Abstract— A new design of a dual band all-dielectric transmit array for 5G and satellite communications is presented for establishing a link between ground and satellite terminals. This solution aims at working at satellite ka-band and it is fabricated using 3D printing technology. The goal is to achieve a High Gain above 35dBi while maintaining a good Side Lobe Level with an extremely low cost and simple solution. According to the literature, similar designs have not been reported yet. Using a finite number of 64 dielectric cells, their distribution throughout the transmit array aims at minimizing the error coming from the dual band independent phase correction functions for each band. Two transmit arrays are designed and full wave simulations allowed to confirm fulfillment of specifications. Results show a solution achieving gains of 36.73 dBi at 20 GHz and 37.30 dBi at 30 GHz while maintaining SLL below -19 dB. This design is also compared to the same sized transmit arrays optimized for single band resulting in a difference in gain of -1.20 dBi at 20 GHz and -2.54 dBi at 30 GHz for the dual band case.

Index Terms— All-dielectric, ka-band, 3D-printing, transmit array, satellite communications, dual band.

I. INTRODUCTION

Worldwide, wireless communications have become an enormous part of peoples’ lives. From the regular mobile cellular networks to satellite communications or even broadcast radio, wireless communications enabled societies to create sustainable interconnectivity among people and devices. The development of these technologies created simpler and more effective ways to spread information and data, allowing them to inevitably become an indispensable mean in the rise of new technologies [1].

Satellite communications can play a major role in overall communications as they complement terrestrial networks in reaching users with broader coverage. Studies have correlated the integration of terrestrial and satellite communications for different applications, resulting in technical gains depending on the scenario [2]. Particularly with the emergence of 5G technology, the role of satellites will prove extremely important, since the combination with terrestrial cellular networks will provide an increasingly powerful network, supporting innovation in different industries and services [3]. Additionally, there is a need to shift towards millimeter wave bands in order to achieve higher capacity rates.

Antennas are crucial to the implementation of the requirements for these types of applications, so that free space loss and blockage can be overcome at these frequencies and the radiation characteristics can be maintained. Depending of the type of satellite orbit used for communication, a significantly high gain is required as well as possibly tracking of the satellite’s position in order to maintain a stable link. For this, beam steering mechanisms can be enforced in the antenna. Mechanical beam steering solutions are very simple and low cost but normally very bulky [4][5][12]. An alternative are electronic steering mechanisms which can be attained with the use of phase shifters or phased array antennas, resulting in more expensive and coverage limited solutions [13][14].

Some types of applications do not require the need for tracking mechanisms since the characteristics of the satellite orbit are sufficient to guarantee correct communication.

Geostationary (GEO) orbits require very high gain antennas to counteract the high path loss between ground and satellite terminals. The biggest advantage is that the satellite appears to be fixed to a ground terminal since the period of the orbit is approximately the same as the Earth’s rotation, thus not needing beam steering.

The most commonly used antennas for satellite communications are parabolic reflectors with a horn antenna feed. Depending on size, the gain of these antennas can be very high. However, sometimes they are extremely bulky. Contrary to the parabolic reflector, this solution does not reflect the energy coming from the feed. Instead, it lets the radiation propagate through the structure obtaining the same outcome as parabolic reflectors. The use of reflect arrays and transmit arrays has become of great interest since they can achieve comparable high performance while reducing the overall size of the antenna [4][5][6][7][8][9]. Particularly, a structure design of flat dielectric lenses or transmit arrays is being more frequently used for these types of applications [8][10][11]. Transmit arrays allow for a bigger flexibility regards its placing as a ground terminal, allowing to position them at a wall of a building for example due to the ability to steer the beam or even having a collimated beam with a tilt.

A dual band scenario must also be enforced. By integrating both the uplink and downlink into the same antenna, there is a reduction in the number of antennas needed for communication with the satellite. A similar concept has been introduced in [16], however this solution is designed for lower values of gain and the transmit array’s cells are formed by metallic elements embedded in a multi-layer dielectric substrate. This type of structure has a few fabrication challenges, therefore an even simpler transmit-array structure becomes attractive.

The objective of this article is to design and fabricate an all-dielectric dual band transmit array for GEO applications particularly for the 20GHz and 30GHz bands. This solution aims at being a ground terminal for communication with satellite applications. Looking into previous literature, the design of this all-dielectric characteristic has not yet been reported in the state of the art. Nonetheless, a gain of 35dBi for both frequency bands and a side lobe level (SLL) of less than -15dB are the specifications to achieve in the proposed solution. Furthermore, this type of transmit array can easily be manufactured with a reduced cost by using 3D printing technology at a small scale or by using an injection moulding technology at a large-scale mass production.

The design of the solution must be then compatible with the 3D printer [15] available at Instituto de Telecomunicações (IT) in IST. The extruder nozzle has 0.4 mm diameter, the positioning precision is 12.5 micron, land layer resolution can go down to 40 micron. The dielectric cells infill is 100%.

