Leakage Suppression for PCB-Waveguide Interconnects Using Hybrid Metasurfaces

DAVID GONZÁLEZ-GALLARDO
Leakage suppression for PCB-waveguide interconnects using hybrid metasurfaces

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Abstract

A metasurface is a two dimensional periodic structure which possesses macroscopic equivalent electromagnetic properties such as anisotropy, negative refractive index and Electromagnetic Band Gaps (EBG). EBGs have been used for the design of waveguide filters and gap waveguides. In this thesis, a transition between stripline and waveguide technology is designed for the standard 5G frequency band of 28 GHz. In addition, a metasurface will be designed and placed in the surroundings of this transition with the objective of suppressing the leakage that appears in the air gap between the Printed Circuit Board (PCB) and the waveguide flange.

The transition is embedded in the PCB technology while the metasurface is built in two different technologies at each layer: PCB and holey metal. There are non existing studies of this hybrid composition of a metasurface, therefore, a thorough study of this structure has been carried. The proposed technology simplifies the connection between elements, avoiding the leakage due to imperfect connection between layers.

Commercial software CST Microwave Studio Suite is used for simulations. A prototype has been built and measurements have been carried out in order to compare the simulation results with a real manufactured design.
**Abstract**


Övergången är implementerad i mönsterkortsteknik medan metaytan är en hybrid-lösning realiserad i två olika tekniker, i respektive transmissionssedningslager: mönsterkort och metall. Eftersom det saknas tidigare publicationer av denna typ av hybrid-lösning har en detaljerad studie genomförts. Den föreslagna lösningen förenklar anslutningen mellan mönsterkort och vågledare, och understrycker det läckage som skulle kunna uppträda i gapet mellan de två lagren.

För simuleringarna har CST Microwave Studio använts och en prototyp har tillverkats och mätts för att verifiera den framtagna övergången och lösningen för undertryckning av läckage.
Abstract

Una metasuperficie es una estructura periódica de dos dimensiones que posee propiedades electromagnéticas macroscópicas equivalentes como anisotropía, índice de refracción negativo y Bandas Electromagnéticas Prohibidas (EBG). Las estructuras EBG han sido utilizadas para el diseño de filtros en guías de onda y guías de onda GAP. En este trabajo de fin de master se diseña una transición de stripline a guía de onda para la banda de frecuencia estándar de 28 GHz usada en 5G. Además, una metasuperficie será diseñada y situada alrededor de la transición con el objetivo de eliminar las fugas que aparecen en el espacio entre la placa de circuito impreso (PCB) y la brida de la guía de onda.

La transición está integrada en tecnología PCB mientras que la metasuperficie se construye sobre dos medios: PCB y metal. No hay estudios previos sobre esta composición de metasuperficies híbridas, por tanto, se ha llevado a cabo un estudio exhaustivo de esta estructura. La tecnología propuesta simplifica la conexión entre elementos, evitando las fugas debidas a una conexión imperfecta de las distintas capas.

Se ha usado el software comercial CST Microwave Studio para las simulaciones. Se ha construido un prototipo y se han llevado a cabo medidas para comparar las simulaciones con el diseño real fabricado.
Abstract

Una metasuperfície és una estructura periòdica de dues dimensions que té propietats electromagnètiques macroscòpiques equivalents com l’anisotropia, l’ànxer negatiu i les Bandes Electromagnètiques Prohibides (EBG). Les estructures EBG han estat utilitzades per al disseny de fíltres en guies d’ona i guies d’ona GAP. En aquest treball de fi de màster es dissenya una transició de stripline a guia d’ona per a la banda de freqüència estàndard de 28 GHz utilitzada en 5G. A més, una metasuperfície serà dissenyada i situada al voltant de la transició amb l’objectiu d’eliminar les fugues que apareixen en l’espai entre la plaça de circuit imprès (PCB) i la brida de la guia d’ona.

La transició està integrada en la tecnologia PCB. Per contra, la metasuperfície es construeix sobre dos medis: PCB i metall. No hi ha estudis previs sobre aquesta composició de metasuperfícies híbrides, per tant s’ha dut a terme un estudi exhaustiu d’aquesta estructura. La tecnologia proposada simplifica la connexió entre elements, evitant les fugues degudes a una connexió imperfecta de les diferents capes.

S’ha utilitzat el software comercial CST Microwave Studio per a les simulacions. S’ha construït un prototip i s’han dut a terme mesures per a comparar les simulacions amb el disseny real fabricat.
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Contents

1 Glossary 13

2 Introduction 15

3 Theoretical Overview 17
  3.1 Waveguides .................................................... 17
  3.2 Planar Transmission Lines .................................... 19
  3.3 Transitions: State of the Art ................................. 20
    3.3.1 Radiating Element ........................................ 20
    3.3.2 Cavity Backshort ......................................... 21
    3.3.3 Probe Insertion .......................................... 22
  3.4 Metasurfaces .................................................. 23
    3.4.1 Eigenmode Analysis ...................................... 24
    3.4.2 Dispersion Diagram ...................................... 25
  3.5 Glide Symmetry ............................................... 29

4 Transition Design 32
  4.1 Integration in the PCB ........................................ 34
    4.1.1 Cavity filling ............................................ 34
    4.1.2 Cavity boundaries ....................................... 37
  4.2 Leakage Study ................................................ 43

5 Leakage Reduction Using an EBG 45
  5.1 Metasurface Topology ....................................... 45
  5.2 Unit Cell Design .............................................. 47

6 Complete Structure 59
  6.1 Single transition ............................................. 59
  6.2 Single transition with metasurface ......................... 62
  6.3 Comparison with already existing technologies .......... 66
  6.4 Back to back transition .................................... 69
  6.5 Back to back transition with metasurface ............... 72

7 Manufacturing, measurements and tolerance analysis 75
  7.1 Manufacturing ............................................... 75
  7.2 Measurements ................................................. 78
  7.3 Tolerance analysis ........................................... 84
    7.3.1 Rotational misalignment ................................ 84
    7.3.2 Lateral misalignment ................................... 85
    7.3.3 Dielectric thickness .................................... 87
List of Figures

1. Rectangular waveguide ........................................... 17
2. TE modes in a RWG: (a) TE_{10}, (b) TE_{20}, (c) TE_{30} ........ 18
3. (a) Microstrip line, (b) Stripline, (c) Coplanar line .......... 19
4. Example of a transition using a radiating (resonant) element . 20
5. Example of a transition using a cavity backshort .................. 21
6. Example of a probe insertion transition where a probe is directly inserted in the waveguide .......................... 22
7. Example of a basic metasurface generated by a periodic holey unit cell ......................................................... 23
8. Holey unit cell example. The dimensions are: \(a = 3.5 \text{ mm}, g = 0.05 \text{ mm}, h = 2 \text{ mm} \) and \(r = 1.4 \text{ mm}\). ......................................................... 26
9. Top view of several unit cell example .............................. 27
10. Example of a dispersion diagram .................................. 28
11. Example of glide symmetry in a simple structure ................. 29
12. Unit cell applying 2D glide symmetry ............................. 29
13. Dispersion diagram of the unit cell after applying 2D glide symmetry ......................................................... 30
14. (a) 2D glide symmetry, (b) 1D Glide symmetry .................. 31
15. First approach of the transition design: air filled cavity backshort ................................................................. 32
16. Designed Stripline ................................................... 33
17. \(S_{11}\) and \(S_{21}\) of the air filled cavity backshort ................. 33
18. Stack up of all the dielectric and metal layers of the PCB. Dielectric constant varies depending on the thickness of each layer. ........... 35
19. Cross-section (x-plane) of the dielectric-filled cavity ............. 35
20. Top view of the transition. Cavity lid has been hidden in order to allow the view of the inner part of the cavity ......................... 36
21. \(S_{11}\) and \(S_{21}\) of the dielectric-filled cavity ....................... 36
22. Cross-section of the transition. The grid shaped by the intersection of vias and metal layers is clearly shown .......................... 37
23. Top view of the cavity-backshort with via position. Dielectric layers and cavity lid have been hidden for helping in the view .............................. 38
Cavity-backshort design including pads for guarantying the contact between vias and metal layers.

Cavity backshort top view. No EM fields go out of the cavity due to the use of vias.

Top view of a metal layer. It can be seen the new shape of the cavity due to the addition of the pads.

Top view of the transition showing the position of vias. Distances are expressed to the center of the vias. Via diameter = 0.3 mm.

Model of the simulated transition.

$S_{11}$ and $S_{21}$ of the dielectric-filled cavity after replacing metallic walls by vias.

Some examples of studied cavity lids: elliptical lids (a) and (b), flat lid (c).

Cross-section of the transition showing the non-perfect connection between PCB and waveguide flange.

$S_{21}$ parameter of the transition for several air gap thicknesses.

Metasurface topology: (a) Rectangular and (b) Circular.

Holey unit cell on the left and its IBZ on the right.

Dispersion diagram of the holey unit cell. The dimensions are $r = 2.6 \text{ mm}$, $h = 1.5 \text{ mm}$, $a = 6 \text{ mm}$ and $g = 0.1 \text{ mm}$.

Glide-symmetric holey unit cell on the left and its IBZ on the right.

