the design for its manufacture.

Figure 4-21: Phases at the output ports for input ports 1 (a) and 2 (b) after phase adjustments. Phase differences are referenced to the phase at the Port 6.

4.3.2 Experimental Validation of the Butler Matrix and Its Components

This section presents the results obtained both in simulation and in measurement of the manufactured prototypes. The prototypes were manufactured using SLA. The validation of prototypes includes the individual components of the Butler matrix to verify their behaviour. Subsection 4.3.2.1 presents only the prototypes of individual components, while measurements and simulations of the complete Butler matrix are shown in subsection 4.3.2.2.

4.3.2.1 Individual Components

All components of the Butler matrix are manufactured in two pieces. The piece in Figure 4-22(a) contains a short straight section, a right angle bend, a 45° bend, a 90° phase shifter and a 3dB- 90° hybrid. The prototype in Figure 4-22(b) presents a long straight section, a power divider and a crossover. The difference in length between the short and long straight sections is 24 mm, so it is possible to obtain an estimate of the measured losses per unit length of the gap waveguides. The images in Figure 4-22(c-d) show a manufactured transition to WR-10 in detail.

Figure 4-22: Pictures of the prototypes manufactured: Piece with bend, 45° bend, phase shifter, short straight section and hybrid 3dB-90° (a). Piece with power divider, long straight section and crossover (b). Detail of the manufactured WR-10 transition: bottom view (c) and top view (d).

All the measurements of the prototypes presented in this section show a high degree of similarity to the simulations. One of the critical points in the prototypes is the interface between the used GGW structure and the WR-10 flange. Figure 4-23 and Figure 4-24 depict the simulated and measured S parameters of the short and long straight sections, respectively. The measured reflection coefficient is 15 dB higher than the simulation level, but remains below -20 dB in a 20 GHz band. In relation to the transmission in Figure 4-23(a) and Figure 4-24(a), there are very low losses up to 100 GHz, from which the pins lose the ability to block propagation as seen in the dispersion diagram in Figure 4-3(b). In Figure 4-23(a), some measured values are above 0 dB around 80 GHz that are produced by small alignment error during the calibration and measuring processes. In Figure 4-25, taking into account the length difference of 24 mm between straight sections, we estimate losses between 0.05 and 0.1 dB/cm that fit well with the losses obtained in simulation considering a copper metallization with a roughness of 0.3 μm .

Figure 4-23: Simulated and measured S parameters of the short straight section: S_{21} (a) and S_{11} (b).

 Figure 4-24: Simulated and measured S parameters of the long straight section: S_{21} (a) and S_{11} (b).

Figure 4-25: Estimated losses per centimetre derived from the transmission coefficient for the short and long straight sections (24 mm difference).

The results of the 90° bend and the 45° bend in Figure 4-26 and Figure 4-27 are also very satisfactory. The measured reflection is below -20 dB from 80 GHz to 100 GHz with a transmission coefficient around -0.2 dB. The same applies to the 90° phase shifter, although in this case the matching level shown in Figure 4-28(b) is slightly worse in both measurement and simulation. The phase difference between the right angle bend and the 90° phase shifter is represented in Figure 4-29. The measurement of the phase difference fits very well with the simulation and with the result obtained in the design of Figure 4-12. The measured phase difference at 94 GHz is 91° , with a variation of $\pm 5^\circ$ between 91 and 97 GHz.

 Figure 4-26: Simulated and measured S parameters of the right-angled bend: S_{21} (a) and S_{11} (b).

 Figure 4-27: Simulated and measured S parameters of the 45° bend: S_{21} (a) and S_{11} (b).

 Figure 4-28: Simulated and measured S parameters of the phase shifter: S_{21} (a) and S_{11} (b).

Figure 4-29: Simulated and measured phase shift of the phase shifter.

With respect to the power divider, the comparison of simulation and measurement is shown in Figure 4-30. Both the level and shape of the measured S parameters meet the simulation expectations and the reflection coefficient is maintained at around -20 dB in an 8 GHz bandwidth. The measurements for the 3dB-90° hybrid in Figure 4-31 and Figure 4-32 and the crossover in Figure 4-33 show a small deviation of about 0.5 GHz upwards in frequency with respect to the simulation. However, the performance of both devices is acceptable with a band of about 8 GHz with a reflection below -15 dB and an isolation better than 20 dB for the hybrid coupler and better than 30 dB for the crossover at 94 GHz. Due to the frequency shift, the hybrid coupling has an unbalance of 0.4 dB at 94 GHz. The phase difference between its outputs in Figure 4-32 is around 90° with a maximum error of 5° in an 8 GHz band. Finally, the crossover transmission parameter represented in Figure 4-33(c) has a level around -0.3 dB at 94 GHz due to copper losses and a small resonance at 92.35 GHz. This value coincides with the resonance found in the design of the crossover without pins (see Figure 4-18(b)).

Figure 4-30: Simulated and measured S parameters of the power divider: S_{21} , S_{31} (a) and S_{11} (b).

Figure 4-31: Simulated and measured S parameters of the power divider: S_{31} , S_{41} (a) and S_{11} , S_{21} (b).

Figure 4-32: Phase difference between outputs 3 and 4.

Figure 4-33: Simulated and measured S parameters of the crossover: S_{11} (a) S_{21} , S_{31} (b) and S_{41} (c).

4.3.2.2 Final Butler Matrix

This subsection presents a comparison between simulation and measurement of the final design of the Butler matrix and the manufactured prototype in Figure 4-34. The results obtained in simulation and measurement are depicted from Figure 4-35 to Figure 4-37. The reflection and isolation parameters between the input ports are shown in Figure 4-35 with a reflection coefficient below -15 dB in a 6 GHz bandwidth and an isolation better than 20 dB. At 94 GHz, the reflection is below -25 dB at both input ports and the isolation between them is greater than 30 dB.

Figure 4-34: Picture of the complete manufactured Butler matrix.

Figure 4-35: Simulated and measured S parameters of the Butler matrix: S_{11} & S_{22} (a) and S_{21} (b).

Figure 4-36 shows the results of the transmission parameters from the input ports to the output ports. Figure 4-36(a-b) presents the simulation and measurement of the transmission from port 1 and Figure 4-36(c-d) the transmission from port 2. Two resonances appear around the frequencies 92.5 GHz and 95.5 GHz in simulation and measurement. The reasons for the appearance of these resonances are discussed in detail in section 4.3.3 and solutions are proposed to avoid them. Both resonances are outside the working band, which shows an average output level of -6.8 dB. The additional 0.8 dB is due to losses in copper metallization. The minimum amplitude error appears at 94.17 GHz with values below ± 0.44 dB for the port 1 and ± 0.3 dB for the port 2.

Figure 4-36: Simulated and measured S parameters of the Butler matrix: Simulation of transmission from port 1 (a) and port 2 (c) to outputs. Measurement of transmission from port 1 (b) and port 2 (d) to outputs.

The full wave simulations of the complete Butler matrix presented in Figure 4-37(a) and Figure 4-37(c) are practically identical to the results obtained in the design stage of section 4.3.1.8 in Figure 4-21 by grouping the phase shifts introduced by each component. In the measurements in Figure 4-37(b) and Figure 4-37(d), phase deviations present a maximum error of 8° for difference configuration and 7° for sum configuration at 94 GHz. However, due to the asymmetry of the path lengths, the phase balance degrades rapidly when separated from 94 GHz with an approximate ratio of 30° per GHz. Apart from that, there is a ripple of ± 1 dB in amplitude and $\pm 7^\circ$ in phase which is mainly due to the great difference in lengths between paths. Compared to simulations, this ripple increased due to errors in manufacturing and measurement processes, but only by ± 0.3 dB extra in amplitude and $\pm 1.5^\circ$ extra in phase. Despite this, the prototype fulfils well the desired conditions at the outputs for both input ports between 93.75 GHz and 94.4 GHz with a maximum error of 1.5 dB in amplitude and 20° in phase.

Figure 4-37: Simulated and measured phases at the outputs of the Butler matrix: Simulation for port 1 (a) and port 2 (c). Measurement for port 1 (b) and port 2 (d). Phase differences are referenced to the phase at the Port 6.

Table 4-III shows a comparison with other feeding networks designs for monopulse operation or beamforming. The elements included in the comparison are the type of comparator or beamformer used, the number of pieces, the type of transmission line, the manufacturing technology used, the size, characterization of S parameters and performance of the network in terms of amplitude and phase balance and insertion losses.

The first thing to keep in mind is the difference between designs with magic tee and Butler matrix. The proposals with magic tee present good performance in all aspects, but this type of structures typically require a multi-level distribution network. Butler matrices can be designed on a single level, reducing manufacturing complexity and cost. In addition, the proposals with magic tee [92], [107]-[109] are made of metal with expensive manufacturing processes. Only the case in [108] exploits the benefits of 3D printing with DMLS technology. However, the authors claim that the roughness of the printed surfaces is 25 μm using the aluminium alloy AlSi10Mg, whose conductivity is $2.04 \cdot 10^7$ S/m. With these properties of the material, the losses obtained at the frequency of 15 GHz of the design in [108] are of 0.07 dB/cm, but in band W would rise to around 1 dB/cm. For comparison, losses measured in our prototype are around 0.05 dB/cm at 94 GHz. This high roughness makes the use of DMLS at 94 GHz unfeasible. The proposal in [92] takes advantage of gap waveguide for the distribution network. The design uses multiple layers using CNC milling for the manufacturing process, which is much more expensive than additive manufacturing.