An alternative fabrication process to 3D printing is injection moulding. This process allows the mass production of these types of transmit arrays at a very low cost. In terms of commercialization this
process would be ideal, however, for testing the 3D printing method is more appropriate.

The proposed prototype is composed of a horn antenna and a transmit array. The main challenge is to design a single transmit array for both frequencies that enforce the correct phase shift at each position for each frequency. The method proposed in this work can be extended to any type of dielectric material and independent of the size of the transmit array. It allows creating an all-dielectric transmit array for dual band applications at the expense of some losses in terms of magnitude of the radiation patterns.

Two different software programs were used in the development of this work: KH3D [17] and CST Microwave Studio [18]. As a first approach to the evaluation of the solution KH3D, it is a very fast way of accessing the potential of the solution. It is based on the hybrid use of Geometrical Optics and Physical Optics. After that, the full wave simulations are performed with CST, which a full-wave solver.

This paper is organized as follows. An analytical formulation for the whole problem is addressed in section II. First the formulation regarding the integration of the unit cells in the transmit array is explained; and after the design challenges relative to the dual band scenario are stated. Section III addresses the design of the dielectric unit cells which will compose the transmit array. In Section IV, the design and simulation of two dual band transmit arrays for different focal distances is detailed. Finally, Section V presents the experimental measurements of the manufactured transmit array.

II. DESIGN OVERVIEW

This section presents the analytical formulation for a collimated beam dual band transmit array. Transmit arrays are planar structure illuminated by a feed located at a distance $F$ from the transmit array. The transmit array is composed of a finite number of unit cells that are responsible for introducing a shift on the phase characteristic of an incident wave in order to compensate the different paths distances across the whole array geometry. With this phase shift, the different existing electric paths between the feed and each and every unit cell integrating the transmit array is compensated, resulting in a collimated beam as seen in figure 1.

A. Unit cells and Transmit array

Figure 1 represents the position of both the feed and transmit array. The centre of the transmit array is the origin of the $(x,y,f)$ coordinate system. Coordinates $x$ and $y$ change according to the position in the array and $z$ is the coordinate that goes from the bottom to the top part of the array.

For the collimation of the beam to occur, every point at the output of the transmit array has to have the same optical phase independently of the varying path lengths from the feed to the input of the transmit array. These varying path lengths, can be represented by a transmit array phase correction function $\phi_{TA}(x,y,f)$, that depends only of the position on the transmit array $(x,y)$ and the operation frequency $f$ [16]:

$$\phi_{TA}(x,y,f) = k_0(x \sin \alpha_0 - \sqrt{x^2 + y^2 + F^2})$$

where $k_0 = 2\pi f / c$ is the free space wave number, $F$ is the distance from the feed to the centre of the array which is a constant and $\alpha_0$ is the zenithal direction arbitrary tilt angle for the collimated beam. When $\alpha_0 = 0$, which corresponds to the broadside Fresnel correction, $\phi_{TA}(x,y,f)$ reduces to the physical length travelled by a ray from the feed phase centre to up to each point in the transmit array.

The transmit array was first set to be a $D \times D$ square with a maximum height of $H$. However, a solution was found to make the transmit array smaller by transforming it into a circle with radius $D$. An agreement between the radiation lost or spill over by cutting the corners of the square and the total size of the transmit array had to be made. Therefore, in figure 2 is shown the transmit array’s final shape which has variable height throughout the array.

Generally speaking, the variation of the focal distance can make impact in two different ways:

- Directivity of the radiation pattern;
- Distortion of the main beam and increase in SLL;

From the work developed on [8], the approximation of the directivity of a transmit array for a diameter of $D_A$ is given by:

$$D_{MAX} = \frac{\eta (\frac{4\pi\sigma}{\lambda})^2 \tanh \left(\frac{D_A^2}{4\sigma}\right) \cos \alpha_0}{\lambda}$$

where $\eta$ defines the aperture efficiency with all the involved losses and $\sigma$ is a parameter related to the edge field taper of the feeding system. With the variation of the focal distance, the edge field taper will vary and change the directivity of the radiation patterns as well as the spill over effect.

The shape of the radiation pattern as well as the SLL typically have better outcomes when the focal distance is larger, however, it also increases the spill over which can lead to increasing side lobes. Balance between these two factors has to be ensured in order to have the best possible outcome when it comes to simulating, testing and measuring the transmit array.
B. Dual Band design challenges

The dual band scenario implies coordination between the designs of the unit cells at both frequencies. When talking about dual band scenario three main factors have to be taken into consideration when designing the transmit array [16]:

- Need to compensate several wavelengths of phase error which results from a very high gain requirement;
- Unfavourable ratio between frequency bands;
- Need to find all the phase pairs for all positions in the transmit array;

In figure 3, a representation of the phase correction function for the transmit array to help visualize the behaviour of equation 1 across the transmit array. By observing this figure, it is clear that there are periodic $2\pi$ rad phase jumps throughout the transmit array. This results from a $2\pi$ rad phase wrapping used to compensate for the higher phase error values. It can be shown that the phase wrapping introduces an operation bandwidth limitation [16].