Dispersion diagram of the unit cell with glide-symmetric holes. The dimensions are $r = 2.6 \text{ mm}$, $h = 1.5 \text{ mm}$, $a = 6 \text{ mm}$ and $g = 0.1 \text{ mm}$.

Stackup view of the unit cell for upper and lower plates.

Different configurations for the metal layer of the upper plate. Vias are drawn as darker grey dots.

Holey unit cell with glide-symmetric air filled holes in both plates.

Dispersion diagram of the hybrid unit cell with glide-symmetric air filled holes in both plates. Dimensions are: $r_{\text{etch}} = r_{\text{cav}} = 2.8 \text{ mm}$, $r_{\text{wg}} = 2.8 \text{ mm}$, $h = 1.5 \text{ mm}$, $a = 7 \text{ mm}$ and $g = 0.1 \text{ mm}$.

Holey unit cell on the left and its IBZ on the right.
39b Dispersion diagram of the hybrid unit cell. Dimensions are: $r_{etch} = 2.5 \text{ mm}$, $r_{cav} = 3 \text{ mm}$, $r_{wg} = 2.8 \text{ mm}$, $h = 1.5 \text{ mm}$, $a = 7 \text{ mm}$ and $g = 0.1 \text{ mm}$.

40 IBZ of both unit cells: (a) 2D glide symmetry, (b) 1D glide symmetry.

41 Demonstration of why only the first region of the dispersion diagram is studied.

42a Hybrid unit cell with 1D glide symmetry on the left and its IBZ on the right.

42b Dispersion diagram of the hybrid unit cell with 1D glide symmetry. Dimensions are: $r_{etch} = 2.5 \text{ mm}$, $r_{cav} = 3 \text{ mm}$, $r_{wg} = 2.8 \text{ mm}$, $h = 1.5 \text{ mm}$, $a = 7 \text{ mm}$ and $g = 0.1 \text{ mm}$.

43 Model of the studied unit cell.

44 3D view of the studied unit cell.

45 $S_{21}$ parameter of the hybrid unit cell with 1D glide symmetry.

46 Cross-section of the complete structure.

47 Model of the simulated transition.

48 $S_{11}$ and $S_{21}$ parameters of the single transition.

49 Study of $S_{11}$ parameter of the single transition without metasurface for several air gap thicknesses.

50 Study of $S_{21}$ parameter of the single transition without metasurface for several air gap thicknesses.

51 (a) Single transition including metasurface, (b) Top view of the PCB plate and (c) Top view of the waveguide plate.

52 Top view of the transition including the metasurface around. The plane has been cut at the stripline layer and substrate layers have been hidden.

53 $S_{11}$ and $S_{21}$ parameters of the single transition with metasurface.

54 Study of $S_{11}$ parameter of the single transition with metasurface for several air gap thicknesses.

55 Study of $S_{21}$ parameter of the single transition with metasurface for several air gap thicknesses.

56 $S_{11}$ comparison when using or not the metasurface.

57 $S_{21}$ comparison when using or not the metasurface.

58 Illustration of a choke flange.
<table>
<thead>
<tr>
<th>Page</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>59</td>
<td>$S_{21}$ comparison between the choke flange and the designed meta-surface.</td>
</tr>
<tr>
<td>60</td>
<td>Illustration of the corrugations.</td>
</tr>
<tr>
<td>61</td>
<td>$S_{21}$ comparison between the corrugations and the designed meta-surface.</td>
</tr>
<tr>
<td>62</td>
<td>Back to back transition.</td>
</tr>
<tr>
<td>63</td>
<td>$S_{11}$ and $S_{21}$ of the back to back transition having no air gap.</td>
</tr>
<tr>
<td>64</td>
<td>Study of $S_{11}$ parameter of the back to back transition without meta-surface for several air gap thicknesses.</td>
</tr>
<tr>
<td>65</td>
<td>Study of $S_{21}$ parameter of the back to back transition without meta-surface for several air gap thicknesses.</td>
</tr>
<tr>
<td>66</td>
<td>Back to back transition including the metasurface on both.</td>
</tr>
<tr>
<td>67</td>
<td>$S_{11}$ and $S_{21}$ of the back to back transition including a metasurface around.</td>
</tr>
<tr>
<td>68</td>
<td>Study of $S_{11}$ parameter of the back to back transition with meta-surface for several air gap thicknesses.</td>
</tr>
<tr>
<td>69</td>
<td>Study of $S_{21}$ parameter of the back to back transition with meta-surface for several air gap thicknesses.</td>
</tr>
<tr>
<td>70</td>
<td>$S_{11}$ comparison when using or not the metasurface in the back to back transition.</td>
</tr>
<tr>
<td>71</td>
<td>$S_{21}$ comparison when using or not the metasurface in the back to back transition.</td>
</tr>
<tr>
<td>72</td>
<td>Top view of the metal layer containing the striplines.</td>
</tr>
<tr>
<td>73</td>
<td>Top view of the PCB.</td>
</tr>
<tr>
<td>74</td>
<td>Top view of the metal block.</td>
</tr>
<tr>
<td>75</td>
<td>$S_{11}$ parameter of several transitions with $g = 0$.</td>
</tr>
<tr>
<td>76</td>
<td>$S_{21}$ parameter of several transitions with $g = 0$.</td>
</tr>
<tr>
<td>77</td>
<td>30 $\mu$m thickness paper sheet placed on top of the metal block.</td>
</tr>
<tr>
<td>78</td>
<td>65 $\mu$m thickness metal tape placed around the screw holes.</td>
</tr>
<tr>
<td>79</td>
<td>$S_{21}$ parameter for several air gap thicknesses.</td>
</tr>
<tr>
<td>80</td>
<td>Comparison of the $S_{21}$ parameter for the back to back transition without metasurface and for several air gap thicknesses.</td>
</tr>
</tbody>
</table>
Comparison of the $S_{21}$ parameter for the back to back transition with metasurface and for several air gap thicknesses. .................. 81

a) H-plane crosstalk study and b) E-plane crosstalk study. ............... 82

$S_{21}$ parameter showing the H-plane crosstalk. .......................... 82

$S_{21}$ parameter showing the E-plane crosstalk. .......................... 83

Top view of the single transition with metasurface showing a rotational misalignment of 5°. .................................................. 84

$S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for several angular rotations. ........... 85

Top view of the single transition with metasurface showing a lateral misalignment along x-axis. ................................................. 86

$S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for several translations over x-axis. .... 86

$S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for several translations over y-axis. .... 87

$S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for a varying dielectric thickness. .... 88

$S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for a varying R-5785 dielectric constant. .......... 89

$S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for a varying R-5680 dielectric constant. .......... 89
1 Glossary

Abbreviations

AMC  Artificial Magnetic Conductor
EBG  Electromagnetic Band Gap
EM   Electromagnetic
IBZ  Irreducible Brillouin Zone
IoT  Internet of Things
SIW  Substrate Integrated Waveguide
MSL  Microstrip Line
PCB  Printed Circuit Board
PEC  Perfect Electric Conductor
PMC  Perfect Magnetic Conductor
RWG  Rectangular Waveguide
SL   Strip Line
TE   Transversal Electric Mode
TEM  Transversal Electromagnetic Mode
TM   Transversal Magnetic Mode
TRL  Thru, Reflect and Line
VNA  Vector Network Analyzer
WG   Waveguide
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$c$</td>
<td>Speed of light in vacuum</td>
</tr>
<tr>
<td>$\epsilon_r$</td>
<td>Relative permittivity</td>
</tr>
<tr>
<td>$\lambda$</td>
<td>Wavelength</td>
</tr>
<tr>
<td>$\lambda_g$</td>
<td>Guided wavelength of the waveguide</td>
</tr>
<tr>
<td>$\alpha$</td>
<td>Attenuation constant</td>
</tr>
<tr>
<td>$\beta$</td>
<td>Phase constant</td>
</tr>
<tr>
<td>$\gamma$</td>
<td>Propagation constant</td>
</tr>
<tr>
<td>$k_0$</td>
<td>Wave number in free space</td>
</tr>
<tr>
<td>$\mu_0$</td>
<td>Vacuum permeability</td>
</tr>
<tr>
<td>$\epsilon_0$</td>
<td>Vacuum permittivity</td>
</tr>
<tr>
<td>$\omega$</td>
<td>Angular frequency</td>
</tr>
<tr>
<td>$V_g$</td>
<td>Group velocity</td>
</tr>
<tr>
<td>$\tan \delta$</td>
<td>Loss tangent</td>
</tr>
<tr>
<td>$Z_0$</td>
<td>Characteristic impedance</td>
</tr>
</tbody>
</table>
2 Introduction

5G will be the fifth generation of cellular mobile communications whose main benefits will be an increase in data rate (20 Gbps) and capacity (1000x increase), reduction of latency (1 millisecond) and cost. 5G has additional targets such as energy saving and massive device connection. All these improvements will immediately open the possibility to new technologies such as IoT (Internet of Things), autonomous vehicles and smart cities. 5G will be released by ITU (International Telecommunication Union) in different phases, being the first one in March 2019 and the second one a year after, in March 2020 ([1] - [4]).

