Design and Simulation of a Slot Waveguide Array Antenna (SWAA) for Satellite Communications

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Abstract. One of the most used antennas today, is that of slots in waveguides. These antennas are an important type within the microwave range with various applications in electronic communications systems, radar, and satellite applications, due to its high power capacity, high efficiency, low weight, compact structure and few lateral lobes, all added together to its proven resistance to radiation, high temperatures and vibration, characteristics without which they can not be used in aerospace applications. The following lines show the design and simulation of a Slot Waveguide Array Antenna (SWAA) for satellite communications in the 30 GHz band. The antenna will be approximately 170x170 mm, constituted by 24x24 slots. It will be fed by waveguides in a lower layer connected to the radiating guides through slots. The modeling and design of the different elements that make up the power grid as T-junctions and elbows in the E and H planes were carried out.

1. Introduction
At present, different types of arrays have been developed [1, 2, 3, 4], all with some restrictions, mainly in terms of efficiency and bandwidth [5]. The increase in operating bandwidth of slit antennae in waveguides is currently stimulated by the development and applications of synthetic aperture radars (SAR) [5, 6, 7]. According to Takeshima and Isogai [8], the larger the number of slots in an array, the lower its bandwidth. Therefore in [9,10] it was found that by reducing the number of slots in an array and using subarrays fed by waveguides the bandwidth can increase. In [11], it was found that by managing conventional longitudinal slots and subarray techniques it is possible to achieve bandwidths of approximately 15%. In [12] to extend the bandwidth of the array, it went away in two subarrays fed by a convex waveguide power splitter. In [13] a geometry of slots for large arrays that offer an increase in the bandwidth of the slot on the conventional geometry is exhibited, this new model was compared with the conventional one and its advantages and disadvantages were discussed in large arrays applications of slots. While most researches are based on narrow grooves, the coupling between grooves is very important, since this coupling can optimize bandwidth and power handling capacity, Sinha [14] and Sangster [15] have investigated the coupling of wide grooves. Distribution networks constructed with various power dividers are widely used to power the array's radiating elements with the correct amplitudes and phases [16, 17, 18]. They have the advantage of having low losses, high power handling and are used to feed arrays of microstrip antennas [19] and slots [20], but their use has been restricted by having large relative electrical dimensions [21].
The aim is to design and simulate a Slot Waveguide Array Antenna (SWAA) for satellite communications, which will operate in the 29.5-31 GHz band. The power coefficients will be separable in such a way that the use of linear current distributions to independently generate the radiation pattern with different beam width values in the main planes of the antenna. A distribution of currents associated with a Taylor diagram of 20 dB will be used. The antenna will be 170x170 mm, consisting of 24x24 slots and taking into account that the separation between slots is 7,031 mm, 48 linear subarrays of 12 slots each will be required, for a total of 576 slots. The way of feeding has been raised by waveguides in a lower layer coupled to the radiating guides by means of slot couplings.

In the following lines, we seek to develop a waveguide slot antenna with vertical polarization and broadside radiation, to operate in the 29.5-31 GHz band at an operating frequency of 30 GHz, with a gain greater than 30 dBi, a beam width of 3 degrees in both planes, a fractional bandwidth of 5%, a Lobe Level (NLPS) of 20 dB and an ROE less than 1.4: 1. A Slot Waveguide Array Antenna (SWAA) was selected. A distribution of currents associated with a Taylor diagram of 20 dB will be used in this design. An efficiency greater than 55% is expected.

2. Approach of the solution (design of the complete array) and results.

2.1. Antenna size
To determine the size of the antenna, the antenna beam width specifications are evaluated, as well as their gain and efficiency, taking the least restrictive selection criteria.

2.2. Beam width (BW)
Taking into account that a beam width (BW) of 3° (0.05236 rad) is required in both planes, we can determine approximately the length in the planes of the array, using the equation:

\[ L = f_B \lambda \frac{0.88}{BW} \]

Where \( \lambda \) is the wavelength in the vacuum and \( f_B \) is the spreading factor of a Taylor array given by:

\[ f_B = 1 + 0.632\left(\frac{2}{R}\cosh\sqrt{(\cosh^{-1}R)^2 + \pi^2)}\right) \]  

For a lobe level of 20dB, \( R = 10 \) and the spreading factor is practically unity. For the frequency of operation the wavelength is 10 mm, therefore.

\[ L = (10 \text{mm}) \left(\frac{0.88}{0.05236 \text{rad}}\right) = 168.1 \text{mm} \]

An antenna array of 170 x 170 mm is required.