![Figure 3: Phase correction function (in radians) visualization across the entirety of the transmit array](image)

In the dual band scenario this is particularly difficult due to the independent $2\pi$ rad phase jumps needed at each band. With this in mind it is possible now to represent the phase correction function variation with the position along the transmit array for a wrapped situation. Figure 4 demonstrates exactly how the phase wrapping would behave on a transversal cut for $y = 0$ and $f=600$ mm. An offset value is added to the whole function so that at the center of the x axes both function are zero. It is seen that there are infinite combinations of 20 GHz and 30 GHz phase correction values versus $x$.

![Figure 4: Different wrapped phase correction functions across a transversal cut for $y=0$](image)

III. UNIT CELLS DESIGN

One of the foundation principles of the design and implementation of transmit arrays, is the individual analysis of each component in the array. The transmit array is divided into a finite number of cells which will recreate the phase correction function needed to collimate the beam. Each unit cell has to be individually simulated in order to acquire a full comprehension of its behaviour when integrated within an array. This Section efforts the design and full wave numerical simulation analysis of the group of unit cells which will compose the transmit array.

The unit cells geometric representation can be seen in figure 5. It is composed of a dielectric part and a vacuum part and its shape is a parallelepiped with given thicknesses. These unit cells are designed to present different phase shift values attained from the different dielectric heights each unit cell has. Unlike the work presented in [16], where the unit cells have all the same height and the metallized components create the phase shift required, this solution relies only on the path length travelled through the cell height. This allows a much more simple and low cost solution at the expense of not having all the possible phase combinations at both frequencies.

The base of the parallelepiped is a 5x5 mm square meaning that the width of the cell is $\lambda/3$ and $\lambda/2$ at 20 GHz and 30 GHz respectively. This discretization is determined by the wavelength requirement at both frequencies and the study of the step of the cells determined that there were insignificant differences on the unit cells behaviour until the width of 5 mm.

![Figure 5: Schematic of the two ports and unit cell simulation setup](image)

There are four main characteristics of unit cells that will have to be taken into account when designing and simulating with the final objective to integrate them in a transmit array:

- The phase generated at the cell output port for a given incident wave, considering the cell length, the dielectric characteristics and the required combined behaviour at 20 GHz and 30 GHz;
- The losses from propagation and reflection occurring throughout the unit cells. This will directly affect the magnitude of the signal at the output of the unit cell;
- The thickness of the cells. Not only to avoid having a very large and heavy final product but also because a big discrepancy in heights of adjacent cells can cause additional phenomena to occur, for example reflections.
- The unit cell dual-band design requires an additional concern, there is independent phase behaviour at each band as explained in section II which causes its overall project to be significantly more difficult than a regular single band transmit array.

The purpose of the unit cells is to compensate the electric path between the focal point and each position in the array. A priori, the compensation needed for each position is known from the phase correction function introduced in section II. Therefore, by simulating the unit cells the goal is to obtain the phase behaviour at both frequencies that matches the required phase compensation coming from equation 1.

By approximating the phase behaviour of each unit cell only to the phase shift coming from the propagation in different materials, it is possible to calculate the phase delay introduced by each and every type of unit cell. Of course this approximation does not correspond to the real scenario since it does not take into account the multiple reflections within the dielectric material which affect directly the
phase of the cell. The approximation of the phase delay in an all-dielectric unit cell is given by:

$$\phi_{\text{dielectric}} = Ar(\varepsilon)^{ \phi_{\text{air}} = Ar(\varepsilon)\sqrt{\varepsilon dt_{\text{air}}} = k_0\varepsilon \sqrt{\varepsilon dt_{\text{air}}}$$

where \(k_0 = 2\pi f/c\) is the spatial frequency of the wave or wave number with \(f\) and \(c\) representing the frequency and speed of light in the vacuum respectively, \(\sqrt{\varepsilon}\) is the refractive index of the dielectric material and \(d_{\text{dielectric}}\) is the thickness (in mm) of the dielectric part of the unit cell.