In order to support the new requirements of speed and capacity new frequency bands should be available, particularly moving to higher frequencies. This increase in frequency has a direct consequence on all the circuitry associated which will have a smaller physical size. Therefore new circuits, antennas, transmission lines and the rest of elements needed in the communication process must be designed. Another direct consequence of the increase of the frequencies is a reduction in the wall-penetration capability and the increase in the path loss.

Since different transmission technologies will be needed to deploy 5G, interconnection between them is a key point which must be properly addressed.

Hollow waveguides have been widely used as a standard hardware technology for the design of passive microwave components and antenna arrays. They are entirely made of metal and exhibit attractive features like low loss, good isolation and high power handling capability. However, as frequency increases, tolerance requirements become more strict with the consequent need of more precise manufacturing techniques which means that this solution becomes expensive and difficult to be mass-produced. Any possible gap in waveguide joints can cause strong field leakage with a resulting increase in the overall system loss. Furthermore, the integration of waveguide components and antennas with the active parts of the radio system is not straightforward due to the difference of interface, i.e. some parts are substrate-based and others are fully made of metal. Therefore, the design of low loss interconnects between the different interfaces and the development of suitable packaging solutions are key challenges from the integration point of view in order to build a radio system.

It has been previously proved that metasurfaces based on glide symmetry are suitable for the design of wideband and low loss lens antennas as well as providing broad scanning performance in azimuth ([5], [6]). Moreover, periodic patterns built with glide-symmetric unit cells can be used as an EBG structure to suppress
any possible leakage in the presence of a gap between two metallic plates ([7], [9], [10]). We here propose to use this property to investigate several options of low loss interconnects between a metal layer and a multilayer Printed Circuit Board layer. This will facilitate the integration of waveguide-based filters and antennas with the active components of the system. The stopband provided by the glide-symmetric EBG structure will remove any possible leakage and ensure a smooth transition between the different interfaces.

This document is organized as follows: Section 3 covers a theoretical overview of the basic concepts related with this project such as waveguides and planar transmission lines, transitions and metasurfaces. In Section 4 the design of the transition is deeply explained showing the keypoints on this design and how a non-perfect connection between waveguide and PCB can produce losses. Section 5 focuses on the study and design of the unit cell which will form the metasurface as well as the topology that this metasurface will have on the complete structure. Results after joining both structures are described in Section 6. Section 7 describes the manufacturing and measurement results, including a tolerance analysis. Finally, Section 8 summarizes the conclusions obtained with this project and states the future working lines.
3 Theoretical Overview

This chapter provides a general overview of theoretical concepts and already existing technologies related to the transition presented in this thesis. The two types of transmission lines which are employed in the design of the transition are rectangular waveguides and planar transmission lines. Since a transition between these two technologies will be designed, it will be shown how the transition technology is working nowadays, presenting some examples and references of already designed transitions.

3.1 Waveguides

Waveguide-based transmission lines has been extensively studied during a long period of time. Therefore, only some basic concepts of this type of transmission line will be discussed. For a more comprehensive overview of waveguide technology, see [10] - [13].

Waveguides can have a variety of cross-sectional shapes, but in this document, only the rectangular waveguide (RWG) will be used, which is simply a rectangular pipe with conducting inner surfaces, Fig. 1. The material chosen for building a waveguide should typically be a good conductor, and this can be realized by using either solid metal walls or metalized walls with the bulk material being for example some plastic material. Rectangular waveguides are typically classified with respect to their “width” $a$ (the wider dimension of the waveguide) and “height” $b$ (the narrower dimension), the former controlling the useful band of operation of a waveguide.

![Rectangular waveguide](image)

**Figure 1: Rectangular waveguide**

Transmission lines support the propagation of electromagnetic (EM) waves through them in the form of modes. A mode is a possible configuration of EM waves which satisfies the boundary conditions of the transmission line. Specifically,
the lowest order mode that an RWG supports is the TE\(_{10}\) mode, whose cut-off frequency (the frequency above which EM waves can propagate) is directly related to the “width” \(a\) of the RWG as

\[ f_c = \frac{c}{2a} \tag{1} \]

where \(c\) is the speed of light in the medium inside the waveguide, which is vacuum for the vast majority of waveguides.

In order to understand the principle of waveguide propagation it is necessary to solve the EM problem, and for that, Maxwell equations are needed. The solution to this can be found in [11]. As it is explained, only transverse electric (TE) and transverse magnetic (TM) modes can propagate through an RWG, since only one conductor is being used; no transverse electromagnetic (TEM) mode propagation is supported. Some examples of modes supported by an RWG are illustrated in Fig. 2, where red arrows represent the direction of the electric field and blue lines represent the distribution of the magnitude of the electric field over the wider dimension of the waveguide.

![Figure 2: TE modes in a RWG: (a) TE\(_{10}\), (b) TE\(_{20}\), (c) TE\(_{30}\)](image)

As can be observed, WG is a bulkier and heavier technology than other transmission media such as transmission lines using wires or planar transmission lines. This makes this technology more expensive and not suitable for all applications.

On the other hand, a direct benefit from a bigger conductor area is that the current density is much smaller and, therefore, the ohmic losses are also drastically reduced (Joule effect). It implies that the power which can be handled by a waveguide is much higher than for other transmission line technologies.
3.2 Planar Transmission Lines

A planar transmission line can be defined as a transmission line which whose conductors are flat. Some examples of planar transmission lines are microstrip lines, stripline, and coplanar waveguide (Fig. 3). Their configuration is simple, and low-cost when using printed circuit board (PCB) technology. This kind of transmission lines is used for filters, impedance matching devices, couplers, and many more components. They are generally composed of flat metallic strips, one or more ground planes which are parallel to the strips, and layers of dielectric substrates which separate them.

![Figure 3: (a) Microstrip line, (b) Stripline, (c) Coplanar line](image)

Striplines support TEM mode propagation, which is non-dispersive (if the dielectric substrate is not dispersive). On the contrary, microstrip and coplanar lines present non-TEM modes due to the use of different dielectrics (air on top of the line).

The use of microstrip lines is more widespread, because the lack of upper dielectric makes it easier to work with them, while the non-dispersive characteristic of striplines make them more suitable for wideband applications.
3.3 Transitions: State of the Art

The aim of this section is to offer a theoretical background of the most common transitions used for interconnecting the two already studied transmission technologies: planar transmission lines and waveguides.

Since the propagation of EM fields is different for each transmission line technology, an interconnecting component, which should be able to handle this difference, is needed. Its main objective is to carry out an efficient transition between two transmission line configurations with as low insertion loss as possible and avoiding reflections.

The main types of transitions are described: Radiating Element, Cavity Backshort, and Probe Insertion. Some examples of these technologies are introduced, and the advantages and disadvantages of each of them are also explained.

3.3.1 Radiating Element

This type of transition is also known as matching element. It consists on the use of an element, generally a patch antenna, which is located inside the waveguide and which generates an electric field. As was shown previously, the fundamental propagating mode inside the waveguide is the TE$_{10}$ mode, while the mode supported by a stripline is a TEM mode.

Figure 4: Example of a transition using a radiating (resonant) element
The resonance frequency of the radiating element transition depends strongly on the radiating element used and, given the use of a patch antenna, the bandwidth of this technology is limited. The position of planar transmission line with respect to the radiating element is used in order to tune the impedance matching.

As mentioned before, this transition can be characterized as narrowband (since the radiating element will only resonate around a particular frequency) but is easy to design. Several examples of this transition type can be found in [14], [15].

### 3.3.2 Cavity Backshort

The cavity backshort transition has been used for designing various microstrip to waveguide connections [14], [16]. The cavity backshort transition is designed by using a quarter-wavelength ($\lambda/4$) cavity on top of the dielectric substrate of the planar transmission line, as illustrated in Fig. 5.

![Figure 5: Example of a transition using a cavity backshort](image)

The mode transformation principle of this transition is almost the same as the one used between coaxial cables and waveguides. The $E$ field generated by the inserted probe enters the backshort and is then reflected back by the conducting plane. When the reflected field arrives back at the probe position, it has traveled half a wavelength ($\lambda/2$) and, because of the phase-shift at the reflecting plane, it adds in phase with field radiated by the probe. It is important to note that since
this field is propagating inside a waveguide, all wavelengths previously mentioned should be $\lambda_g$ (the guided wavelength of the TE$_{10}$ mode of the waveguide).

A probe can be added at the end of the line in order to tune the resonance frequency. In this transition, the size of the cavity is generally the same as the size of the used waveguide, but choosing different lengths of the cavity sides, the frequency band of the transition can also be tuned.

Regarding impedance matching, both the insertion of the probe and the cavity have a strong dependence on it. Cavity backshort transitions have generally a wider bandwidth than the transitions discussed in Section 3.3.1.

### 3.3.3 Probe Insertion

This particular type of transition is designed just using a planar transmission line ending with a probe with a determined shape. Both elements (planar transmission line and probe at its end) are introduced in the waveguide in a particular position which excites the TE$_{10}$ mode. Fig. 6 shows a basic example of this kind of transition. A detailed example of this transition can be found in [17] and it is generally used with stripline technology.