The rest of the proposals in Table 4-III use Butler matrices. The design of most of them is based on SIW manufactured using PCB technology. The presence of the dielectric substrate introduces losses greater than 3 dB even in Ku band [110]. In [111], Butler matrix is based on

rectangular waveguide manufactured with CNC milling at 94 GHz. The distribution network is divided into two parts with a cut in E-plane to minimise leakage, but still requires very good alignment and electrical contact between parts. This increases the thickness of the distribution network for connection with WR-10 flanges on the sides and the weight of the parts. The proposal in [112] presents an atypical Ka-band SISL-based design with five stacked layers of dielectric substrate. Losses are relatively low, but not sufficient for W-band. Two proposals use microstrip technology at 60 GHz with manufacturing based on LTCC [113] and LCP [114]. Both cases have very high insertion losses and use amplifiers to compensate them.

With respect to bandwidth, the proposed magic tee typically achieves relative bandwidths of 20% with an isolation of more than 30 dB. Butler matrix designs can reach bands above 2% and isolation above 20 dB. In general, the amplitude unbalance is between 0.2 dB and 1 dB and the phase unbalance between 2° and 10°. However, many of these results are given only in simulation since the measured results are from a prototype with the comparator network or beamforming integrated with an antenna.

The prototype presented in this work takes advantage of the high precision of current SLA 3D printing techniques with copper plating. The result is very smooth surfaces with a metallization roughness around 0.3 μm . This technology applied to gap waveguide allows to drastically reduce the manufacturing costs, to decrease the complexity of the assembly and to obtain the lowest insertion losses of the Table 4-III. The VSWR of 1.2 and isolation better than 23 dB improves the values obtained from some proposals. The amplitude and phase unbalance is better than ± 1.5 dB and $\pm 20^\circ$ respectively in a relative band of 0.7%. These values, worse than those obtained in other proposals, are mainly due to the asymmetry of the Butler matrix. This design consideration was chosen considering that the working band around 94 GHz is very narrow (± 200 MHz, 0.4%) for a monopulse radar, so the values obtained are good enough for the operation of the system. In Table 4-III, we include the values at 94.17 GHz that are comparable to the rest of the proposals. This minimum deviation in frequency allows to validate the viability of the SLA technology for prototyping at W band.

TABLE 4-III: COMPARISON WITH PREVIOUS WORKS

Ref.	Feeding network ¹				Physical features		Scattering Parameters ²				Network performance ³		
	Compare/ Beamform	Pieces	Type	Manuf.	Size (mm)	Size (λ_0)	f_0 (GHz)	BW (%)	VSWR	Iso. (dB)	AI (dB)	PI (°)	IL (dB)
[107]	MT	19	RW	DB	51x51x4	13.3x13.3x1	78.5	21.9	2	n. a.	0.2	n. a.	n. a.
[108]	MT	1	RW	DMLS	240x240x31	12x12x1.55	15.1	12.9	1.5*	30	0.2*	1.5*	0.15*
[109]	MT	11	RW	SM	40x37.4x2.65	12x11.3x0.8	90.5	15	2	45*	n. a.	n. a.	0.9*
[92]	MT	3	GW	CNC	55x55x8	17.4x17.4x2.5	95	21	2	50	0.5	2	4.4
[90]	BM	1	SIW	PCB	28x30x0.5	8.8x9.4x0.157	94	1.7	2	n. a.	0.2*	5.2*	1.9*
[89]	BM	1	SIW	PCB	130x125x0.5	41x39.4x0.16	94.5	1.6	1.4	n. a.	2.66*	n. a.	7
[111]	BM	2	RW	CNC	58x58x>20	18.2x18.2x>6.3	94	4.3	1.4	21	0.2	n. a.	1
[115]	BM	1	SIW	PCB	93x100x0.5	10.9x11.7x0.06	35	2.9	1.7	25	0.5	n. a.	3
[110]	BM	1	SIW	PCB	140x154x1	8.2x9x0.06	17.53	6	2	25	0.5	5.5	5
[106]	BM	1	SIW	PCB	27x18x0.25	5.4x3.6x0.05	60	6.7	1.5	13.5	2	12	2.5
[112]	BM	5	SISL	PCB	35x34x2.65	3x2.9x0.23	25.5	7.8	1.7	16	1	8	1.1
[113]	BM	1	MS	LTCC	4.2x2.5x1	0.88x0.5x0.2	62.5	1.6	2	n. a.	n. a.	10	7
[114]	BM	1	MS	LCP	58x58x0.05	11.6x11.6x0.01	60	3.3*	2*	17*	1*	10*	2.5*
Our work	BM	1	GW	SLA	160x100x3.9	50.1x31.3x1.2	94	0.7	1.2	23	1.5	20	0.8
							94.17	-	1.12	33	0.44	8.5	

¹ Feeding network: MT: Magic Tee, BM: Butler Matrix. Transmission line: RW: Rectangular Waveguide, GW: Gap Waveguide, SIW: Substrate Integrated Waveguide, SISL: Substrate Integrate Suspended Line, MS: Microstrip. Manufacturing technology: DB: Diffusion Bonding, DMLS: Direct Metal Laser Sintering (3D-Print), SM: Silicon Micromachining, CNCM: Computer Numerical Controlled Milling, PCB: Printed Board Circuit, LTCC: Low Temperature Cofired Ceramics, LCP: Liquid Crystal Polymer, SLA: Stereolithography (3D-Print).

² Scattering Parameters: f_0 : central frequency, BW: relative bandwidth, Iso.: Isolation between input ports.

³ Network performance: AI: Amplitude imbalance, PI: Phase imbalance, IL: Insertion loss in the feeding network.

* Simulated value.

4.3.3 Design Considerations

This last section is dedicated to the analysis of the problems encountered in the last stage of prototyping, simulation and measurement. All the problems are related to the resonances found in Figure 4-33, Figure 4-35 and Figure 4-36. These resonances appear because higher order modes are being excited in the structure. In the case of the crossover in Figure 4-33, the TE_{10} and TE_{20} modes must be excited, but the dimensions allow the TE_{30} mode to be excited as well. The resonances found in Figure 4-35 and Figure 4-36 for the complete Butler matrix are mainly due to an excess of width in the straight sections $S4$ and $S5$ (see Figure 4-19) allowing TE_{20} mode propagation. Table 4-IV shows the Wg widths of the basic waveguide gap structure (Figure 4-3(a)) for which modes TE_{10} , TE_{20} and TE_{30} begin to propagate at 94 GHz. As can be seen, the width of the crossover is above 5.1 mm and the width of sections $S4$ and $S5$ is above 3.5 mm. Therefore, the resonances found must be eliminated by correcting these dimensions. We propose to use the values $n=2$ and $n=4$ instead of the current $n=3$ and $n=7$ for the hybrid coupler and crossover design. In this way, their dimensions are reduced but they continue allowing a good matching in band. For the phase adjustment in sections $S4$ and $S5$, we propose the substitution of the bends of section W (Eq. (4-10)) by 90° phase shifters to have to correct -47.7° instead of 132.3° . With this change, the width Wg of sections $S4$ and $S5$ should be 3.15 mm instead of 3.6 mm for phase adjustment.

TABLE 4-IV: WIDTH MODIFICATIONS

Mode	Wg for $f_c=94$ GHz	Component modified	Previous width	New width
TE_{10}	1.9 mm	Hybrid 3dB- 90°	5.6 mm	4.88 mm
TE_{20}	3.5 mm	Crossover	5.77 mm	4.68 mm
TE_{30}	5.1 mm	Sections $S4$ and $S5$	3.6 mm	3.15 mm

4.4 Design and Characterization of the Monopulse Antenna

This section presents the design, simulation and measurement of the gap waveguide coupling element to the RLSA antenna, as well as the full monopulse antenna prototype with the integration of the Butler matrix.

4.4.1 Design of the Antenna Coupled to a Circular Cavity

4.4.1.1 Design of the RLSA Antenna

A design of the RLSA antenna with T-shaped radiating elements, shown in Figure 4-38, arranged in concentric rings, has been proposed. In total, the antenna has 10 rings of radiating elements that have been optimized based on two algorithms: a global and a local algorithm that reduces the number of iterations [103]. In the optimization process, a quality factor is considered given by the uniformity of amplitude and phase and the radiated power value. The algorithm optimizes the difference between directivity and spill-over power for the sum pattern (power after the last slot ring). The analysis is performed using a fast Method of Moments algorithm with only 1 basic function per slot and analytical expressions for autoimpedance of slots and feed pins [116]. This algorithm allows relatively large antennas to be optimized in a reasonable time. The result of optimization is the length and radial position of the slots for each ring. The angle of the slots with respect to the radial wave front is 45° , so that the excitation of the two slots of the radiating element is identical.

Figure 4-38: Radiating element and parameters.

The RLSA antenna is implemented on a PTFE substrate Neltec® NY9217 with dielectric constant of 2.17 and a thickness of 0.762 mm. Since the rings of radiating elements must be separated from each other by a wavelength inside the dielectric, which corresponds to a phase of 2π required for uniform phase configuration, it is necessary to use a material with a dielectric constant greater than 1 to avoid grating lobes. It is recommended to use dielectric constant values greater than 1.7 so that the separation between consecutive concentric rings is less than $0.75 \cdot \lambda$ in free space.

The final distribution of the radiating elements in the 10-rings RLSA antenna is shown in Figure 4-39. The diameter of the antenna is around 45 mm. All the design parameters of the slots are depicted in Figure 4-40, which includes the distance from the centre of the pair of slots of each ring, their length and separation between them in millimetres and with respect to the wavelength in the dielectric and the rotation angle of each radiating element in each ring.

Figure 4-39: 10-ring RLSA antenna. The red circles in the centre of the scheme represent four probes excited with the corresponding phases and same amplitude.

Figure 4-40: Design parameters of the RLSA antenna: (a) distances of the slots to the centre of the antenna, (b) slot length and separation between slot pairs and rings, (c) slot length and separation between slot pairs and rings normalized with respect to the wavelength in the dielectric and (d) rotation angle of the slots in each ring.