Additionally, the phase of a unit cell composed entirely of air or vacuum can describe in a similar way replacing \(\varepsilon\) by 1:

$$\phi_{\text{air}} = Ar(\varepsilon)^{\phi_{\text{air}} = \sqrt{\varepsilon dt_{\text{air}}} = k_0d_{\text{air}}}$$

The final unit cells will either be all-dielectric or a combination of air and dielectric. This combination with different values for the height of each material will allow creating different phase shifts according to the need. This results in variable thickness unit cells of dielectric which can be defined by using equations 3 and 4.

$$\phi_{\text{dielectric-air}} = k_0\varepsilon \sqrt{d_{\text{dielectric}}} + k_0d_{\text{air}}$$

With equation 5 it is possible to calculate the height or thickness of any cell according to the phase shift wanted at each position. The height of a specific cell in relation to a reference cell chosen arbitrarily is given by a combination of two equations 5 and solving the equation to find \(d_{\text{dielectric}}\):

$$d_{\text{dielectric}} = \frac{(\phi_1-\phi_{\text{ref}})+k_0\varepsilon dt_{\text{ref}}+k_0d_{\text{ref}}}{k_0\varepsilon \sqrt{\varepsilon d_{\text{air}}-k_0}}$$

It is now possible to calculate the value for the height of each unit cell related to the phase shift needed, \(\phi_1 - \phi_{\text{ref}}\), and then with the CST simulations, confirm it behaves the way it is supposed to.

![Diagram](image.png)

Table 1: Calculations of the height of dielectric correspondent to each phase shift based on equation 6

<table>
<thead>
<tr>
<th>Phase shift</th>
<th>Height calculated at 20GHz [mm]</th>
<th>Height calculated at 30GHz [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>ref</td>
<td>0.27</td>
<td>0.27</td>
</tr>
<tr>
<td>45°</td>
<td>2.94</td>
<td>2.05</td>
</tr>
<tr>
<td>90°</td>
<td>5.60</td>
<td>3.83</td>
</tr>
<tr>
<td>135°</td>
<td>8.27</td>
<td>5.60</td>
</tr>
<tr>
<td>180°</td>
<td>10.94</td>
<td>7.38</td>
</tr>
<tr>
<td>225°</td>
<td>13.61</td>
<td>9.16</td>
</tr>
<tr>
<td>270°</td>
<td>16.27</td>
<td>10.94</td>
</tr>
<tr>
<td>315°</td>
<td>18.94</td>
<td>12.72</td>
</tr>
<tr>
<td>360°</td>
<td>21.61</td>
<td>14.50</td>
</tr>
</tbody>
</table>

The reference cell was chosen to be the smallest possible dielectric thickness that could be printed by the 3D printer. This resulted in a \(d_{\text{ref}} = 0.27\text{ mm}\). By replacing, in equation 6, the aforementioned values, it is possible to create a table with the results at each frequency. Table 1 shows a linear relation between the height of the cell and the phase shift the cell is characterized by. As can be seen in table 1, the two frequencies have a different height evolution according to the phase shift value. This particular effect is the main problem of the dual-band scenario because the phase of the cells at 20 GHz and 30 GHz vary independently as explained in section II.

For the purpose of this work, the dielectric originally chosen was Polylactic Acid (PLA), which is a biodegradable plastic from the polymers family and has a permittivity of 2.9. The choosing of this dielectric material was also favourable because its characteristics where very well defined already in the IT laboratory environment. With this in mind, table 1 demonstrates the evolution of the heights for the unit cells in increments of 45° phase shifts when compared to the reference cell.

In terms of outdoor applications, since the PLA is a biodegradable plastic it would not be fit to integrate the transmit array, however, in terms of testing it is very appealing for reasons aforementioned. Another plastic such as acrylonitrile butadiene styrene (ABS) would be more fit since it is not biodegradable and it is very commonly used in moulding processes. The 3D printer available in laboratory does not work well with this material that is why the chosen material was the PLA.

CST simulations were performed for a minimum height of 0.27 mm and maximum height of 21.375 mm using a step of 0.335 mm which will allow minimizing very big height differences between adjacent cells. This resulted in 64 evenly distributed unit cells with a phase shift between each other of around 5º/6º at 20 GHz and 8º/9º at 30 GHz. Two ports were placed according to figure 5 to simulate an electromagnetic wave transmission and allow the computation of the parameters responsible for determining the magnitude and phase response of the unit cell using periodic boundary conditions.

With the CST simulations, it was confirmed that the mathematical calculations performed above and presented in table 1 where approximately correct with small insignificant phase discrepancies. To have a better representation of the phase behaviour of each unit cell designed at both frequencies, a representation is made in figure 6 showing for every unit cell the pair of phases which correspond to its simulation results.

This representation was first introduced in [16] for a dual band transmit array for different applications. In this case, the phase of each unit cell at both frequencies is treated like a Cartesian \((x, y)\) pair, therefore, a representation of the 64 unit cells can be done by assuming each unit cell, or each pair, is denoted by one cross in the representation. The lines in figure 6 correspond to the areas of unrepeated unit cells. In order to have all required pairs of phase combinations at 20 and 30 GHz, the lines have to be fully populated with the unit cells simulated. This means that a dual band transmit array can be populated with a finite number of discrete phase cells. Figure 6 is our basic groundwork for the population of the transmit array and, for this reason, every unit cell designed must comply with the sloped lines in the graph.