![Figure 6: Example of a probe insertion transition where a probe is directly inserted in the waveguide](image)

This transition is easily manufactured and offers good behaviour in terms of bandwidth. Unfortunately, in order to excite the TE$_{10}$ mode inside the waveguide, the line should be inserted as it is shown in Fig. 6. On this study, we want to place the waveguide below a PCB (which contains the planar transmission line). Therefore, using this type of transition is not an option for us.
3.4 Metasurfaces

A metasurface, also known as an artificial surface, can be defined as an engineered periodic structure which, due to its shape and material characteristics, offers macroscopic properties beyond those of its individual constituent materials. Such surfaces modulate the behavior of electromagnetic waves through specific boundary conditions, rather than the constitutive effective parameters in three-dimensional (3D) space. Particularly, these surfaces can generate boundary conditions which do not exist intrinsically in nature. For example, a metasurface can generate what is known as a Perfect Magnetic Conductor (PMC). Typical configurations of metasurfaces are holey metallic structures [8], metallic pins [18], and arrays of patches [19]. Fig. 7 provides an example of a basic artificial surface.

![Figure 7: Example of a basic metasurface generated by a periodic holey unit cell](image)

Figure 7: Example of a basic metasurface generated by a periodic holey unit cell

The study of these structures can be done in two ways:

- With a finite structure and the Transmission Coefficient ($S_{21}$). Using a simulation software, like CST, it can be done by building a structure with several unit cells and placing two ports on both ends of the propagation path that we are interested on analyzing.

- Studying a unit cell using an Eigenmode Analysis (see section 3.4.1)
The type of metasurface used (periodicity and type of unit cell) will determine its behaviour but they are generally divided into three groups [20]:

- **Soft-surfaces**: they stops wave propagation only in the transversal direction of the propagation (for example, they stops propagation in the orthogonal direction of a corrugation). Boundary conditions are \( E_t \approx E_n \approx 0 \) being \( E_t \) the tangential component and \( E_n \) the normal component of the electric field.

- **Hard-surfaces**: this second type of metasurface support the propagation of waves along certain directions (for example, along the corrugations in a corrugated surface). Boundary conditions are \( \partial E_t / \partial n \approx \partial E_n / \partial n \approx 0 \) being \( n = x \) or \( y \).

- **Electromagnetic Band-Gap (EBG) surface**: are a particular kind of metasurface which presents forbidden frequency bands. This is a particular frequency band in which it does not exist any kind of propagation in the gap between the two surfaces (it is also known as stopband, see section 3.4.2). Due to this bandgap property, these type of metasurfaces have several applications such as mutual coupling and back radiation reduction, filtering, and surface wave guiding.

### 3.4.1 Eigenmode Analysis

This kind of analysis is carried out without any excitation and it exploits the information provided by the energy stored in the structure. This analysis is done in terms of internal field for each mode and several modes can be analyzed. It solves the eigenvalue problem associated with a determined structure making the assumption of having an infinite periodic structure.

The boundary conditions used in this analysis are “periodic” in x-y plane and PEC (\( E_t = 0 \)) in z-direction (top and bottom planes). A number of points are studied and used to built the dispersion diagram, where the behaviour of the structure is shown in terms of the modes which propagates through itself.
3.4.2 Dispersion Diagram

Using an eigenmode analysis on CST the dispersion diagram is obtained. A dispersion diagram consist on a graph which shows the frequency necessary to obtain a certain phase shift over one cell of the periodic structure. This dispersion term is related with the propagation constant, \( \gamma = \alpha + j \beta \), which is composed by the two terms:

- Attenuation constant: \( \alpha \)
- Phase constant: \( \beta \)

In free space conditions, the function \( \beta(\omega) \) for the lossless case (\( \alpha = 0 \)) is

\[
\beta(\omega) = k_0 = \omega \sqrt{\mu_0 \varepsilon_0}
\]

(2)

where \( k_0 = 2\pi/\lambda_0 \) is the wave number in free space.

However, the previous equation can only be used under the expressed conditions (free space and lossless). When waves are propagating through other structures, it is difficult to find an expression for \( k \). Here is where the eigenmode solver is used, since it solves the eigenvalue equations. The wave number is obtained after solving this problem. It is important to remark that several solutions can be found for an eigenvalue equation, each one of these solutions is known as a mode. The dispersion diagram is thus representation of the relation between \( \beta \) and \( \omega \) for each of these modes.

Therefore, in the horizontal axis of the dispersion diagram, \( \beta d \) is depicted, where \( \beta \) is the propagation constant in the structure expressed in rad/m and \( d \) is the distance travelled. The former product is swept from 0 degrees to 180 degrees, and then a normalization factor of \( \pi \) is applied, so \( 0 \leq \beta d/\pi \leq 1 \).

When this kind of analysis is done, since the studied structure is periodic in both the x- and y-directions, the propagation in both directions should be analyzed. In particular, this analysis has to be carried out over what is known as the Irreducible Brillouin Zone (IBZ). This concept will be explained for a unit cell of the metasurface shown in Fig. 7, which corresponds to a single hole. For this example, a square unit cell will be used, as shown in Fig. 8.
The irreducible Brillouin zone is defined as the minimum section of the unit cell that can be used to obtain the whole unit cell just by mirroring this minimum section. On this example, this minimum section would be a triangle whose vertices are denoted Γ, X, and M, see Fig. [9]. Therefore, a dispersion diagram with three different intervals will be obtained. In the first interval, which goes from Γ to X, the frequency of a mode propagating in X direction is increased. When a frequency shift (βd) of 180 degrees has been obtained, that frequency is kept for the mode in X direction and start increasing the frequency for a mode in Y direction, until another 180 degree phase shift is obtained (in Y direction this time). This corresponds to the analysis from points X to M. Finally, a third interval of the diagram (from M to Γ) represents the propagation both in X and Y directions at the same time. The total frequency in this process is obtained knowing that $k = \sqrt{k_x^2 + k_y^2}$. 

Figure 8: Holey unit cell example. The dimensions are: $a = 3.5 \text{ mm}$, $g = 0.05 \text{ mm}$, $h = 2 \text{ mm}$ and $r = 1.4 \text{ mm}$. 

26
As was mentioned before, this three interval case is limited to structures like the one in the picture, where the unit cell is equal in both X and Y directions (a square unit cell). If a rectangular unit cell is being studied (which does not present equal length in X and Y directions), four intervals would be needed in order to calculate the dispersion diagram, since the IBZ will be a square instead of a triangle.

Fig. 10 represents a simple example of a dispersion diagram corresponding to the unit cell of Fig. 8. This study is based on [8] and the chosen dimensions are: \( a = 3.5\text{mm} \), \( g = 0.05\text{mm} \), \( h = 2\text{mm} \) and \( r = 1.4\text{mm} \).
From this graph it is possible to obtain a large amount of information with a quick glance. The most attractive part of this diagram in relation with this project is what is known as the stopband and it corresponds to the frequency band where no modes are propagating. Finally, an extra information that can be obtained looking at this graph is the Group Velocity \( V_g = \partial \omega / \partial \beta \). Therefore, for transmission purposes, one would like to aim for a constant slope diagram in order to have a low dispersive mode.
3.5 Glide Symmetry

Glide symmetry is a type of higher symmetry. A periodic structure possesses higher symmetries if it is created with more than one operator ([21], [22]). Single operators are, for example, translation, reflection and rotation ([23] - [26]). Glide symmetry is obtained just by applying a reflection over a plane and translating this reflected object half of its period ([27] - [30]). This can be observed in Fig. 11. Glide symmetries have been investigated in slotted lines, co-planar technology and printed bifilar lines ([31] - [33]).

![Glide Symmetry Diagram](image)

Figure 11: Example of glide symmetry in a simple structure

Regarding the unit cell of the example, Fig. 12 shows how the unit cell of Fig. 8 will look after applying glide symmetry on it.

![Unit Cell Diagram](image)

Figure 12: Unit cell applying 2D glide symmetry.
When using glide-symmetric structures, a number of characteristics can be observed when studying the dispersion diagram of this kind of unit cells, all of them are shown in Fig. 13 and they are:

- Increase in the bandwidth of the stopband (compare with Fig. 10).
- Stopband appears between second and third mode.
- Mode 1 and 2 join at point X of the dispersion diagram.

Figure 13: Dispersion diagram of the unit cell after applying 2D glide symmetry.

Note that in this study two different glide symmetries will be mentioned; in order to explain them, the XY-plane is selected (Fig. 14): 

1. **2D glide symmetry**: reflection operation is applied, as well as a translation of the structure half a period of the unit cell in both X and Y directions.

2. **1D glide symmetry**: reflection operation is applied but translation is applied only along one of the two axis (X or Y).
2D glide symmetry is the one mainly used in previous studies as \([7]\) and \([8]\). But due to the constrains of the design, a deeper study of 1D glide symmetry will be studied during this document.

Figure 14: (a) 2D glide symmetry, (b) 1D Glide symmetry.
4 Transition Design

During this chapter, we will design a transition between stripline and waveguide technology. The requirements are:

- Reflection coefficient $S_{11}$ below $-20 \, dB$ for at least 20% bandwidth.
- Insertion losses ($S_{21}$) as low as possible on that frequency range.