In Figure 4-40(c), it can be seen that the length of the slots is around half a wavelength in the dielectric. This length starts in the first rings slightly below their resonant length to reduce the radiated power in the first stages. The length of the slots grows with each ring to achieve uniform illumination. In addition, the separation between the pair of slots of each radiating element is kept constant at a value of a quarter wavelength in the dielectric. This provides two advantages: the two slots are excited orthogonally, so circular polarization is achieved, and the waves reflected by each slot cancel in the centre of the antenna because they combine in counterphase, allowing a reduction of reflection. The RLSA antenna is designed for lossless exponential illumination, so that the real lossy substrate used with the antenna provides uniform illumination. The normalized power decay inside the lossless RLSA is shown in Figure 4-41(a). In Figure 4-41(b), the phase along a circumference centred on the origin of the RLSA results in a constant phase for the difference configuration and a progressive linear phase for the sum configuration. Figure 4-41(c,d) show the amplitude and phase error in a circumference centred within the RLSA for the sum and difference patterns, respectively. A maximum error of ± 2 dB in amplitude and $\pm 15^\circ$ in phase is observed for the sum diagram and an error of ± 1.5 dB and $\pm 5^\circ$ for the difference diagram.

Figure 4-41: Power and field values inside the RLSA: (a) power decay, (b) E-field phase in a circumference around the RLSA, (c) amplitude and phase errors for the sum pattern and (d) amplitude and phase errors for the difference pattern.

The sum and difference radiation patterns at 94 GHz obtained with the RLSA antenna in Figure 3-38 are shown in Figure 4-42. The radiation diagrams include the co-polar component (LHCP) and the cross-polar component (RHCP) for phi cuts of 0° , 45° , 90° and 135° . The SLL value for the sum diagram is -13.5 dB and -16 dB for the difference diagram. The relation CP-XP is 35 dB and 22 dB for the sum and difference patterns, respectively.

Figure 4-42: Radiation diagrams at 94 GHz: (a) CP and XP sum patterns and (b) CP and XP difference patterns.

The directivity as a function of the frequency for the direction of maximum radiation can be found in Figure 4-43. Since the radiating elements used are narrow-band in nature, the obtained aperture efficiency reaches values above 90% around 94 GHz, but falls rapidly to values of 30% at 90 GHz and 98 GHz.

Figure 4-43: (a) Directivity for the sum (Σ) and difference (Δ) patterns in the direction of maximum radiation. (b) Directivity, gain and efficiencies for the sum pattern.

4.4.1.2 Design of the Coupling Gap Waveguide Circular Cavity

Once the RLSA antenna and the gap waveguide structure have been defined, the dimensions of the circular cavity are selected for obtaining the resonance of TM_{0nm} and TM_{1nm} modes at 94 GHz. TM_{0nm} resonant modes have a radially symmetrical field distribution, so they are suitable for generating a uniform phase pattern inside the RLSA for Δ configuration. In the same way, TM_{1nm} modes can be excited with an arbitrary radial orientation so that we can induce the rotation of the fields inside the cavity. This rotary field distribution can be coupled to the RLSA and generate the Σ pattern. Figure 4-34 shows the design parameters of the circular cavity and a scheme of the fields inside the cavity. The calculation of the dimensions is made from the expression (4-11) for the resonance frequency of the TM modes of interest in a classical cylindrical cavity [4].

Figure 4-44: Circular cavity parameters. Representation of the TM_{011} and a rotating TM_{110} modes. E-fields in red and H-fields in dashed blue.

$$f_{nml} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{p_{nm}}{a}\right)^2 + \left(\frac{l\pi}{d}\right)^2} \quad (4-11)$$

The parameter p_{nm} represents the m -th root of the Bessel function of order n . The p_{nm} values listed in Table 4-V correspond to the first three roots of the Bessel functions of order 0 and 1. With these values, the resonance frequencies of the modes TM_{011} , TM_{021} , TM_{031} , TM_{110} , TM_{120} and TM_{130} can be calculated as a function of the radius a and height d of the circular cavity. The resonance variation is shown in Figure 4-45(a-c) as a function of d for different values of a . Figure 4-45(d) shows the cumulative 94 GHz frequency deviation for modes TM_{021} and TM_{120} . Since the resonance frequency of modes TM_{1m0} does not change when the height is changed, it is easy to find the radius and height values where both modes TM_{0m1} and TM_{1m0} resonate at 94 GHz. Applying the expressions (4-12) and (4-13), the dimensions of the hollow circular cavity are obtained to excite the desired modes. The resulting dimensions for the selected modes in the ideal circular cavity are in Table 4-VI.

TABLE 4-V: BESSEL ROOTS VALUES

Bessel root	Value	Bessel root	Value
p_{01}	2.405	p_{11}	3.832
p_{02}	5.520	p_{12}	7.016
p_{03}	8.654	p_{13}	10.173

Figure 4-45: Resonance frequencies in function of radius and height of the circular cavity: (a) values for TM_{110} and TM_{011} modes, (b) values for TM_{120} and TM_{021} modes, (c) values for TM_{130} and TM_{031} modes and (d) Cumulative frequency error for TM_{120} and TM_{021} modes respect to 94 GHz.

$$a = \frac{c \cdot p_{1m}}{2\pi \cdot f} \quad (4-12)$$

$$d = \frac{c \cdot l}{2\sqrt{f^2 - \left(\frac{c \cdot p_{0m}}{2\pi \cdot a}\right)^2}} \quad (4-13)$$

The expression (4-12) for calculating the radius a only depends on the resonance frequency f and the chosen mode TM_{1m0} (p_{1m}). The expression (4-13) for the height d uses as variables the previously calculated radius a , the resonance frequency f and the chosen mode TM_{0m1} (p_{0m} and $l=1$).

TABLE 4-VI: PARAMETER VALUES OF THE IDEAL CIRCULAR CAVITY

Desired modes	Radius	Height	Frequency
$\text{TM}_{011} + \text{TM}_{110}$	1.95 mm	2.05 mm	94 GHz
$\text{TM}_{021} + \text{TM}_{120}$	3.56 mm	2.58 mm	94 GHz
$\text{TM}_{031} + \text{TM}_{130}$	5.16 mm	3.03 mm	94 GHz

After preliminary calculations of the dimensions of an ideal circular cavity, we applied the values of radius and height obtained to a circular cavity based on gap waveguide. The dimensions of the pins are the same as those used in the design of the Butler matrix, with a diameter of 0.5 mm, height of 0.9 mm and gap of 0.5 mm with a separation between pins of 1.2 mm. The cylindrical wall of the cavity is replaced by four concentric rows of pins to ensure good field insulation inside the cavity. With this modification of the structure it is necessary to readjust the dimensions to generate the desired modes at 94 GHz. Starting from the ideal cavity, three different cavities have been designed for the three configurations of the Table 4-VI. For each configuration, the field of the TM_{0nm} mode and that of the TM_{1mn} mode with its corresponding $\text{TM}_{1mn|\pi/2}$ turned 90 degrees are shown. In the three cases, the values of the cavity radius and height have been adjusted to excite these modes at 94 GHz.

The structures in Figure 4-46 have the following dimensions: 1.88 mm radius and 2.08 mm height (Figure 4-46 (a)), 3.51 mm radius and 2.60 mm height (Figure 4-46 (b)) and 5.12 mm radius and 3.04 mm height (Figure 4-46 (c)). Of these three options, the first one is discarded due to its small dimensions, since the excitation by means of four GGW inputs makes it impossible to keep a circular structure. The third option, while allowing for a better access of the GGW inputs, has unwanted modes that excite at frequencies very close to the working frequency (93.5 GHz and 94.7 GHz). Therefore, the structure chosen is that corresponding to Figure 4-46 (b), which allows access through GGW and unwanted modes are excited at frequencies far enough apart (several gigahertz). The dimensions and resonance frequencies of the cavities in the Figure 4-46 are compared in the Table 4-VII.

TABLE 4-VII: PARAMETER VALUES OF THE GGW CIRCULAR CAVITY

Modes	Radius	Height	Previous mode	Next mode
$TM_{011} + TM_{110} + TM_{110 \pi/2}$	1.88 mm	2.08 mm	87 GHz	97 GHz
$TM_{021} + TM_{120} + TM_{120 \pi/2}$	3.51 mm	2.60 mm	91.3 GHz	100.7 GHz
$TM_{031} + TM_{130} + TM_{130 \pi/2}$	5.12 mm	3.04 mm	93.5 GHz	94.7 GHz

Figure 4-46: Amplitude of the electric field for the TM_{0nm} and TM_{1nm} modes excited at 94 GHz within three cavities of different sizes. (a) presents the electric field distributions for the smallest case studied. (b) shows the chosen structure with the desired modes. Finally, (c) depicts the electric field amplitude of the largest case studied. No further cases have been studied because numerous unwanted modes appear close to the working frequency.

In order to incorporate the four GGW inputs into the basic cavity structure, it is necessary to create four appropriately matched openings. In addition, the excited fields inside the cavity must be coupled as much as possible to the parallel plate waveguide (PPWG) of the RLSA antenna. For this purpose, a circular slot centred in the RLSA ground plane is used, which is in contact with the upper part of the cavity. The radius of this slot will determine the coupling of each of the modes of interest. An optimization process of the structure is carried out to reduce the reflection coefficient at GGW inputs and reduce amplitude and phase errors of field distributions generated inside the PPWG. To reduce computation time and facilitate the calculation of these errors, a PPWG without slots has been considered as a first approximation. As the radiating elements of the RLSA are arranged so that reflections generated by the slots are cancelled in the centre of the antenna, the effect on the reflection coefficient when adding these slots to the design should be negligible. The final structure with the four GGW inputs and the coupling slot to the PPWG is shown in Figure 4-47 and Figure 4-48.