Another very important parameter is the transmission magnitude coefficient (TMC) as explained in the beginning of this section. This coefficient translates the overall magnitude transmission loss that each cell imposes. In order to better analyse this coefficient, figure 7 was produced, showing a graph of all the 64 unit cells simulated and
the transmission magnitude coefficient evolution regarding the phase shift of every unit cell relative to the reference cell defined earlier.

Figure 7: TMC behaviour evolution with the increase in the phase shift value relative to a reference cell

It is clear that there is a relation between the TMC and the phase shift relative to the reference cell. As the thickness of the cell gets bigger, the phase shift gets bigger, and the overall coefficient increases, this translates into two oscillating behaviour with an offset and a slope.

In the application proposed in this work, due to the unit cells being defined by variable thickness dielectric material, there are not enough degrees of freedom to impose simultaneously a phase and magnitude condition. Therefore, the phase condition was given priority in relation to the magnitude condition. This will allow to have a better collimation of the beam at the expense of a higher reflection loss related to the unit cells used in the transmit array. With that being said, all the 64 unit cells found will be considered for implementation on the transmit array.

However, figure 6 shows that not all lines needed for correct assignment of the pairs of phase combinations where found. This means that in certain areas of the transmit array, where the pairs of phase combinations not found are needed, there will be a certain degree of error in the collimation of the beam.

If the TMC is a specific requirement or the overall size of the unit cells needs no be reduced, then a combination of dielectric and metal cells should be studied and employed like shown in [16].

IV. TRANSMIT ARRAY DESIGN AND SIMULATIONS

In this section, two transmit arrays, using the cells developed in section III are designed and full wave simulated in order to assess their viability when it comes to complying with the requirements. The two transmit arrays differ on the distance in the focal distance.

The targeted directivity is of 35 dBi considering a typical edge taper illumination of around 7-11 dB and a typical aperture efficiency of 50%. Replacing these values in equation 2 and solving it in order of $D_x$, the diameter value for both frequencies is between 550 mm and 800 mm. The chosen diameter was 600 mm since it is in the lower part of the interval mentioned before and it allows for a smaller dimension solution.

There exists a trade-off between the value of $F$, the phase error to compensate in the transmit array due to the missing pairs of phase combinations and the feed edge taper. As $F$ increases, the phase error decreases since the phase jumps become less frequent; however, the feed edge taper gets worse increasing the spill over which can increase the SLL and decrease the gain. The variation of $F$ will have an impact on the outcome of the solution and for this reason two focal distances are chosen for simulation, 500 mm and 600 mm.

Transmit array design is a more complex work than the unit cells themselves, especially in the dual band scenario. To have a transmit array prepared for transmission and reception at both frequencies, an automated simulation process must be applied to ensure the final product has the best possible outcome. The goal is to populate the transmit array with a finite number of discrete phase unit cells so that the phase correction function of the resulting transmit array is the closest possible to the continuous function.

For this, a horn antenna is integrated with the transmit array in the CST software to create the final prototype antenna.

A. Transmit array for focal distance of 600 mm

In order to optimize the choosing of the cells for each position in the array, the process of populating was done manually. What this means is that, according to the distance from each position in the transmit array to the centre of the transmit array, a specific cell was selected which would ensure the fewer errors in phase as possible, in 20 GHz and 30 GHz. By manually choosing the cells, the optimization of the distribution of the unit cells throughout the array has the lower error it can be obtained.

With the cells at disposal, it is not possible to guarantee an error free solution as explained in Section III. However, when choosing the cells for the most critical cases, the criteria chosen had the purpose of balance the error at both frequencies as represented in figure 8.

It is clearly visible that there are two types of areas within the different rings across the transmit array. The area in blue corresponds to the use of all cells developed in Section III which ensure a 360° phase wrap at 20 GHz and a 540° phase wrap at 30 GHz, meaning that combinations between the two phases are existent. After the first transition, the yellow and green regions correspond to cells with non-existing perfect combinations of phase at 20 GHz and 30 GHz. Meaning that these combinations of phases could not be syntheised with the used cell geometry, therefore, existing cells were used trying to optimize the solution at both frequencies according to the manual criteria stated before. That’s why a clear jump between error occurs as the position gets further away from the centre of the transmit array. A mean error of 40.9° and 34.8° at 20 GHz and 30 GHz respectively was calculated by relating the phase function correction with the phase imposed by the chosen unit cell at each position in the array.

Figure 9: Front and side view of the designed transmit array

Figure 9 presents the CST simulation model, showing the transmit array and the horn feed. The feed edge taper illumination is around 7 dB for this focal distance. Two plane symmetries are applied since the transmit array is symmetrical in both x and y axes.
Since there are no similar transmit arrays in the literature to compare the results with, a good comparison of the performance of the designed transmit array is to relate the dual band with single band performances. This means that single band transmit arrays were optimized for each frequency using the same unit cells developed in section III. This comparison will allow to understand how much is the loss that this work’s solution because of its dual band characteristics. The criteria from which the cells were chosen was as follows: at each position evaluate which unit cell from the library available corrected the phase better according to the phase correction function. By doing this, the phase imposed by each cell throughout the array for each frequency almost resembles a perfect phase correction function. Obviously, this scenario cannot be perfectly replicated since the phase values imposed by the library of unit cells available are discrete.