After studying and testing all explained transitions (section 3.3), the cavity backshort option is chosen. It presents a wideband behaviour and, as it will be demonstrated later, after applying some changes to that cavity it will be easily embedded inside the PCB structure.

![Diagram of transition design](image)

Figure 15: First approach of the transition design: air filled cavity backshort

Fig. 15 shows this design. The two transmission media which should be connected are:

- **Stripline**: built using Megtron 7 as substrate. Two varieties of this material are used:
  1. R-5785(N) Core: $\epsilon_r = 3.34$, $\tan \delta = 0.003$ and height of 0.2 $mm$.
  2. R-5680(N) Prepreg: $\epsilon_r = 3.32$, $\tan \delta = 0.003$ and height of 0.132 $mm$.

In order to have a characteristic impedance ($Z_0$) of 50$\Omega$, the width of the stripline is 0.15 $mm$ and a standard metal thickness of 18 $\mu m$.
- **Waveguide**: Standard waveguide WR28 whose dimensions are: width \(a = 7.112\) mm and height \(b = 3.556\) mm. This waveguide presents a frequency band of operation suitable for this project, having a cutoff frequency of the lowest order mode (TE\(\text{_{10}}\)) of 21.077 GHz and a cutoff frequency of next mode of 42.154 GHz.

After tuning different parameters such as cavity size, position and insertion of the stripline inside the cavity, the behaviour of this transition is shown in terms of S-Parameters. The graph below shows the reflection coefficient and the insertion losses for the range of frequencies around 28 GHz. The dimensions are: cavity sides keep the dimensions of WR28 waveguide, cavity height of 1.5 mm and stripline insertion on the cavity of 1.6 mm.
4.1 Integration in the PCB

Compactness is a key point on this project, therefore the use of a bulky metallic cavity as the one previously designed is not a suitable choice. The options that PCB technology offers for embedding the transition are shown in this section. This integration has been done attending two features: the inner part of the cavity (dielectric-filled cavity) and the cavity boundaries (metallic cavity walls and lid).

4.1.1 Cavity filling

The inner part of the cavity is filled with a dielectric material. This presents several advantages in relation to the air filled one:

- Removing the dielectric from the cavity requires the use of milling (increase the cost).
- A dielectric substrate filling the cavity would reduce its height. Ideally, the cavity height should be $\lambda_g/4$. The use of a material with a higher $\epsilon_r$ implies a reduction of the height by a factor of $\sqrt{\epsilon_r}$.
- Having a dielectric layer on top of the already existing layers makes possible to build the complete transition using PCB technology, with the consequent cost reduction.

Originally, a single dielectric substrate was used for the whole cavity. However, it was necessary to divide this thick block in several layers due to manufacturing constrains. Each substrate layer is isolated by a thin metal layer of approximately 18 $\mu$m. The substrate layers are laminated with standard thicknesses and the manufacturer requires that a layer of R-5680(N) (prepreg) is placed between two consecutive layers of R-5785(N) (core). Combining these constrains the final stack for the PCB is obtained, it is illustrated in Fig. 18. The cavity height is $\lambda_g/4$. 

34
Figure 18: Stack up of all the dielectric and metal layers of the PCB. Dielectric constant varies depending on the thickness of each layer.

Megtron 7 is also used for the cavity filling. Fig. 19 shows the cross-section of the cavity backshort while Fig. 20 shows a top view of the transition. Finally, Fig. 21 shows S-parameters of the structure.

Figure 19: Cross-section (x-plane) of the dielectric-filled cavity.
Figure 20: Top view of the transition. Cavity lid has been hidden in order to allow the view of the inner part of the cavity.

Figure 21: $S_{11}$ and $S_{21}$ of the dielectric-filled cavity.

Comparing the results of the air filled cavity (Fig. 17) with the dielectric filled cavity (Fig. 21), it is demonstrated that the use of dielectric inside the cavity improves the transition in terms of bandwidth while the increase of insertion losses is negligible. The dimensions of this design are: cavity sides 6 $mm$ (wide side), 3.3 $mm$ (narrow side) and insertion of the stripline in the cavity 1.35 $mm$. 
4.1.2 Cavity boundaries

We consider as boundaries both, cavity walls and cavity lid.

Regarding the walls of the cavity, fully metallic walls should be avoided when manufacturing (it is not a common structure on PCB). These walls can be easily replaced by the use of vias surrounding the space of the cavity. This approach does not behave as well as the ideal metallic wall but, using a narrow pitch between vias and making use of the several metal layers which isolate the dielectric layers of the PCB, an good enough behaviour can be obtained. Fig. 22 clearly shows how this “grid” is built using both, vias and metal layers.

![Cross-section of the transition. The grid shaped by the intersection of vias and metal layers is clearly shown.](image)

Fig. 22: Cross-section of the transition. The grid shaped by the intersection of vias and metal layers is clearly shown.

Fig. 23 presents the top view of the transition after replacing the metallic walls by vias. As it is highlighted, there is a missing via in the position of the stripline, it is designed in this way in order to allow the insertion of the stripline inside the cavity. A second row of vias is placed in left and right side. The purpose of this second row is to avoid the possible leakage which is produced due to the use of a bigger pitch between vias on these two sides.
It exists an important design constraint regarding vias. If they are located in the edge of a metallic layer (this is clearly observed in Fig. 23), a pad should be added in order to guarantee the contact between vias and metallic layer. The use of this pad is shown in Fig. 24 where again, dielectric layers and cavity lid have been hidden to make easier the visualization of the image.

Figure 23: Top view of the cavity-backshort with via position. Dielectric layers and cavity lid have been hidden for helping in the view.

Figure 24: Cavity-backshort design including pads for guaranteeing the contact between vias and metal layers.
It can be seen in Fig. 25 that the use of vias mimics the behaviour of a metallic wall, avoiding that EM fields leak out from the cavity.

Figure 25: Cavity backshort top view. No EM fields go out of the cavity due to the use of vias.

S-parameters of the structure are shown in Fig. 29. When comparing them with the ones of Fig. 21 it can be seen that the bandwidth (in terms of reflection coefficient) is smaller. There are different reasons for this bandwidth reduction:

- **Metal layers opening**: cavity dimensions are determined by the size of the openings on each metal layer. On this design, openings in all metal layers are equal (wide side = 6 mm and narrow side = 3.6 mm) but different combinations can be built.

- **Use of pads**: the shape of the cavity walls is not flat anymore. Fig. 26 shows how metal layers look now. Diameter of the pads is 0.6 mm.

Figure 26: Top view of a metal layer. It can be seen the new shape of the cavity due to the addition of the pads.
• **Via position**: the position of the vias is an important design parameter. Wide side of the cavity determines the cut-off frequency of the transition and, in turn, vias position determine the size of the two sides of the cavity. Therefore, there is a direct relation between the position of the vias and the frequency band of the transition. Fig. 27 shows the position of vias for this particular design. These positions have been chosen trying to obtain the widest possible bandwidth.

• **Vias shape and pitch**: ideally, vias should be in contact to mimick the behaviour of an ideal metal wall. However, due to manufacturing constrains and the use of the pads, a certain pitch between vias should be kept. If this pitch is small enough, changes should be negligible. On the contrary, the use of cylindrical vias involves a slightly change on S-parameters since they differs from the ideal flat surface which a metallic wall presents. Via pitch is shown in Fig. 27.

![Figure 27: Top view of the transition showing the position of vias. Distances are expressed to the center of the vias. Via diameter = 0.3 mm](image)

All these constrains introduce several new tuneable parameters: via pitch and position, pad radius and number of vias among others. The transition model used for the simulation is illustrated in Fig. 28. Fig. 29 presents the obtained S-parameters for the chosen values shown in Fig. 26 and 27. On this model the length of the stripline is 9 mm while the length of the waveguide is 10 mm.
Finally, a study of different shapes of the cavity lid has been performed. Fig. 30 shows some of the lids studied during this step. No major improvements are obtained changing the shape of the lid (Fig. 30 (a) and (b)). In addition, the use of a different shape could imply more difficulties in the manufacturing process. Therefore, a flat lid is chosen (Fig. 30 (c)). It is built as one more metal layer of the whole PCB structure.
Figure 30: Some examples of studied cavity lids: elliptical lids (a) and (b), flat lid (c).
4.2 Leakage Study

Manufacturing processes are not perfect, that means that all techniques used to create PCBs and waveguides present some inaccuracies. This implies that, generally, simulations offer a better result than the real performance of the manufactured object. During this section, the effect of having a “non-perfect” connection in the transition between a stripline and a waveguide is studied.

The main methods used to join these two structures are generally the use of some type of conductive adhesive, soldering or screwing both surfaces. An air gap or air bubbles between both surfaces can be generated during these processes. The immediate consequence of this non-perfect connection is the loss of power by means of leakage through this air gap. In order to model this air gap, a simple separation between PCB and waveguide will be simulated. Fig. 31 illustrates the transition including a simple model of the air gap due to an imperfect connection.