Figure 4-47: Top view of the final gap waveguide circular cavity. Length values are in millimetres.

Figure 4-48: Lateral cross-section of the circular cavity coupled to the parallel plate dielectric substrate.

The coupling slot between the circular cavity and the PPWG in the final design is not circular, but wavy. This is due to the shape of the surface currents in the upper face of the circular cavity. Since the slot must simultaneously couple modes TM_{021} and TM_{120} , the radius of this slot must approximately coincide with the maximum of the surface current in the radial direction. Figure 4-49 depicts the amplitude of the current in the upper face of the ideal circular cavity for both modes, taking into account that the amplitude of the current for mode TM_{120} (right) also has a rotary pattern with sum mode excitation. The amplitude of these normalized currents in a radial cut are represented in the Figure 4-50. This graph includes the currents for both the ideal circular cavity and the final waveguide gap cavity of the Figure 4-47 and Figure 4-48. The maximums in the curves in Figure 4-50 are separated a distance of 1.9 mm and 3.6 mm for TM_{021} and TM_{120} modes. This difference prevents an efficient simultaneous coupling of both modes using a circular slot.

Figure 4-49: Representation of the surface current at the top face of the ideal circular cavity for the TM_{120} and TM_{021} modes at 94 GHz.

Figure 4-50: Radial normalized surface current for TM_{120} and TM_{021} modes at 94 GHz in the top face of the ideal cavity and in the final gap waveguide cavity in Figure 4-48.

Figure 4-51(a) shows the structure of the gap waveguide cavity with different shapes and sizes of coupling slot to the PPWG. The two circular shapes correspond to the slots with a diameter of 1.9 mm and 3.6 mm, coincident with the maximum currents of the Figure 4-50. A wavy slot is superimposed over them to find an efficient simultaneous coupling. The design parameters of this wavy slot are presented in detail in the Figure 4-51.

The combination of the S parameters of the cavity with four ports is shown in the Figure 4-52. The results for circular coupling slots of 1.9 mm and 3.6 mm diameter can be seen in the Figure 4-52(a), together with an additional result for an average diameter of 2.75 mm. As expected, matching at 94 GHz using a 1.9 mm slot is acceptable for the symmetrical mode (TM_{021}), but very bad for the rotary mode (TM_{120}), and vice versa when using the 3.6 mm slot. With the average diameter slot, a bad matching is achieved in both cases. However, when using the wavy slot in the Figure 4-51(b), it is possible to achieve a good match for both configurations simultaneously. The final results with this wavy slot are shown in the Figure 4-52(b), which also compares the reflection coefficient of the cavity coupled to the PPWG without RLSA and with the integrated RLSA. The only effect of introducing the RLSA is an improvement of the reflection level for the sum configuration (rotary mode).

Figure 4-51: Coupling slot from the circular cavity to the PPWG. (a) Representation of the circular slots for good matching of the TM_{120} mode (dashed curve), the TM_{021} mode (dotted curve) and both of them (solid curve). (b) Design parameters of the final wavy slot for simultaneous matching of the TM_{120} and TM_{021} modes at 94 GHz.

Figure 4-52: Matching of the rotary (TM_{120}) mode and symmetric (TM_{021}) mode for various shapes of the coupling slot. (a) Circular slots with various diameters and (b) final wavy slot.

It should be noted that the results obtained are inherently narrow-band due to the nature of the resonant cavity. In particular, during the process of optimization of the cavity and the coupling slot it has been intended to maintain a greater bandwidth for the sum mode at the cost of narrowing the band of the difference mode. The aim of this is to reduce the probability that the sum mode will have losses caused by poor matching due to a frequency deviation in manufacturing. A drop in gain in the main beam for this reason means a reduction in the system's performance in detecting targets of a certain size at a given distance.

Another important factor is the reduction of amplitude and phase errors of the fields inside the RLSA to achieve a radiation pattern with low secondary lobes for the Σ diagram and a pronounced null in the Δ diagram. These errors have been obtained by calculating the difference between the ideal and simulated electric field phase and amplitude values along a circumference with 6 mm radius in the central plane of the PPWG. It is of interest for the application to achieve better results in rotary distribution than in uniform distribution. This is because the first one is associated with Σ pattern, whose good performance is more critical than Δ pattern. The Σ pattern supports a higher level of radiated power so it is convenient to keep a low reflection level, as

indicated in the previous paragraph. In addition, the main power transmission is associated with the Σ pattern, so it should radiate as much power as possible and maintain a low secondary lobe level. In the case of the Δ pattern, it is only involved in the estimation of the direction of arrival of the detected signal. Therefore, higher weights have been applied to the optimization factors related to the Σ pattern.

In Figure 4-53, we show graphical representations of the field distributions both in a plane inside the cavity and in the central plane of the PPWG. On the one hand, in Figure 4-53(a1-3) we have the rotary field distribution in the cavity induced by 90-degree sequentially rotated excitation from the four GGW inputs (Figure 4-53 (a1)), together with the field coupled to the PPWG in amplitude (Figure 4-53 (a2)) and phase (Figure 4-53 (a3)). As can be seen, a rotary pattern with progressive linear phase change is obtained. On the other hand, in Figure 4-53 (b1-3) we present the radially uniform field distribution in the cavity excited by four GGW equiphased inputs (Figure 4-53 (b1)), which is coupled to the PPWG generating the field distribution of Figure 4-53 (b2) in amplitude and Figure 4-53 (b3) in phase. In this case, a radially uniform phase distribution is coupled inside the PPWG.

The amplitude and phase errors in a circumference with radius $R=6$ mm are represented in Figure 4-54. The maximum amplitude and phase errors achieved in the final design are as follows: for rotary distribution there is an amplitude error of 0.10 dB and a phase error of 1.0 degrees, while for homogeneous distribution the amplitude error is 0.14 dB and the phase error is 4.2 degrees. Amplitude and phase errors are summarized in Table 4-VIII. The results obtained are very good because the feeding structure of the RLSA is highly radially symmetrical.

Figure 4-53: Field representation of the rotary mode (a1-3) and the homogeneous mode (b1-3). (a1) and (b1) represent a snapshot of the amplitude of the electric field inside the cavity structure, where the rotation and phase shifting of the inputs can be appreciated in (a1) and the uniformity and equiphase feeding of the inputs in (b1). (a2) and (b2) show the electric field amplitude of the rotary and uniform PPWG coupled modes, respectively. Finally, (a3) and (b3) depict the phase of the electric field for the rotary and uniform PPWG coupled modes, respectively.

Figure 4-54: Amplitude (a) and phase (b) errors of the z -component of the E-field for the rotary and symmetric modes inside the PPWG for wave-fronts at a radius $R=6\text{mm}$.

TABLE 4- VIII: PHASE AND AMPLITUDE ERRORS FOR EZ IN THE PPWG FOR WAVE-FRONTS AT $R = 6 \text{ mm}$

Type of field	Maximum errors	
	Phase error (°)	Amplitude error (dB)
ROTARY	± 1.0	± 0.10
SYMMETRIC	± 4.2	± 0.14

The final PCB has the radiating elements of the antenna located on the external face (Figure 4-55(a)) arranged in concentric circular rings and a wavy coupling slot (Figure 4-55(b)) in the internal face. Both faces of the PCB must be precisely aligned.

Figure 4-55: 10-rings RLSA antenna (a) and wavy coupling slot (b).

4.4.2 Experimental Validation of the Antenna Coupled to a Circular Cavity

This section presents the assembly and measurements of the 3D printed circular cavity together with the PCB for the RLSA antenna. The results obtained are validated by comparison with simulations of the prototype in CST Microwave Studio. The manufactured prototype consists of two parts: the 3D printed plate with the circular cavity based on gap waveguide and the PCB with the RLSA antenna. A picture of the prototypes is shown in the Figure 4-56.

Figure 4-56: Manufactured circular cavity coupled to the RLSA antenna: (a) 3D-printed gap waveguide with top view RLSA and (b) Bottom view of the RLSA antenna with coupling slot.

A more detailed view of the coupling slot and the manufactured radiating elements is shown in the Figure 4-57. The coupling slot (Figure 4-57) presents the wavy pattern described in the design stage and a central hole of 0.35 mm in diameter to align the PCB with the 3D printed plate.

Figure 4-57: Micro-photographs of the PCB: (a) wavy coupling slot and (b) T-shaped radiating elements. The central hole in (a) has a diameter of 0.35 mm for alignment.

One of the critical parts of manufacturing is to get a good alignment between the two parts that form the antenna. The only alignment requirement that must be met is that the centre of the RLSA on the PCB matches the centre of the circular cavity of the 3D printed part. To achieve this, a pair of holes have been made in the PCB and the circular cavity to align with a 0.3mm diameter pin traversing the two pieces from side to side. For manufacturing reasons, the hole in the PCB has been defined with a diameter of 0.35 mm to maintain a clearance that allows the pin to slide. Similarly, the hole in the cavity has been defined with a diameter of 0.45 mm, since a smaller hole is likely to close completely during the 3D printing process and subsequent copper plating. Images of the cavity, the coaxial probe with the 0.3 mm pin and the alignment with the RLSA are shown in Figure 4-58.

Figure 4-58: Alignment considerations: (a) zoom of the circular cavity with a central alignment hole, (b) a coaxial connector with 0.3 mm diameter probe and (c) alignment with the RLSA antenna.