The simulated far field realized gain for the E and H planes for the two different frequencies, 20 GHz and 30 GHz respectively, along with the single band simulations are presented in figure 10 and 11.

![Figure 10: Simulated far field realized gain radiation pattern at 20GHz for both planes](image)

![Figure 11: Simulated far field realized gain radiation pattern at 30GHz for both planes](image)

Analysing the two figures above it is possible to say that both the E and H plane behaviour is very similar differing slightly in the side lobes. There is a bigger deformation of the radiation pattern between the two scenarios at 30 GHz than at 20 GHz, this shows that the phase error across the array is affecting one frequency more than the other. The realized gain for the dual band solution at 20 GHz and 30 GHz frequencies is 36.7 dBi and 37.3 dBi respectively. Additionally, the SLL for both frequencies is very low and below -19 dB. The main lobe is directed into the required angle which is 0°. The radiation efficiency is of 30.31% and 15.12% for the 20 GHz and 30 GHz frequencies respectively. A decrease in the radiation efficiency can also be observed from the single to dual band scenario, this is one of the consequences of dual band design as stated before. A table is done in order to compare the simulations results between the dual band and single band scenarios; RG corresponds to the realized gain, SLL to the side lobe level and RE to the radiation efficiency.

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Frequency [GHz]</th>
<th>Plane</th>
<th>RG [dBi]</th>
<th>SLL [dB]</th>
<th>RE [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Single Band and F= 600 mm</td>
<td>20</td>
<td>E</td>
<td>37.93</td>
<td>-27.70</td>
<td>39.32</td>
</tr>
<tr>
<td></td>
<td>30</td>
<td>E</td>
<td>39.84</td>
<td>-31.20</td>
<td>27.13</td>
</tr>
<tr>
<td></td>
<td>30</td>
<td>H</td>
<td>36.73</td>
<td>-21.80</td>
<td>30.31</td>
</tr>
<tr>
<td>Dual Band and F= 600 mm</td>
<td>20</td>
<td>E</td>
<td>37.30</td>
<td>-19.00</td>
<td>15.12</td>
</tr>
<tr>
<td></td>
<td>30</td>
<td>H</td>
<td>-19.00</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

From table 2 there are also some differences as expected. The realized gain of the dual band transmit array reduces 1.20 dB and 2.54 dB at 20 GHz and 30 GHz respectively from the single band transmit arrays. A bigger loss at the higher frequency is encountered, however the losses in relation with an all-dielectric single band transmit array are very small, especially when taking into consideration the practicality of having only one transmit array performing both the uplink and downlink. This reinforces the need to fabricate and measure an all-dielectric dual band transmit array.

B. Transmit array for focal distance of 500 mm

The process of choosing the cells for this section was similar to the anterior section. All the same steps were taken when it came to populating the transmit array with specific unit cells, the only difference was on the distance from the feed to the transmit array, having to translate the feed 100 mm in the direction of the transmit array. This different F testing will allow to understand the differences in performance of the transmit array when changing the focal distance. Therefore, the representation on the CST is the same as for image 9 but the only difference is the focal distance which decreases by 100mm and the geometry of the transmit array is completely new.

In terms of the representation of the error of the chosen cells relative to the theoretical phase correction function as shown for the previous section in figure 8, the results for this array are very similar. The mean error was of 42.4° and 35.5° at 20GHz and 30GHz respectively, calculated by relating the phase function correction with the phase imposed by the chosen unit cell at each position in the array. These mean errors are a little bit bigger than for the previous section for reasons explained in the beginning of this section. The simulated far field realized gain for the E and H planes for the two different frequencies, 20GHz and 30GHz respectively, are represented in figure 12 and 13.

Analysing the two figures above it is possible to say that both the E and H plane behaviour is very similar differing slightly in the side lobes. Like in the previous case, the deformation of the far field radiation pattern is bigger at 30 GHz than at 20 GHz. The realized gain for the 20 GHz and 30 GHz frequencies is 36.3 dBi and 36.6 dBi respectively for the case of 500 mm of focal distance. Additionally, the SLL for both frequencies is very low and below -19 dB. The main lobe is directed into the required angle which is 0°. The radiation efficiency is of 27.01% and 12.86% for the 20 GHz and 30 GHz frequencies respectively. A decrease in the radiation efficiency can be observed from the 600 mm to 500 mm of focal distance as a consequence of a decrease in the realized gain of 0.43 dB.
dBi and 0.74 dBi at 20 GHz and 30 GHz respectively. Table 4 summarizes the results.