![Figure 31: Cross-section of the transition showing the non-perfect connection between PCB and waveguide flange.](image)

As can be seen in Fig. 32, the power received at the transition end is directly depending on the thickness of the gap between PCB and waveguide. This is expressed in the form of bigger insertion losses when using the transition. These insertion losses are directly related with $S_{21}$ parameter:

$$IL = -20\log_{10}(S_{21})dB$$  \hspace{1cm} (3)
Figure 32: $S_{21}$ parameter of the transition for several air gap thicknesses.

Fig. 32 clearly demonstrates how the amount of received power decreases when the gap aperture increases. Assuming an ideal case of no air gap the insertion losses are around 0.4 dB at 28 GHz, while in the worst considered case (having 0.1 mm air gap thickness), leakage increases the amount of lost power and, which is more critical, some peaks of losses appear at certain frequencies.
5 Leakage Reduction Using an EBG

Once the transition is working according to the specifications and the negative effects of an air gap in the transition have been demonstrated, the leakage will be mitigated. In order to reduce the leakage, an EBG structure will be placed surrounding the transition (§31, §35). In this section, the topology of the metasurface is characterized through the study of its unit cell.

5.1 Metasurface Topology

One of the aims of this project is the suppression of the leakage demonstrated in section 4.2. In order to achieve this goal, a metasurface is designed. Before choosing the metasurface, the available space must be considered. The topology of the metasurface can have different configurations. Two of them are illustrated in Fig. 33. Each color refers to each of the two possible plates in which the hole can be positioned (upper and lower layers).

![Metasurface Topology](image)

Figure 33: Metasurface topology: (a) Rectangular and (b) Circular.

Some advantages and disadvantages of these two topologies are:

a) **Rectangular topology**:

- Advantage: easily integrable with an array structure sharing unit cells.
- Disadvantage: corner unit cells are different and their behaviour should be studied.
b) **Circular topology:**

- Advantage: uniform behaviour of all unit cells.
- Disadvantage: bulkier when for array structure. Unit cells cannot be reused for several transition (overlap between them).

Considering these advantages and disadvantages, the option (b) has been chosen. Using this topology, only a single unit cell has to be designed, which is a big advantage in terms of design simplicity. Despite that option (a) is easier to integrate, the amount of space which is saved is not that substantial. Therefore, using a circular topology still makes possible to use this technology in array structures.
5.2 Unit Cell Design

Here, we describe the design of a unit cell which accomplish the requirements of this project. This is, presenting a stopband for the whole bandwidth in which the transition is working for a given air gap that could appear between the waveguide and PCB.

As starting point, we design a unit cell like the one in Fig. 34a. All dimensions are adapted for centering the stopband around 28 GHz. In [7] is stated that: “the center of the stopband is related to the periodicity “a”, which fixes the central frequency”. Therefore, due to the direct dependence between the size of the unit cell (a) and the location of the stopband (central frequency), the required dimensions are: \( r = 2.6 \, mm \), \( h = 1.5 \, mm \), \( a = 6 \, mm \) and \( g = 0.1 \, mm \).

The dispersion diagram of this particular unit cell is illustrated in Fig. 34b. The stopband generated by this unit cell does not completely match the requirements of having a stopband bandwidth of 20% at 28 GHz.

Figure 34a: Holey unit cell on the left and its IBZ on the right.
Figure 34b: Dispersion diagram of the holey unit cell. The dimensions are $r = 2.6 \, mm$, $h = 1.5 \, mm$, $a = 6 \, mm$ and $g = 0.1 \, mm$.

The next step is to introduce glide symmetry to increase the bandwidth in order to accomplish the requirements for the design. Fig. 35a shows the glide-symmetric unit cell.

Fig. 35b shows the dispersion diagram of the new structure, for the same dimensions of the case of non-glide. Glide symmetry significatively increases the bandwidth with respect to the non-glide.

Figure 35a: Glide-symmetric holey unit cell on the left and its IBZ on the right.
However, this unit cell cannot be used for our objective which is to avoid the propagation between a PCB and a waveguide. Each plate of the unit cell must be built on a different structure. Fig. 36 illustrates the cross section required for the unit cell. The bottom plate is made of metal while the top plate is a structure made of several layers (metal in grey color and dielectric substrate in blue and purple).

A hybrid unit cell is needed in this specific case. In order to enhance the bandwidth both layers must behave similarly to produce an equivalent glide symmetry. This is, we will try to achieve the same dispersion diagram using two different structures. In other words, we will mimic glide symmetry. For this, we use two quotidian tools of PCB manufacturing: metal etching and use of vias.
In order to create the holey structure on the upper plate it is needed to etch the metal from several metal layers of the PCB. This etching can generate (see Fig. 37) different shapes but we decide to use circles to follow the pattern of the lower plate. The radius of the etched circle is defined as $r_{etch}$ and it is a design parameter.

A study of the impact of the hole depth on holey metasurfaces is carried in [7]. The authors demonstrate that a minimum depth is required in order to maximize the bandwidth. Following that, metal is etched from all metal layers but the top one (cavity backshort lid and unit cell lid). On this design, $r_{etch}$ is kept constant for all layers to make it simpler.

Vias are placed around the etched metal to create a cavity (in the same way as we did for the transition). They are used to avoid the propagation of EM waves between metal layers through the dielectric substrate. The cavity works as the drilled hole on the lower plate and it can have different shapes depending on how vias are placed (see Fig. 37). In order to generate a unit cell similar to the studied on Fig. 35a, we place vias creating a ring whose radius is defined as $r_{cav}$ and that can also be tuned. Pitch between vias should be small enough to confine the EM fields in the cavity.
Fig. 37 illustrates some examples of cavities and etched metal for a unit cell. On this thesis, option (a) is chosen but the possibilities offered by this technology are huge.

![Diagram of different configurations for the metal layer of the upper plate. Vias are drawn as darker grey dots.](image)

Figure 37: Different configurations for the metal layer of the upper plate. Vias are drawn as darker grey dots.

The dielectric substrate that is inside the cavity can be removed or kept. If it is removed the similarity between lower and upper plate will be almost perfect. Fig. 38a shows this unit cell. This possibility has been simulated and its results are shown in Fig. 38b.
Figure 38a: Hybrid unit cell with glide-symmetric air filled holes in both plates on the left and its IBZ on the right.

Figure 38b: Dispersion diagram of the hybrid unit cell with glide-symmetric air filled holes in both plates. Dimensions are: $r_{etch} = r_{cav} = 2.8\ mm$, $r_{wg} = 2.8\ mm$, $h = 1.5\ mm$, $a = 7\ mm$ and $g = 0.1\ mm$.

This option is an easy solution for mimicking glide symmetry using hybrid structures. Nevertheless, a milling procedure is needed to remove the dielectric substrate resulting on an increase of cost and manufacturing complexity.

Therefore, unit cell of Fig. 39a is selected. All dielectric substrates are kept and its dispersion diagram is shown in Fig. 39b.
The study of the 2D glide-symmetric unit cell shows that this unit cell achieves the requirements of bandwidth after applying glide symmetry. It has also been proved that the behaviour of a holey fully metallic unit cell can be mimicked using a hybrid unit cell.
Next step is to study a more accurate model of the unit cell. During section 5.1 we decided to use a *Circular Topology* for placing the metasurface around the transition. In order to model this unit cell, we will use “1D glide symmetry”. Fig. 40 illustrates a comparison between both unit cells (2D vs 1D). Notice that the IBZ of this new unit cell changes with respect to all previous studied cases. It has a square shape.

![IBZ comparison (a) 2D glide symmetry, (b) 1D glide symmetry.](image)

Figure 40: IBZ of both unit cells: (a) 2D glide symmetry, (b) 1D glide symmetry.

We do an assumption regarding the understanding of the dispersion diagram when the metasurface is placed on a circular topology. We consider the center of the transition as the source of the leakage. The EM fields propagate between the two plates of the metasurface in the radial direction from that central source. This is illustrated in Fig. 41 by the dotted red arrows. The study of the dispersion diagram can be focused only on its first region (from Γ to X) since most of the propagating waves will follow this path. Therefore, we assume that if we have a proper stopband in that region, leakage will be suppressed.

The unit cell that has glide symmetry only in one axis, does not correspond exactly with the real situation, but it emulates the behaviour of this scenario with higher accuracy than the 2D glide-symmetric unit cell. Fig. 42a illustrates the unit cell with 1D glide symmetry while Fig. 42b shows its dispersion diagram.
Figure 41: Demonstration of why only the first region of the dispersion diagram is studied.

Figure 42a: Hybrid unit cell with 1D glide symmetry on the left and its IBZ on the right.
Figure 42b: Dispersion diagram of the hybrid unit cell with 1D glide symmetry. Dimensions are: $r_{etch} = 2.5 \text{ mm}$, $r_{cav} = 3 \text{ mm}$, $r_{wg} = 2.8 \text{ mm}$, $h = 1.5 \text{ mm}$, $a = 7$ $\text{mm}$ and $g = 0.1 \text{ mm}$.

Previous dispersion diagram shows that a stopband from 22 GHz to almost 31 GHz is obtained on the region from $\Gamma$ to $X$, accomplishing the requirements of 20% bandwidth.