The prototype proposed for validation in this section does not generate the desired sum and difference patterns on its own, as it would be necessary to excite each of its four ports simultaneously with the corresponding phase pattern. In the process of measuring the S parameters, we have a two-port analyser, so the results have been acquired two by two. The final results for the sum and difference configuration have been postprocessed combining the measured S parameters applying the corresponding phase in each port. In the Figure 4-59, the combined measurements for the two modes of operation are represented together with the simulations. The similarity between simulation and measurement is acceptable, with a marked reduction in reflection for the difference mode on the centre frequency and a greater bandwidth for the sum mode. After verifying that there is a good reflection level for both modes on the centre frequency, we performed a full acquisition measurement in an anechoic chamber for each of the four ports. Figure 4-60 represents a photo of the mounting in an anechoic chamber. Similar to the S parameters, the radiation measurements of the four ports must also be combined by applying the corresponding phase to each of them to obtain the sum and difference radiation diagrams.

Figure 4-59: Measured and simulated S-parameters after the combination of 4 ports.

Figure 4-60: 4-port circular cavity with RLSA antenna in anechoic chamber for measurement.

To detect if there is any type of amplitude or phase unbalance in the measurement of the radiation for each port, we have first combined the directivity diagrams obtained for each port assuming that the antenna is perfectly symmetrical and the behaviour with the three remaining ports is identical. The comparison of the four patterns obtained in sum and difference mode is represented in the Figure 4-61, along with the simulated result. The cut for the represented results corresponds to an angle $\phi=45^\circ$. There is a clear variation of the maximum directivity obtained in each case, with a maximum variation of 2.9 dB for the sum diagram and 5.7 dB for the difference diagram. In addition, the maximum measured directivity of the sum diagram is 30.7, 1.8 dB below the simulated value. These results imply the existence of an amplitude unbalance that can have origin in a slight misalignment between RLSA and circular cavity.

Figure 4-61: Measured directivity (co-polar and cross-polar) supposing that the radiation produced by each input port is rotationally symmetric and matches the other three ports. Included the simulation for comparison. Sum patterns in (a) and difference patterns in (b).

With these data, the actual radiation patterns combining the measurements of the 4 ports are depicted in the Figure 4-62. The calculated diagrams are not rotationally symmetrical due to the amplitude unbalance found, although they maintain the desired monopulse behaviour. The

maximum directivity of the sum diagram reaches 29 dB with a null depth in the difference diagram of -15 dB. The simulated CST model, considering losses in both copper and dielectric, produces the radiation diagrams in Figure 4-63 with a maximum of 32.5 dB and a null of -44.5 dB. The difference lies in the unbalance of amplitude and phase present in the measurement and in additional phase errors within the RLSA produced by the tolerances in the dielectric constant of the substrate used. This would also explain the filling of nulls between the secondary lobes. Despite these problems, the maximum null to null ratio between the sum diagram and the difference diagram is 45 dB, with a null deviation of 0.6° from the broadside direction.

Figure 4-62: Measured directivity (co-polar and cross-polar) combining the radiation diagrams generated by the four ports. Sum patterns in (a) and difference patterns in (b).

Figure 4-63: Simulated directivity (co-polar and cross-polar) combining the radiation diagrams generated by the four ports. Sum patterns in (a) and difference patterns in (b).

Finally, the measured axial ratio is about 0.5 dB in the direction of the maximum radiation of the sum beam. The data for the four main cuts and those obtained in the simulations are shown in Figure 4-64. The circular polarization obtained is very good despite the unbalance of amplitudes found, mainly because the radiating elements generate this circular polarization independently. In simulation, the axial relation reaches practically 0 dB in broadside also due to the high rotational symmetry of the antenna.

Figure 4-64: Axial ratio in the four main cuts of the radiation diagram of the antenna in the Figure 4-60. Comparison of measurements and simulation.

4.5 Integration of the Complete Monopulse Antenna

This section deals with the combined operation of Butler matrix coupled to the RLSA through the circular cavity. The complete antenna consists of two pieces: the 3D-printed Butler matrix with the coupling cavity and the dielectric substrate with the RLSA in the top copper foil and the coupling slot in the bottom copper foil. The expanded view of the full monopulse antenna is illustrated in Figure 4-65.

Figure 4-65: Mounting scheme of the complete antenna.

The operation of the complete structure is shown in Figure 4-66. The final structure is presented in Figure 4-66(a), showing a side view of the complete device with the feeding part with Butler matrix on which the PTFE substrate is placed with the printed RLSA. An additional zoom view highlights the internal structure of the cavity in gap-waveguide, the coupling slot and the RLSA antenna. Figure 4-66(b) shows the behaviour of the electrical field distribution inside the Butler matrix when feeding the structure through the Σ input. The Butler matrix imposes conditions that result in the rotating field within the cavity. The Σ radiation pattern generated is shown on the right side of Figure 4-66(b). If the structure is fed through the Δ input, a field configuration is obtained in the Butler matrix that results in uniform mode excitation in the cavity, shown in Figure 4-66(c). Coupling this field distribution to the RLSA generates the difference radiation pattern that appears to the right of Figure 4-66(c).

Figure 4-66: Complete view of the antenna in (a) with the feeding network at the bottom and the RLSA at the top. Operation of the Butler matrix together with the obtained radiations diagrams for the Σ configuration (b) when feeding the structure through the upper input and for the Δ configuration (c) when feeding through the lower input.

It is worth noting the great radial symmetry of both the Σ and Δ radiation diagrams, achieved thanks to the reduced amplitude and phase errors in the coupling between the cavity and the RLSA. The side lobe level in the Σ pattern is small, around -17 dB relative to the main beam. Finally, the null obtained in broadside for the Δ pattern is very pronounced, with approximately

70 dB below the maximum radiation level of the Σ pattern. This big difference is difficult to find in a real prototype, but it is to be expected in simulation. The directivity obtained for the sum pattern is 32.5 dBi. However, since a lossy dielectric material is used to support the RLSA antenna, the gain will be lower. To estimate this effect, the simulation has been made considering the dielectric loss tangent $t_d = 0.008$. The value measured at 10 GHz is 0.0008 according to the datasheet, so we consider one more order of magnitude for the estimation. Copper with a roughness of $0.3 \mu\text{m}$ has also been used as a conductive material in the parallel plates that make up the RLSA. With these considerations, the gain is 31.4 dBi, 1.2 dB less than in the directivity. The losses in the distribution network for the same copper roughness is 0.8 dB, so the total loss is expected to be 2 dB. This means an antenna efficiency of about 80%, which is acceptable.

4.5.1 Results with SLA Technology and Copper Plating

The final prototype of the manufactured monopulse antenna is shown in the Figure 4-67. The dimensions of the assembled prototype are $150 \times 60 \times 4.7$ mm. A photo of the assembly for measurement of S parameters is included in Figure 4-68. Measurements are represented in Figure 4-69 together with the simulated S parameters of the complete structure. The bandwidth difference of the two patterns is clearly observed, with 0.8 GHz at -10 dB for the difference (Δ) pattern and 4 GHz for the sum (Σ) pattern. Both reflection coefficients fit well with the simulations. The measured and simulated isolation between ports is practically identical, with values below -15 dB between 93 and 95 GHz and a value around -25 dB at the central frequency. This validates additive manufacturing at frequencies close to 100 GHz with a high degree of fidelity to the ideal model.

Figure 4-67: Picture of the manufactured monopulse antenna before assembling.

Figure 4-68: Set-up for S-parameters measurement.

Figure 4-69: Simulated and measured S-parameters: (a) reflection coefficients and (b) isolations coefficients.

The measurements of the radiation patterns of the monopulse antenna were made in the spherical system of the anechoic chamber LEHA-UPM. Pictures of the complete system assembly are shown in the Figure 4-70. Figure 4-71 depicts the measured radiation patterns of the Σ (Figure 4-71 (a)) and Δ (Figure 4-71 (b)) configurations at 94 GHz for different azimuth angles, along with the cross-polar component.

Figure 4-70: Monopulse RLSA antenna in anechoic chamber for measurement.

For comparison, the results obtained from the simulations are represented in the Figure 4-72 for the two patterns and the same cuts in φ angle. In simulation, the high symmetry of the coupling cavity to the RLSA generates very stable radiation patterns in φ angle, since it results in very small phase and amplitude errors inside the RLSA. There is some small secondary lobe imbalance produced by small phase imbalances in Butler matrix outputs that excite the cavity. These phase imbalances are difficult to correct, as the external paths are longer than the internal ones, so the Butler matrix is not fully symmetrical. However, the measured radiation patterns show asymmetries due to amplitude and phase errors caused by manufacturing tolerances, assembly and variations of the dielectric substrate properties. The simulated sum diagram has a secondary lobe level for the sum beam better than -15 dB with a maximum directivity of 32.3 dBi, while in measure the directivity drops almost 3 dB to a value of 29.35 dBi. The nulls of the secondary lobes are filled and a marked asymmetry is observed in the cuts of the difference diagram, in which the null presents a value of -21.2 dBi and a deviation of 1.2° with respect to broadside. For a better perspective of the radiation measured on the sphere, 2D radiation diagrams

are represented in the Figure 4-73. With this representation, the produced asymmetries and the deviation of the null radiation of the difference diagram can better observed. The 2D measurements are represented in a range of angles from -50° to 50° in theta and from 0° to 360° in phi for the sum (Figure 4-73(a)) and difference (Figure 4-73(b)) diagrams. For a clearer view of the null, a zoom of the difference diagram is included for a range in theta of -5° to 5° in the Figure 4-73(c). The 2D measurements also correspond to the results for 94 GHz.

Figure 4-71: Measured radiation at 94.5 GHz: (a) sum pattern and (b) difference pattern.

Figure 4-72: Simulated Δ (a) and Σ (b) patterns (Directivity-LHCP) for different azimuth angles.

Figure 4-73: 2D-plots of the measured radiation diagrams: (a) sum pattern, (b) difference pattern and (c) zoom of the difference pattern for better visualization of the deep null.