Table 4: Overall simulation results comparing both scenarios for the different focal distances

<table>
<thead>
<tr>
<th>Scenario</th>
<th>Frequency [GHz]</th>
<th>Plane</th>
<th>RG [dBi]</th>
<th>SLL [dB]</th>
<th>RE [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>F=600</td>
<td>20GHz</td>
<td>E</td>
<td>36.73</td>
<td>-21.80</td>
<td>30.31</td>
</tr>
<tr>
<td></td>
<td></td>
<td>H</td>
<td></td>
<td>-21.80</td>
<td>30.31</td>
</tr>
<tr>
<td></td>
<td>30GHz</td>
<td>E</td>
<td>37.30</td>
<td>-19.00</td>
<td>15.12</td>
</tr>
<tr>
<td></td>
<td></td>
<td>H</td>
<td></td>
<td>-19.00</td>
<td>15.12</td>
</tr>
<tr>
<td>F=500</td>
<td>20GHz</td>
<td>E</td>
<td>36.30</td>
<td>-21.30</td>
<td>27.01</td>
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<td>H</td>
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<td>-21.10</td>
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<td></td>
<td>30GHz</td>
<td>E</td>
<td>36.56</td>
<td>-19.50</td>
<td>12.86</td>
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<tr>
<td></td>
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<td>H</td>
<td></td>
<td>-19.10</td>
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It can be seen that both the realized gain and radiation efficiency are better for the case of 600 mm. In terms SLL, all values for both frequencies are good enough for them to be insignificant in terms of alterations on the correct signal processing. The reason of choice of which transmit array to fabricate is decided strictly by performance simulation results because the antenna height is not such an important factor in these types of applications. Therefore, in the spectrum of this thesis, the transmit array with focal distance of 600 mm is the one selected for the fabrication and measurements processes.

C. Central plate simulations

It is not possible with currently available FDM 3D printers to print the 60 cm diameter transmit-array as a single block. It was divided into nine parts with square projected area of 200 mm x 200 mm as shown in figure 14. The elapsed time for printing each block was in the order of 50 hours. Long duration prints of 100% fill rate blocks such as these, tend to detach from the printer bed in the process, and create deformations.

On the other hand, the existing antenna measurement infrastructure at IT is not equipped for direct far-field measurement of a 60 wavelength aperture antennas. The measurement distance is 1.6 m. Therefore a decision was taken to print and measure the radiation pattern of the central part of the TA only, keeping the focal distance at 600 mm. This experiment is enough to validate the full-wave simulation model, which we postulate then that provides the correct result for the complete 600 mm diameter antenna simulation.

Additional simulations using the CST software were carried out so that the measured results in Section V can be compared with simulations in the same conditions. With this in mind, this section aims at performing the simulation of the structural part in the same scenario of the previous sections along with some adjustments that will be further explained.

The simulations presented below are done taking into considerations all the challenges and structures presented in section V. Additionally, all the supporting material presented in section V b) is considered in the simulations so that the simulation scenario is the closest possible to the real measurements’ scenario. A block of Styrofoam was used to keep the transmit array in place, and a block of absorbing material was integrated surrounding the transmit array so that the spill over radiations would affect the least possible the measurements. The resulting CST representation is presented in figure 15.

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Analysing the two figures above it is possible to say that both the E and H plane behaviour is very similar differing slightly in the side lobes. Because this scenario recreates the scenario used for the measurements, these will be the simulations used to compare to the measured results.

Figure 16: Near field calculated in the E and H plane for 20 GHz at a distance of 1.6 m

Figure 17: Near field calculated in the E and H plane for 30 GHz at a distance of 1.6 m

V. EXPERIMENTAL RESULTS

In this section, the validation of one of the transmit arrays designed and simulated in section IV is performed. First a brief description of the prototype definition and fabrication is made. After that, the supporting structures that were fabricated are presented. Finally, the measured results are demonstrated.

A. Challenges

The main focus of the fabrication is the transmit array, as stated before, this process is made using a 3D printer available in our laboratory, model Ultimaker 2 Extended+ [15]. The specifications of this individual 3D printer constrain the dimension of the size of the plates to be printed. As stated before, the transmit array has to be divided into sub-plates that will after be put together to recreate the final prototype.

Additionally, due to the total time that it takes for printing, the extremities of the printed lens can start to curl and go out of place. On the fabrications of the central plate, one of the corners of the plate started to curl and end up staying a little deformed in relation to the original design as it is shown in figure 18.

This deformation can cause problems when simulating the realized gain for the central plate, therefore, it had to be taken into considerations in the simulations in section IV C).

Finally, regarding the measurements of the radiation pattern of the transmit array, a challenge with the available installations arises due to the size of the transmit array. The far field distance for the fabricated central plate is much bigger than the dimensions of the anechoic chamber available for measurements. With this restriction in terms of space, the measurements of the radiating field coming from the dielectric lens need to be performed at a smaller distance than the far field distance. This distance is the largest that can be attained inside the anechoic chamber and it is defined at 1.6 m away from the transmit array.