In order to prove that the assumption of only studying the region from $\Gamma$ to $X$ is correct, a second study will be carried out regarding $S_{21}$ parameter. Fig. 43 illustrates the model used for this study. We place a port on each side of the unit cell emulating a wave propagating along the unit cell following the path described by the red dotted arrow (radial propagation from the source). Fig. 44 shows a 3D view of the model. Waveguide ports are placed on two sides of the unit cell and PMC boundary conditions on the other two.
Fig. 45 shows the $S_{21}$ parameter of the modeled hybrid unit cell with 1D glide symmetry. We see how the stopband matches the frequency range defined in Fig. 42b, approximately from 23 GHz to 33 GHz (similar values as the obtained in the dispersion diagram). It is also shown that both studies match on the frequencies in which modes 2 and 3 start to propagate (around 22 GHz and 33 GHz respectively).
Figure 45: $S_{21}$ parameter of the hybrid unit cell with 1D glide symmetry.
6 Complete Structure

Both, transition and unit cell have been designed. In this section we will join them in order to generate the complete structure. A single and a back to back transition are studied with and without the addition of the metasurface.

Fig. 46 illustrates the cross section of all the different layers of the final structure as well as how they will be joined together with the use of screws. The upper plate is just made of a PCB while the lower plate is made of a metal block where the metasurface is drilled. Waveguide flanges are screwed to this metal block.

![Cross-section of the complete structure.](image)

6.1 Single transition

The first step is the study of a single transition without the use of the metasurface around. Fig. 47 shows how this transition looks like.

![Single transition diagram](image)
Fig. 48 shows the reflection coefficient ($S_{11}$) and insertion losses ($S_{21}$) of the single transition. The behaviour of reflection coefficient and insertion losses are shown on Fig. 49 and Fig. 50 for a variable air gap thickness.

Figure 48: $S_{11}$ and $S_{21}$ parameters of the single transition.
Figure 49: Study of $S_{11}$ parameter of the single transition without metasurface for several air gap thicknesses.

Figure 50: Study of $S_{21}$ parameter of the single transition without metasurface for several air gap thicknesses.
6.2 Single transition with metasurface

Now we place the metasurface around the transition to check that leakage is properly suppressed. Fig. 51 shows a single transition with the designed metasurface around.

![Figure 51: (a) Single transition including metasurface, (b) Top view of the PCB plate and (c) Top view of the waveguide plate.](image)

Holes are placed at a distance, $d = 7.5 \, \text{mm}$, of the center of the transition. Due to the size of the unit cell a total amount of six holes will be placed on both, upper and lower layer. They are rotated an angle $\alpha = 60^\circ$ obtaining the desired 1D glide symmetry. Fig. 52 illustrates how the complete structure looks.

For the design of the unit cell, 24 vias are used to emulate the cavity (as described in section 5.2). Holes on the upper layer are placed in such a way that stripline interacts as little as possible with the metasurface around it. Two different colors have been chosen for helping in the distinction of the two plates: yellow for the metal layers of the PCB (upper plate) and dark grey for the metal block (lower plate).
Figure 52: Top view of the transition including the metasurface around. The plane has been cut at the stripline layer and substrate layers have been hidden.

Fig. 53 shows $S_{11}$ and $S_{21}$. There is a slight change in $S_{11}$ when comparing with Fig. 48, this is mainly due to the presence of metasurface’s vias (see Fig. 52). These vias interact with the stripline, creating a small reflection. Regarding reflection coefficient and insertion losses, Fig. 54 and Fig. 55 show the behaviour of the structure when the air gap is increased.

Figure 53: $S_{11}$ and $S_{21}$ parameters of the single transition with metasurface.
Figure 54: Study of $S_{11}$ parameter of the single transition with metasurface for several air gap thicknesses.

Figure 55: Study of $S_{21}$ parameter of the single transition with metasurface for several air gap thicknesses.

Fig. 55 demonstrates that leakage is mostly suppressed after the use of the metasurface. Insertion losses are almost identical for every studied air gap thickness, which makes the transition more robust (in terms of insertion losses) to an imperfect connection. In order to show a clearer comparison between using or not the metasurface, Fig. 56 overlaps the information of Fig. 49 and Fig. 54 while
Fig. 57 combines the information obtained in Fig. 50 and Fig. 55.

Figure 56: $S_{11}$ comparison when using or not the metasurface.

Figure 57: $S_{21}$ comparison when using or not the metasurface.
6.3 Comparison with already existing technologies

Previous section proves that leakage can be suppressed by the use of a metasurface. However, there are other technologies which already try to suppress leakage in the connection between waveguides. Particularly, we will study choke flanges and corrugations.

Choke flange is illustrated in fig. 58. It is built milling a narrow ditch around the waveguide. Its frequency band operation is delimited by the air gap between the two flanges and the depth of the ditch. It is milled only on one of the flanges. Fig. 59 shows the comparison between using a choke and the designed metasurface.

Figure 58: Illustration of a choke flange.
Figure 59: $S_{21}$ comparison between the choke flange and the designed metasurface.

Regarding the corrugations, this metasurface is built by milling several ditches (thicker than the one in the choke flange) around the waveguide opening. Fig. 60 illustrates an example of this technology. The parameters that control the behaviour of the corrugations are the depth of the corrugations (which should be of $\lambda/4$) and the thickness of both, ditch and metal between two consecutive ditches. Tuning these parameters we can design a corrugation which properly works on the desired frequency band. Fig. 61 shows a comparison between the use of corrugations and the designed metasurface.

Figure 60: Illustration of the corrugations.
We observe that the behaviour of the choke flange is not as good as the one obtained using the metasurface designed during this project. The obtained stop-band bandwidth is not as big as we need and it can be seen the direct dependence on the thickness of the air gap. Regarding the corrugations, we can see how the performance obtained by both technologies is almost the same. However, the use of ditches (for both choke and corrugations) implies the use of milling. It increases substantially the manufacturing costs of these “anti-leakage” structures compared with the simple use of drilled holes that we propose on this project.

Figure 61: $S_{21}$ comparison between the corrugations and the designed metasurface.
6.4 Back to back transition

The single transition is not a good option for manufacturing due to the use of the waveguide port that is used to feed the stripline during simulations. We now study a back to back transition (waveguide to stripline to waveguide). This design considers a stripline of an approximate length of $10\lambda$ between both single transitions. This distance has been chosen due to several reasons:

- Having a long enough distance between transitions in order to avoid possible unwanted interactions between them.
- During the assembly process after manufacturing, waveguide flanges should be screwed to the metal block and these flanges present a considerable size. Therefore, waveguides cannot be placed so close to each other.
- Knowing the characteristics of the dielectric substrate, losses per unit length can be calculated and subtracted from the model. It makes that the distance of this line has no impact on the results.

Fig. 62 shows the back to back transition. No changes have been applied to the previous structure but mirroring and displacement of the whole single transition.

![Figure 62: Back to back transition](image)

Fig. 63 shows reflection coefficient and insertion losses of this structure while Fig. 64 and Fig. 65 show how these two parameters evolve when increasing the thickness of the air gap between lower and upper plate.
Figure 63: $S_{11}$ and $S_{21}$ of the back to back transition having no air gap.

Figure 64: Study of $S_{11}$ parameter of the back to back transition without metasurface for several air gap thicknesses.
Figure 65: Study of $S_{21}$ parameter of the back to back transition without metasurface for several air gap thicknesses.
6.5 Back to back transition with metasurface

The final step of this simulation study is the use of the metasurface in the back to back transition. As it was done previously, this new structure is obtained just by mirroring and displacing the structure studied during previous section. Fig. 66 illustrates the complete structure.

Figure 66: Back to back transition including the metasurface on both

Fig. 67 shows reflection coefficient and insertion losses of the back to back transition including the metasurface. Fig. 68 and Fig. 69 show how reflection coefficient and insertion losses change depending on the air gap thickness.

Figure 67: \( S_{11} \) and \( S_{21} \) of the back to back transition including a metasurface around.
Figure 68: Study of $S_{11}$ parameter of the back to back transition with metasurface for several air gap thicknesses.

Figure 69: Study of $S_{21}$ parameter of the back to back transition with metasurface for several air gap thicknesses.

Fig. 70 overlaps the information of figures 64 and 68 regarding reflection coefficient while Fig. 71 compares Fig. 65 and Fig. 69 regarding insertion losses.
Figure 70: $S_{11}$ comparison when using or not the metasurface in the back to back transition.

Figure 71: $S_{21}$ comparison when using or not the metasurface in the back to back transition.

The leakage suppression effect is more remarkable on this back to back structure and it can be clearly seen in Fig. 71.
7 Manufacturing, measurements and tolerance analysis

During previous sections several simulations on CST Microwave Studio were carried out in order to demonstrate both objectives of this project: the design of a stripline to waveguide transition working at 28 GHz, and the leakage suppression using a hybrid metasurface. During this section, we will explain manufacturing issues as well as real measurements and some tolerance analysis.

7.1 Manufacturing

The manufacturing process can be splitted in two different parts, each one corresponding to one of the main components of the whole structure: PCB layer and waveguide layer.

The layout of every metal layer was designed, containing the information regarding etched metal, vias, pads, screws and layer thicknesses. MEGTRON7(N) was chosen as dielectric substrate for both, core (R-5785(N)) and prepreg (R-5680(N)) layers. Fig. 72 illustrates these metal layers, particularly the second metal layer containing the stripline.