The measurements of directivity and gain variation of the sum beam with frequency are shown in the Figure 4-74. Figure 4-74(b) includes the simulated directivity considering cumulative errors and deviations. Considering an error of $30 \mu\text{m}$ RMS in the slots length (value provided by our manufacturer), a deviation of the dielectric constant of the substrate from 2.17 to 2.318 and a misalignment between Butler matrix and RLSA of 0.15 mm (as explained in Figure 4-76), the simulated directivity is very close to the measured one. A maximum gain of 28.86 dBi is observed at 93.4 GHz with an approximately constant decrease to 95 GHz. The gain value at 94.5 GHz is 27.8 dBi. Fig. 27(a) compares the final measurements and simulated directivity, gain and radiation efficiency, which is about 70% between 93 and 95 GHz in simulation and measurement. A comparison of the axial ratio of the measured and simulated sum beam as a function of frequency can be seen in the Figure 4-75. The measured results show a very good circular polarization, with an axial ratio below 0.5 dB between 93.5 GHz and 95.5 GHz. The minimum of 0.26 dB is given at 94.08 GHz, very close to the simulated values around 0.2 and 0.3 dB.

Figure 4-74: Measured (a) gain between 90 and 98 GHz and (b) directivity, gain and radiation efficiency between 93 and 95 GHz.

Figure 4-75: (a) Comparison between measurements and final simulated directivity, gain and radiation efficiency. (b) Measured and simulated axial ratio vs frequency.

To finalize the analysis of the results obtained, a study of tolerances in the alignment between the two parts of the antenna has been carried out. The necessary misalignment of 0.15 mm has been sought to obtain the measured deviation of 1.5° of the null in the difference diagram. The Figure 4-76 shows the superposition of the circular cavity, the coupling slot and the radiating elements with an ideal alignment and a displacement of the PCB with the antenna of 0.15 mm in y-direction.

Figure 4-76: Superposition of ideal alignment and 0.15 mm displacement in y direction between circular cavity, coupling slot and RLSA antenna.

Figure 4-77: 2D radiation diagrams for ideal and adjusted simulations: sum beams for ideal simulation (a) and simulation with error in slots length, dielectric constant deviation and 0.15 mm misalignment (c). Difference beams for ideal simulation (b) and simulation with error in slots length, dielectric constant deviation and 0.15 mm misalignment (d).

For the ideal case, the sum diagram and the difference diagram are well centred, with a high rotational symmetry. In the sum diagram the secondary lobes appear well defined with deep nulls between them. When applying the 0.15 mm mismatch in Figure 4-77(c,d), together with the other errors and deviations analysed in Figure 4-74(b), the symmetry is broken and the secondary lobes are blurred in the same way as in the 2D representation of the measurements in the Figure 4-73. In addition, the null of radiation of the difference diagram has deviated to 1.5° . A clear unbalance of the level of the conical beam of the difference diagram is observed, with a maximum of 28.5 dB and a minimum of 22.5 dB in the cut $\varphi=75^\circ$.

A comparison with other works is conducted in Table 4-IX. The elements included in the comparison are the type of comparator and antenna used, the number of pieces, parts or layers that compose the antenna, the type of transmission line, the manufacturing technology used, the size, characterization of S-parameters and antenna performance. The proposals with magic tee present acceptable performance in all aspects, but this type of structures typically require a multi-level distribution network with several layers. Butler matrices can be designed on a single level, reducing manufacturing complexity, alignment errors and cost. In addition, the proposals with magic tee [92][107][109] are made of metal with expensive manufacturing processes. The proposal in [92] takes advantage of gap waveguide for the distribution network. The design uses multiple layers using CNC milling for the manufacturing process, which is much more expensive than additive manufacturing. The rest of the proposals in Table 4-IX uses Butler matrices. The design of most of them is based on SIW manufactured using PCB technology [89][90]. The presence of the dielectric substrate results in antenna efficiency below 20%. In [111], Butler matrix is based on rectangular waveguide manufactured with CNC milling at 94 GHz. The feeding network illuminates a Cassegrain reflector system that must be very well aligned. With respect to bandwidth, the magic tee designs typically achieve relative bandwidths of 20% with an isolation of more than 30 dB. Butler matrix designs can reach bands above 2% and isolation above 20 dB. Only full metal proposals without reflector components presents acceptable antenna efficiency above 60%.

The prototype presented in this work takes advantage of the high precision of current SLA 3D printing techniques with copper plating. The 0.83% bandwidth is limited by the difference pattern due to the inherently narrowband circular cavity used. This design consideration was chosen considering that the working band around 94 GHz is very narrow (± 200 MHz, 0.4%) for a monopulse radar, so the values obtained are good enough for the operation of the system. The isolation is better than 20 dB and the antenna efficiency is 70%, better than most of the works. The maximum-null relation is about 50 dB and an axial ratio better than 0.5 dB LHCP. These results highlight over the other works in the comparison.

TABLE 4-IX: COMPARISON WITH PREVIOUS WORKS

Ref.	Feeding network ¹			Physical features		Scattering Parameters ²				Antenna performance ³					
	Comparator /Antenna	Type	Manuf.	Size (mm)	Number Pieces	f_0 (GHz)	BW (%)	VSWR	Iso. (dB)	G_Σ (dBi)	XP_Σ (dBi)	ND_Δ (dB)	SLL_Σ (dB)	Pol./AR (dB)	η (%)
[89]	BM/Array	SIW	PCB	130x125x0.5	1	94.5	1.6	1.4	n. a.	25.8	n. a.	45.8	-10	Linear	15.2
[90]	BM/Reflector	SIW	PCB/CNC	140x140x55	2	94	1.7	2	n. a.	35.6	-21.3	25	-13	RHCP/1.5	19.1 ^a
[92]	MT/Array	GW	CNC	55x55x9	4	95	21	2	50	30.5	-27	40	-13	Linear	60
[107]	MT/Array	RW	DB	50.9x50.9x5.8	29	78.5	21.9	2	n. a.	32.6	n. a.	53	-13	Linear	83
[109]	MT/2x2-SHA	RW	SM	40x37.4x2.65	6	90.5	15	2	45*	14.2	n. a.	49	-10	Linear	69
[111]	BM/Cassegrain	RW	CNC	135x135x46.7	5	94	4.3	1.4	21	35.1	-23	40	-17	RHCP/1.2	18.3 ^a
Our work	BM/RLSA	GW	3D-SLA	150x60x4.7	2	94	0.83	2	20	27.8	-30.8	50	-14	LHCP/0.5	70

¹ Feeding network: MT: Magic Tee, BM: Butler Matrix, SHA: Stepped Horn Antenna, RLSA: Radial Line Slot Array. Transmission line: RW: Rectangular Waveguide, GW: Gap Waveguide, SIW: Substrate Integrated Waveguide. Manufacturing technology: DB: Diffusion Bonding, SM: Silicon Micromachining, CNC: Computer Numerical Controlled Milling, PCB: Printed Board Circuit, 3D-SLA: Stereolithography (3D-Printing).

² Scattering Parameters: f_0 : central frequency, BW: relative bandwidth, Iso.: Isolation between input ports.

³ Antenna performance: G_Σ : Gain of the SUM beam, XP_Σ : Cross-polar level of the SUM beam relative to co-polar, ND_Δ : Null depth of the DIFF beam relative to the SUM beam, SLL_Σ : Side Lobe Level of the SUM beam, Pol./AR: Polarization and axial ratio if circular polarized, η : radiation efficiency.

* Simulated value, ^a Total efficiency (factor of aperture and radiation efficiencies)

4.5.2 Results with DMLS Technology

To test another additive manufacturing technology, the Butler network with coupling cavity has been remade using Direct Metal Laser Sintering (DMLS) technology. The same PCB with the RLSA has been used as for the prototype measured with the part manufactured with SLA. This comparison between technologies is interesting to validate their suitability in W-band antennas. DMLS technology consists of applying heat to metal particles by means of a laser beam so that they are sintered together. This is done layer by layer with a thickness of about 25 μm to form the complete piece. The metal particles are composed of the AlSi10Mg alloy which provides good casting properties, high hardness and strength and light weight. Images of the manufactured prototype are shown in Figure 4-78. A polishing process has been carried out on the external face of the prototype to improve the contact of the WR-10 ports with the connectors. The inner face has not been polished due to the complexity of the design and risk of breaking some of the pins. Figure 4-79 shows a detailed comparison of the inner face Figure 4-79(a) and outer face Figure 4-79(b) around the alignment hole with the RLSA in the centre of the circular cavity. This hole has a diameter of 0.45 mm. On the inner side, the stacked layers with high surface roughness and the orientation of the printing angle are clearly visible.

Figure 4-78: Pictures of the manufactured Butler matrix with DMLS. (a) Complete Butler matrix, (b) detail of the circular cavity and (c) detail of the transition to WR-10.

Figure 4-79: Microphotographs of the internal (a) and external (b) surfaces of the printed part using DMLS. The images are taken in the area of the alignment hole in the circular cavity, with a diameter of 0.45 mm, as a size reference. The external surface is polished to eliminate the ripples produced by the layers to improve the connection of the WR-10 flanges.

Figure 4-80 illustrates a comparison of the WR-10 ports manufactured with SLA (Figure 4-80(a)), DMLS (Figure 4-80(b)) and commercial flange with CNC machining (Figure 4-80(c)). There are large variations in the port shape of the part manufactured with DMLS. The SLA prototype has a very flat port walls, even better than the CNC machined part. The latter has a slight convexity of the long edges of the waveguide. Another important point in the comparison is the weight of the complete prototype, using SLA or DMLS. Figure 4-81 illustrates the measurement of the weight of the two prototypes on a precision digital scale. The values of 93 grams for the prototype with DMLS (Figure 4-81(a)) and 55 grams for the prototype with SLA (Figure 4-81(b)) have been obtained. This large difference is due to the lower density of the resin

used in SLA compared to the metal powder in DMLS. This reduced weight is an advantage in favour of SLA manufacturing.