B. Supporting Structures

In parallel to the fabrication of the transmit-array, a supporting structure was built in order to hold the primary feed and the transmit array in the correct positions to be used during the measurements. This supporting structure guarantees that the primary feed is placed relatively to the transmit-array according to setup described in section IV a).

The anechoic chamber available for measurements at the laboratory has a pre-defined structure in place which has to be adapted to the scenario of the work of this thesis. The transmit array is fixed in its position with the help of a Styrofoam block with a small square compartment. In order to hold the Styrofoam block in its place, two wooden pieces are put together and integrated in the supporting structure already existing in laboratory. Figure 19 describes all this setup and additional structures.

Figure 18: Perspective and side view of the central plate focusing on the deformed corner

The structure used for simulation also integrated a block of an absorbent material so that the exceeding taper coming from the feed influenced the least possible in the measurements. This absorbent material was fixed in its position using a small 3D printed shelf. The absorbing material used was ECCOSORB AN-77 [19] which was available in the laboratory and also present in the library of material in CST for the simulations.

As it is possible to see in figure 20, the transmit array is inserted in a cavity in the Styrofoam which then is covered by the absorbing...
material with a window for the transmit array. The structure was successfully manufactured and all dimensions were identical to the CST design.

![Figure 20](image1.png)

**Figure 20:** Representation of the fabricated prototype and supporting structures for measurements

C. Measured Results

The fabrication and assembly of all the parts composing the antenna was successful. According to figure 21, all the parts from the antenna parts to the supporting structure were put together in the anechoic chamber available in laboratory. After setting up all the components, the structure ready for measurements had the display shown in figure 21.

![Figure 21](image2.png)

**Figure 21:** Perspective and front view of the manufactured central plate integrated with all the structures and the horn feed

The measuring system is represented in figure 22 below.

![Figure 22](image3.png)

**Figure 22:** Representation of the setup used in the laboratory for the measurements

Figures 23 and 24 presents the normalized radiation pattern superimposed on simulations for the aforementioned scenario of measurements for 20 GHz and 30GHz respectively. The measured results for the frequency for the 30 GHz band were made at a frequency of 29.45 GHz, therefore, the measured results have bigger deformations and magnitude losses than if they were made at 30 GHz.

For 20 GHz the agreement between both diagrams seems reasonably good, except that the measured realized gain is 1.06 dB below the simulated value. However, there is a bigger difference at 30 GHz of around 3.28 dBi.

![Figure 23](image4.png)

**Figure 23:** Measured and simulated radiation patterns at 20 GHz

![Figure 24](image5.png)

**Figure 24:** Measured and simulated radiation patterns at 20 GHz

![Figure 25](image6.png)

**Figure 25:** Measured cross polarization for 20GHz for both planes

This difference can be mainly explained by the difference in frequency at which the simulation and measurements are made. The simulation is considering a feed at 30GHz and the measurements at 29.4 5GHz. Additionally, the cross polarization is kept at very low levels for both frequencies as represented in figure 25. The cross polarization values are normalized identically to the radiation patterns and below -18 dB for both frequencies.

VI. Conclusions

The main goal of this thesis was to design, simulate, manufacture and measure an experimental fully 3D printed prototype consisting of an all-dielectric dual band transmit array and a primary horn feed. The 3D printed transmit array advanced in relation to the existing
literature by being all-dielectric and having the ability to communicate in two different frequency bands, namely the 20GHz and 30GHz bands resulting in an extremely low cost solution. The horn feeds were already studied and available in the IT laboratory.

In order to reduce the geometry complexity and overall cost of the alternatives already existent in the literature for communicating with GEO satellites, an all-dielectric transmit array populated with a set of 64 unit cells designed was designed.

The difficulty of the dual band scenario is the independent phase wrapping needed at each band which can create a very large number of phase pair’s combinations at both frequencies. By using a finite set of dielectric cells not all the phase pairs needed are available, however, the cells can be distributed so that the error across the whole transmit array is minimized. Furthermore, the dielectric compatible with the 3D printer available in the laboratory was PLA which has the disadvantage of having very high losses.

Two transmit array with different focal distance designs were simulated using full wave simulations in CST. The results obtained from both these transmit arrays were compliant with the specifications defined at the beginning of this work achieving a gain of over 35 dB at both frequencies, as well as, very low SLL of under -19 dB. Of the two transmit arrays, the best performing one was chosen to the manufacturing process. This array was able to achieve a realized gain of 36.73 dB at 20 GHz and 37.30 dB at 30 GHz while maintaining SLL below -19 dB at both frequencies. Measurements of one of the plates of the array showed good results especially for the 20 GHz frequency. The measurements of the full transmit array should be made to confirm the simulations, especially with the promising scenario coming from the measurements.

By looking into the available literature, an all-dielectric dual band 3D printed transmit array was not found. The dual band scenario allows optimizing a very important factor when it comes to communications which is space availability and cost.

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References