Figure 72: Top view of the metal layer containing the striplines.
On previous image, each red dotted rectangle makes reference to a back to back transition. Two areas are distinguished:

a) Back to back transitions which do not have a surrounding metasurface.

b) Back to back transitions which have a surrounding metasurface.

Transitions 3, 4, 5, 6 and 9, 10, 11, 12 have been placed close to each other in order to study the crosstalk that could be produced between them. Transitions 1, 2 and 7, 8 are away from the rest in order to study how they work having any disturbance. Fig. 73 shows the top view of the manufactured PCB, it follows the sketch of previous figure.

Figure 73: Top view of the PCB.

Regarding the metal block, it consists on an aluminum block in which some apertures are milled (WR28 dimensions). The metasurface is drilled around some of these waveguide apertures. In addition, through holes are drilled at the same position where metal is etched on PCB. These holes will be used to join both, PCB and metal block using $M_3$ standard screws. Fig. 74 shows how the manufactured metal block looks. On its other face, some more holes are drilled around the waveguide aperture to attach the standard waveguide flanges. These flanges use standard screws #4-40UNC with length 9.5 mm.
Figure 74: Top view of the metal block.
7.2 Measurements

After manufacturing the prototype, several measurements were done using the VNA (Vector Network Analyzer). Transitions from coaxial to waveguide (WR28) were needed and the calibration was done using the TRL (Thru, Reflect, Line) method.

The first aspect studied of the prototype was its repeatability. In the lack of air gap, the transitions 2, 4, 6, 8, 10 and 12 (see Fig. 72) were studied. Fig. 75 and 76 show their behaviour.

It can be seen that every transition behaves in a similar way in terms of both, $S_{11}$ and $S_{21}$ parameters. The differences can be explained because the number and position of screws placed around each transition. This different shape in the screws placement can cause a mismatch in the performance on each transition, but, in general terms, they all behave similarly.

![Graph showing the $S_{11}$ parameter of several transitions](image)

Figure 75: $S_{11}$ parameter of several transitions with $g = 0$. 

78
Figure 76: $S_{21}$ parameter of several transitions with $g = 0$.

Fig. 79 shows the measurement of $S_{21}$ for different air gap thicknesses of a back to back transition without and with a metasurface. These gaps are:

- **No air gap**: PCB and metal block are just screwed together as tight as possible.

- **30 µm air gap**: it is obtained by using a thin paper sheet. Paper is removed around the transitions and metasurface areas. Fig. 77 shows how it is placed.

Figure 77: 30 µm thickness paper sheet placed on top of the metal block.
- **65 µm air gap**: it is obtained by using a metallic tape. It is placed around the screw holes as it is illustrated in Fig. 78.

Figure 78: 65 µm thickness metal tape placed around the screw holes.

Figure 79: $S_{21}$ parameter for several air gap thicknesses.

Fig. 80 and 81 show a comparison of $S_{21}$ parameter between the obtained measurements and the simulations of the back to back transition without and with metasurface shown during section 6.4 and 6.5.
As can be seen, simulations are more optimistic regarding losses but the results obtained during the measurements are close to the simulated ones. This slight change can be caused by manufacturing tolerances during the PCB manufacturing process. There is a small frequency displacement, it can be caused for a possible misalignment between the PCB and the metal block (see section 7.3.2).
An important application of the metasurfaces is the reduction of the crosstalk between elements. For that reason, a study of the crosstalk has been carried, focusing on two types of crosstalk: H-plane crosstalk (between transitions 4 and 6 (no metasurface) and transitions 10 and 12 (using metasurface)), and E-plane crosstalk (between transitions 3 and 4 (no metasurface) and transitions 9 and 10 (using metasurface)). Matched loads are used in order to avoid reflections. Regarding E-plane crosstalk, ports were placed in the far end of the transition to avoid that port and load touch each other. Fig. 82 shows how this study is done while Fig. 83 and 84 show the obtained results.

Figure 82: a) H-plane crosstalk study and b) E-plane crosstalk study.

Figure 83: $S_{21}$ parameter showing the H-plane crosstalk.
Figure 84: $S_{21}$ parameter showing the E-plane crosstalk.
7.3 Tolerance analysis

During the design of both, the transition and the metasurface, we tuned a big amount of parameters regarding dielectric substrates characteristics, vias position and etched metal among others. In this section some of these parameters are studied, focusing on those that we consider as more critical during the manufacturing or assembly process. Figures below show the behaviour of a single transition including the metasurface with an air gap of \( g = 0.1 \text{ mm} \) in terms of reflection coefficient and insertion losses. The studied parameters are: rotational misalignment between upper and lower plate, lateral misalignment between upper and lower plate on X and Y directions, dielectric inaccurate thickness and dielectric inaccurate constant.

7.3.1 Rotational misalignment

During the assembly of both plates a rotational misalignment can be produced if screw holes are not properly placed or in case of a wrong positioning of steering pins. Fig. 85 illustrate an example of this kind of misalignment. Fig. 86 shows how this affect to the reflection coefficient and insertion losses.

![Figure 85: Top view of the single transition with metasurface showing a rotational misalignment of 5°](image)

Figure 85: Top view of the single transition with metasurface showing a rotational misalignment of 5°
Figure 86: $S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1 \text{ mm}$) for several angular rotations.

It can be seen that the Insertion Losses keep almost constant and that the Reflection Coefficient varies slightly but still below low levels (-20 dB). Therefore, rotational misalignment is not critical.

7.3.2 Lateral misalignment

The second type of studied misalignment is lateral misalignment in both, x and y directions. This is illustrated in Fig. 87. Figures 88 and 89 show how translations over x and y axis affect the performance of the transition.
Figure 87: Top view of the single transition with metasurface showing a lateral misalignment along x-axis.

Figure 88: $S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for several translations over x-axis.
Figure 89: $S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1\ mm$) for several translations over $y$-axis.

Lateral misalignment does not produce major changes regarding Insertion Losses. When the misalignment is produced over $y$-axis, a small frequency shift can be seen.

### 7.3.3 Dielectric thickness

A third study is performed regarding the thickness of the substrate layers. For this study we add the thickness of every dielectric layer ($1.589\ mm$) and we consider the most extremes cases studying from the smallest possible thickness to the thickest one. These tolerances are provided by the manufacturer and are expressed as $1.589\ mm \pm 0.200\ mm$
We can observe that this parameter is critical in the behaviour of the structure. This transition is designed to work for a particular cavity height. This height should be $\lambda_g/4$ and when dielectric thickness is changed the performance of the design deteriorates drastically.

### 7.3.4 Dielectric constant

Finally, we study the effect of a variable dielectric constant. Both, core (R-5785(N)) and prepreg (R-5680(N)), are studied and Fig. 91 and 92 show how this variation affects the performance of the single transition.
Figure 91: $S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1 \ mm$) for a varying R-5785 dielectric constant.

Figure 92: $S_{11}$ (continuous line) and $S_{21}$ (dotted line) of the single transition with metasurface ($g = 0.1 \ mm$) for a varying R-5680 dielectric constant.
8 Conclusions and Future Work

In this thesis, a wideband PCB to waveguide transition for 5G (28 GHz band) is designed, as well as, an EBG structure which properly stops the propagation of any EM wave for the required bandwidth.

The transition is built using a cavity backshort and it performs well from 25.9 GHz to 30.2 GHz, having a reflection coefficient smaller than -20 dB and insertion losses around -0.4 dB. This transition is completely embedded in PCB technology, the cavity backshort is filled using the same dielectric substrate (Megtron7) than the one applied to design the stripline, and its walls are replaced by vias.

Regarding the metasurface, a study of a hybrid unit cell has been performed. This is using different materials for each of its plates. Particularly, metal is used in the lower plate and PCB on the upper plate. It has been proved that these structures can mimic the behaviour of already existing unit cells. The designed unit cell performs well from 22 GHz to 31 GHz having a stopband bandwidth bigger than the required for the designed transition. It has also been compared with already existing technologies such as choke flanges and corrugations. The performance of our approach is similar than the corrugation and better than the choke flange but a more cost-effective manufacturing process is used for this new design.

The complete design was manufactured and tested. Measurements are consistent with the simulations but with higher losses, due mainly to an increase in the reflection coefficient. In addition, crosstalk between several transition has been studied (H-plane and E-plane) showing reduced values using the mentioned metasurface.

This project presents several future working lines. Starting for an improvement on the design of the transition trying to maximize the operational bandwidth. Also, a deeper study of the hybrid unit cell can be carried. All design parameters can be characterized by the study of their behaviour. Furthermore, the use of PCB technology involves that different shapes for both etched metal and via position can be shaped. This opens a huge amount of possibilities when designing the unit cell and it can also be deeply investigated.
Another working line is the use of smaller unit cells. As it was shown on the study of the S-parameters of the 1D glide-symmetric unit cell, there exists a first stopband between modes 1 and 2. If the size of the unit cell is reduced, the bandwidth of this first stopband is increased and, at the same time several smaller unit cells can be placed in the same space. This allows us to place more than one ring of smaller unit cells, adding their EBG behaviour to create a good enough stopband.
9 References

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