Figure 4-80: Comparison of the WR-10 port manufactured with three different technologies. (a) SLA, (b) DMLS and (c) commercial CNC machined flange.

Figure 4-81: Weight comparison of the complete RLSA antenna. (a) With DMSL and (b) with SLA.

The measured S-parameters and radiation diagrams of the prototype with DMSL are shown in Figure 4-82. A homogeneous behaviour over the whole band is observed for the reflection coefficient of the sum and difference port with values ranging from -10 to -5 dB in Figure 4-82(a). The resonant bands associated with the modes in the circular cavity are not appreciated. The absence of the resonant bands is a sign that these modes are not being adequately excited in the cavity, probably because the phase differences in the cavity have changed considerably due to the roughness of the paths in the Butler matrix. The reflection below -5 dB in the whole band seems to indicate that there the roughness and waviness of the stacked layers have a high impact in the matching of the antenna. Furthermore, the isolation between ports is below -20 dB from 92 GHz to 97 GHz, but in simulation the level is below -20 dB only in a narrow band centred at 94 GHz. This apparent improvement is due to the failure to excite the appropriate modes in the cavity and the high losses introduced by the roughness of the material.

Figure 4-82: Measurements of the monopulse antenna with the Butler matrix manufactured with DMLS. (a) S-parameters, (b) radiation diagram exciting the sum port and (c) radiation diagram exciting the difference port. Co-polarization (CP) and cross-polarization (XP).

The radiation diagrams for the sum and difference ports are shown in Figure 4-82(b) and Figure 4-82(c), respectively. Very poor behaviour is observed, with a very undefined beam for the sum port and a non-existent null for the difference port. The measurement of the radiation diagrams confirms that the desired modes are not correctly excited in the circular cavity. Therefore, it is concluded that the high irregularities present in the entire Butler matrix, as well as the roughness of the material due to the agglomeration of metallic particles, prevent the antenna from achieving a good matching, low losses and the required phase conditions in the circular cavity to provide the monopulse function at 94 GHz.

4.6 Conclusion

The W band monopulse antenna designed in this chapter is based on a Radial Line Slot Array (RLSA) antenna fed by a circular cavity connected to a two-input Butler matrix to simultaneously generate the sum and difference patterns. The prototype design is oriented to manufacturing with PCB technology for the RLSA and metallized 3D-printing technology for the feeding structure. This chapter validates the additive manufacturing applied to waveguide gap in W band. The application of additive manufacturing, in particular stereolithography (SLA), on

a Butler matrix has shown an excellent performance with a high degree of similarity between the simulations and the measurements made. The manufacturing and measurement of multiple components corroborates that the results are repeatable. The copper metallization applied on the resin-printed parts offers excellent conductivity and roughness properties, achieving a very low level of losses of approximately 0.05 dB/cm equivalent to a copper surface roughness of 0.3 μm RMS. The great versatility of additive manufacturing allows the construction of parts and shapes that are impossible or very expensive to manufacture with traditional technologies such as CNC. This adds degrees of freedom that facilitate the design stage. In addition, the high degree of complexity of the prototypes presented in this chapter would have meant a great cost of manufacturing time with CNC and the use of tools such as very thin and fragile drills. 3D-printing has resulted in a significant reduction in the cost of prototyping. The final result is a 3D-printed Butler matrix based on gap waveguide with two inputs and four outputs to generate two different phase configurations: 90° progressive rotation and equiphase. The integration of this Butler matrix with a Radial Line Slot Array (RLSA) monopulse antenna was carried out successfully. The final prototype simultaneously generates a sum and difference diagram with the two phase configurations mentioned. The measurements of the Butler matrix show an amplitude unbalance at outputs of ± 1.5 dB and a maximum phase error of $\pm 20^\circ$. These deviations are acceptable for the correct operation of the monopulse antenna. The results of the complete antenna prototype show a good behaviour of the sum and difference patterns with a measured directivity of almost 30 dBi and a radiation efficiency of 70% for the sum beam and a very pronounced null at 1.5° from broadside for the difference beam. The achieved circular polarization is also very low about 0.5 dB in a 2 GHz band. An additional measurement of a Butler matrix manufactured with DMLS shows that the irregularities and roughness of the sintered metal particles result in bad matching, high losses, and high deviation of phase conditions in the circular cavity. This leads to a non-functional monopulse antenna.

Chapter V

Electrically Tuneable Phase Shifter Based on Liquid Crystal and 3D-Printed Gap Waveguide at W Band

5.1 Introduction

The development of millimetre-wave band antennas requires beam reorientation capability to address the needs of new generation communications. To minimize mechanical errors, moveable component failures and increase beam steering speed, the main objective is to get this reconfiguration controlled electrically instead of mechanically by applying certain control voltages. As already introduced in section 1.6, the use of new materials with anisotropic properties dependent on the direction and intensity of the electric field that passes through them is acquiring special importance.

A well-known material that has these properties is liquid crystal (LC). This material is known for its worldwide use in liquid crystal displays in the form of tension-controlled liquid crystal filled cells. Its operation consists of the ability to polarize the visible light that passes through this material depending on the orientation of its molecules. The orientation angle can be controlled by applying a quasi-static electric field, although it can also be controlled with temperature changes depending on the liquid crystal. This change of properties (anisotropy) by the change of direction of these molecules can be applied to much lower frequencies (microwave range and millimetric waves) to modify the dielectric constant of the medium in a certain direction. These promising properties are of great interest for the development of future millimetre-wave systems [117]-[118]. Some work is already done to exploit this characteristic of liquid crystal to design microwave filters [119], phase shifters [120]-[126], polarizers [127] and antennas [39],[128]-[129].

One of the pioneering companies in the development of liquid crystals with improved properties for use in RF is Liquistal® by © Merck KGaA. The aim of this company is to develop new liquid crystal composites that offer high dielectric anisotropy and small loss tangent at microwave and millimetre-wave ranges [40]. Typically, the molecules that compose the liquid crystal have a rod shape with an asymmetric distribution of electric charge, so they behave like a dipole. These dipole molecules tend to align in the direction of the electric field. In the nematic phase of the liquid crystal, the molecules are randomly arranged but with an axis orientation in a preferred direction. This direction can be controlled varying the intensity of the electric field applied. Two different dielectric constants can be defined: a dielectric constant $\epsilon_{\perp}$ (perpendicular to the electric field in a given direction) and a dielectric constant $\epsilon_{//}$ (parallel to the electric field in a given direction). Figure 5-1 represents the transition between the 0-volt non-polarized state (a) and the 30 volts saturated state with maximum polarization (c). It should be noted that the molecules closest to the top and bottom edges of the liquid crystal cell maintain a certain angle,

called the pretilt angle, in all states due to their interaction with the boundary conditions. The calculation of the angle of the molecules at each position of the liquid crystal cell will be discussed in detail in section 5.2.

Figure 5-1: Orientation of the liquid crystal molecules depending on the voltage. Tilting of the molecules for 0 volts (a), 7 volts (b) and 30 volts (c).

The chapter is divided into the following sections. Section 5.2 details the theoretical analysis of liquid crystal, section 5.3 shows the design of the phase shifter and section 5.4 presents the comparison between measurements and final simulations. Finally, the conclusions are drawn in section 5.5.

5.2 Liquid Crystal Analysis

The anisotropic nature of liquid crystals requires the use of tensors to define how the dielectric constant varies in each of the directions of space [118]. A generic tensor that fully characterizes the dielectric constant of an anisotropic material is the two-rank tensor shown in expression (5-1). The components of this tensor depend on the mean value of the tilt angle of the liquid crystal molecules, which varies with the applied voltage. If the liquid crystal is confined between two parallel metal plates and a uniform quasi-static electric field is applied in the z -direction, the molecules can rotate freely in a vertical plane. This plane of rotation can be modified by applying a thin orienting layer on the parallel metal plates that allows the molecules to rotate in the preferred plane. The tensor (5-2) defines the dielectric constant of a liquid crystal contained between two parallel metal plates with a Y-Z rotation plane [130]-[132]. The parameters needed are the perpendicular dielectric constant $\epsilon_{\perp}$, the parallel dielectric constant $\epsilon_{\parallel}$, and the mean tilt angle θ_m of the molecules in the liquid crystal. The dielectric anisotropy is defined as $\Delta\epsilon = \epsilon_{\parallel} - \epsilon_{\perp}$.

$$\bar{\epsilon} = \begin{pmatrix} \epsilon_{xx} & \epsilon_{xy} & \epsilon_{xz} \\ \epsilon_{yx} & \epsilon_{yy} & \epsilon_{yz} \\ \epsilon_{zx} & \epsilon_{zy} & \epsilon_{zz} \end{pmatrix}; \quad \vec{D} = \bar{\epsilon} \cdot \vec{E} \quad (5-1)$$

$$\bar{\epsilon} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} + \Delta\epsilon \cos^2(\theta_m) & \Delta\epsilon \sin(\theta_m) \cos(\theta_m) \\ 0 & \Delta\epsilon \sin(\theta_m) \cos(\theta_m) & \epsilon_{\perp} + \Delta\epsilon \sin^2(\theta_m) \end{pmatrix} \quad (5-2)$$

The $\epsilon_{\perp}$ and $\epsilon_{\parallel}$ values are provided by the liquid crystal supplier, so only the mean angle θ_m has to be calculated as a function of the applied voltage. In a first approximation it is simple to derive the components of the tensor $\bar{\epsilon}$ assuming that $\theta_m=0^\circ$ for the state $V=0$ and $\theta_m=90^\circ$ for $V \uparrow$. The resulting tensors are (5-3) and (5-4), respectively. The resulting effect is the swapping of the components ϵ_{yy} and ϵ_{zz} due to the rotation of the liquid crystal molecules in the Y-Z plane.

$$\bar{\epsilon}_{V=0} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} + \Delta\epsilon & 0 \\ 0 & 0 & \epsilon_{\perp} \end{pmatrix} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\parallel} & 0 \\ 0 & 0 & \epsilon_{\perp} \end{pmatrix} \quad (5-3)$$

$$\bar{\epsilon}_{V \uparrow} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} & 0 \\ 0 & 0 & \epsilon_{\perp} + \Delta\epsilon \end{pmatrix} = \begin{pmatrix} \epsilon_{\perp} & 0 & 0 \\ 0 & \epsilon_{\perp} & 0 \\ 0 & 0 & \epsilon_{\parallel} \end{pmatrix} \quad (5-4)$$

The mathematical analysis for the derivation and the calculation of the expressions for the tilt angle of the liquid crystal molecules in each position of the Z-axis, $\theta(z)$, is the object of another thesis developed in parallel with the present work. In this section, to obtain $\theta(z)$ we only use a simple analytical expression [41] and an integral expression [133] that reduces the errors made by the first approximation. For the application of these expressions, it is first necessary to specify the operating conditions of the liquid crystal cell, thickness and other parameters required for the theoretical analysis. Figure 5-2 represents the behaviour of liquid crystal molecules confined between two parallel plates separated by a distance d . One of the plates acts as a ground plane and a voltage V is applied in the other plate. In Figure 5-2, the electric field applied is $\vec{E} = V/d\hat{z}$. The tilt angle $\theta(z)$ of the molecules acquires a θ_p value called the pretilt angle in the vicinity of the orientation layers ($z=0$ and $z=d$). Value of θ_p varies depending on the manufacturing process [134][135], but typically are between 0° and 10° [118]. The $\theta(z)$ angle increases progressively until it reaches $\theta(z=d/2)$. The inclination of the molecules is symmetrical with respect to the plane $z=d/2$.

Figure 5-2: Scheme of the liquid crystal cell.

The variation of the tilt angle of the molecules with the applied voltage is modelled with the differential equation (5-5) presented in [118]. The parameters k_{11} and k_{33} are the splay and bend elastic constants and $\alpha = \epsilon_0 \Delta \epsilon \cdot E^2$.

$$(k_{11} \cos^2 \theta + k_{33} \sin^2 \theta) \frac{d^2 \theta}{dz^2} + (k_{33} - k_{11}) \left(\frac{d\theta}{dz} \right)^2 \sin \theta \cos \theta + \alpha \sin \theta \cos \theta = 0 \quad (5-5)$$

The analytical solution of this differential equation proposed in [41] assumes the approximation $k_{11}=k_{33}=k_a$, which is close to actual values of common liquid crystals. The resulting differential equation is (5-6), with $A=\alpha/k_a$.

$$k_a \frac{d^2 \theta}{dz^2} = -\beta \sin \theta \cos \theta \rightarrow \frac{d^2 \theta}{dz^2} = -\frac{A}{2} \sin (2\theta) \quad (5-6)$$

The analytical solution for the expression (5-6) results in (5-7)

$$\theta(z) = 2 \tan^{-1} \left(e^{z\sqrt{A}+\theta'_p} \right) + 2 \tan^{-1} \left(e^{-(z-d)\sqrt{A}+\theta'_p} \right) + c \quad (5-7)$$

where $\theta'_p = \ln \tan \theta_p + \sec \theta_p$ and the c is a constant value given by (5-8)

$$c = \theta_p - 2 \tan^{-1} \left(e^{\theta'_p} \right) - 2 \tan^{-1} \left(e^{d\sqrt{A}+\theta'_p} \right) \quad (5-8)$$

If the approximation $k_{11}=k_{33}$ is not good enough a generalized integral formulation was described in [133]. The resulting expression (5-9) can be solved iteratively, where $k^+=k_{11}+k_{33}$. The solution is only valid for $0 \leq z \leq d/2$. However, thanks to the symmetry with respect $z=d/2$, the solution can be extended to $d/2 \leq z \leq d$ as a reflection of the values obtained for $0 \leq z \leq d/2$.

$$\theta(z) = \theta_p + \int_0^z \left[\left(\frac{\alpha}{k_{11} - k_{33}} \right) \cdot \ln \left(\frac{(k_{11} - k_{33}) \cos(2\theta(z)) + k^+}{(k_{11} - k_{33}) \cos(2\theta(d/2)) + k^+} \right) \right]^{1/2} dz \quad (5-9)$$

To calculate the mean tilt angle it is necessary to calculate the mean of all $\theta(z)$ angles obtained, which is represented in the expression (5-10).

$$\theta_m = \frac{1}{d} \int_0^d \theta(z) dz \quad (5-10)$$

The liquid crystal mixture used in this work is GT3-23002 from ©Merck KGaA. Figure 5-3 shows a photo of the bottle with the material. The properties provided by the manufacturer are enclosed in the Table 5-I.

TABLE 5-I: PROPERTIES OF THE MIXTURE GT3-23002

Property	Value @19 GHz, 20°C
$\epsilon_{//}$	3.0169
$\epsilon_{\perp}$	2.3616
$\tan\delta_{//}$	0.0035
$\tan\delta_{\perp}$	0.0128
k_{11}	20.9 pN
k_{33}	25.5 pN

Figure 5-3: Picture of the liquid crystal mixture GT3-23002 by Merk.

In the Table 5-I, values k_{11} and k_{33} are similar. The relationship $k_{11}/k_{33}=0.82$, so the analytical solution (5-7) for $k_{11}=k_{33}$ is a good approximation. To check if it is a good approximation, both methods have been applied to calculate the mean tilt angle. The variation of the tilt angle from $z=0$ to $z=d$ is depicted in Figure 5-4. Curves for analytic method (A) and integral method (I) are superposed. There is a large difference at low voltages (curves for 3V and 6V), but with higher voltages the difference is negligible. In Figure 5-5, this difference can be better appreciated by representing the mean angle as a function of the voltage applied. The difference in the calculated mean tilt angle can be considered negligible with an applied voltage higher than 9 volts.

Figure 5-4: Tilt angle vs. z for various values of applied voltage. Dashed curves correspond to the analytic solution (A) and solid curves correspond to integral solution (I).

Figure 5-5: Mean tilt angle vs. applied voltage using analytic and integral equations.

Finally, the values of the components of the tensor of dielectric constants are calculated by substituting the obtained mean tilt angle in the expression (5-2) for each applied voltage. The results are depicted in Figure 5-6. The value ϵ_{xx} keeps constant at $\epsilon_{\perp}$, but ϵ_{yy} and ϵ_{zz} exchange their values as the voltage increases. Three different stages are observed: the starting stage in which the epsilon values remain practically invariable between 0 V and 5V, the fast polarization stage between 5 V and 14 V in which there is a 75% variation of the dielectric anisotropy and the saturation stage above 14V. In this last stage, a tendency of the components of the dielectric constant to the values $\epsilon_{yy}=\epsilon_{\perp}$, $\epsilon_{zz}=\epsilon_{\parallel}$ and $\epsilon_{zy}=\epsilon_{yz}=0$ occurs when the mean angle tends to 90° . This pattern is best observed in the Figure 5-7 with a voltage variation up to 100 V.

 Figure 5-6: Variation of components of the dielectric constant tensor $\bar{\epsilon}$ with the applied voltage.

Figure 5-7: Saturation effect with voltage up to 100 V. (a) Tilt angle $\theta(z)$ is almost 90° between $z=10$ μm and $z=90$ μm. (b) The mean tilt angle θ_m tends to 90°.

5.3 Liquid Crystal Based Phase Shifter Design at W Band

The phase shifter is integrated in a 3D-printed groove gap waveguide structure similar to the one used in Chapter III. A stepped ridge is introduced to concentrate the electromagnetic field inside the liquid crystal phase shifter. The main details of the structure are shown in Figure 5-8. Figure 5-8(a) shows the electrode over the ridge to confine the fields. The electrode shape includes a low pass filter in order to decouple DC and RF. A pool made of a photopolymer (SU-8) contains the liquid crystal, as shown in Figure 5-8(b).

Figure 5-8: Scheme of the liquid crystal-based phase shifter embedded in groove gap waveguide. The metal electrode in (a) and the liquid crystal with photopolymer in (b).

The liquid crystal pool has a thickness of 100 μm . This pool is encapsulated between a 0.3 mm thick quartz glass, which includes the electrode made of gold, and a 1 mm thick quartz glass with the ground plane in the inner face (see a zoom of the structure in Figure 5-9). By applying a variable voltage to this electrode, it is possible to achieve different dielectric constants and thus produce a variable phase shifting. The working band of the final design covers from 85 GHz to 100 GHz. The dimensions of the transition from the groove gap waveguide input and output sections to the liquid crystal transmission line are illustrated in the Figure 5-10. The height of the steps tries to follow a pattern of Chebyshev, but, due to impedances in the input and output of the transition and physical and manufacturing limitations, some changes have had to be assumed to ensure their manufacturability. The E-shape of the last step is necessary to eliminate some resonant modes that appeared between that step and the quartz wafer. The use of the corrugations and central cuts in the electrode of the Figure 5-10(c) are detailed in the unit cell study explained later in this chapter.

Figure 5-9: Stack up of the phase shifter. It contains the liquid crystal cell in the photopolymer, all encapsulated between a ground plate and a thin glass with the electrode.

Figure 5-10: Dimensions of the stepped ridge in the interface between the GGW and the electrode. (a) Perspective view of the steps in the ridge, (b) cross-section of the transition from GGW to the liquid crystal line and (c) and liquid crystal cell.