2.3. Gain and Efficiency
Taking into account that at least one gain (G) of 30 dBi (1000) and an efficiency (\( \eta \)) of 55% is required, we can determine approximately the length in the planes of the array, using:

\[ A = G \frac{r^2}{4\pi \eta} \]

\[ A = 1000 \left(\frac{(10 \text{mm})^2}{4\pi(0.55)}\right) = 14.468.6 \text{mm}^2 \]

An array of at least 120 x 120 mm is required. Taking these results into account, the beam width criterion is taken as a reference, given that it has a higher value. Therefore the size of the antenna
would be at least 170x170 mm. And taking into account that the separation between slots is 7.031 mm, 
\((\lambda g/2)\) the antenna will be constituted by 24x24 slots, for a total of 576 slots.

2.4. Design subarray
This was based on the specifications of the antenna. The criterion established to define the maximum number of slots in a subarray was to have a bandwidth from 29.5 to 31 GHz, with a fractional bandwidth of 5%. In the first stage of the design a formula was found that determines the maximum number of slots that are required to comply with the bandwidth specification, which will allow us to determine the number of slots of a subarray.

Taking into account that it is desired to operate in the frequency band of 29.5 - 31 GHz, a fractional bandwidth of 5% is required, therefore, using the equation:

\[ \Delta f / f < \frac{1}{2m} \]

Where \( \Delta f / f \) is the fractional bandwidth, confirming that increasing the length of the antenna reduces the operating bandwidth.

We can determine that the separation between the first and last radiation slots of the subarray, must be less than 5 and therefore each subarray, will consist of six linear arrays, each of 12 radiation slots. Therefore, a total of 8 subarrays divided into four quadrants are required, for a total of 48 linear arrays of 12 slots each.

2.5. Generation of excitation coefficients and displacement of the array slots \((x_n)\)
For the generation of the coefficients the Taylor method with \( n = 4 \) is used and for obtaining the displacements of the slots of the array they were based on their electrical equivalent. The resonant arrays are designed for broadside operation and to obtain said radiation all slots must be excited in phase, this is achieved by spacing the slots \( \lambda g/2 \) alternating their location around the central axis of the guide, as shown in figure 1.

![Figure 1. Longitudinal resonant arrays and their electrical equivalent [22]](image)

The equivalent circuit of the array consists of \( N \) conductances connected through a transmission line with spacings of \( \lambda g/2 \), a short circuit is located at a distance \( \lambda g/4 \) or \( 3\lambda g/4 \), from the last slot, for minimize the reflected wave and this produces that the equivalent input conductance of the array \( g_e \) is simply the sum of all the individual conductances; Then, you have to:

\[ g_e = \sum_{n=1}^{N} g_n \quad (2) \]

If it is the desired excitation coefficient for the e-nth slot in the array. It can be shown that the relative excitation coefficient of said e-nth slot is proportional to \( \sqrt{g_n} \) and therefore, it would be given by:

\[ g_n = Ka_n^2. \]
Where $K$ is a constant to be determined and to radiate all the available input power, it must be fulfilled that the equivalent input conductance of the array $g_e$ equals 1, having that: $K \sum_{n=1}^{N} a_n^2 = 1$.

Therefore, the constant $K$ is determined, knowing that the excitation coefficients are already known. To later use the previous equation and define $g_n$ and obtain the individual conductances of each slot and finally, using the equation that equals 1, the equivalent input conductance of the array $g_e$, determine the displacement $x_1$ of said slots. The slots are designed to be resonant in their frequency of operation, for this reason they are of lengths close to a $\lambda/2$, allowing to maximize the bandwidth of the array [23].

Table 1 shows in its second and fourth column the coefficients for the 24 slots ($N = 24$) of the first row of the array for an NLPS of -20 dB, and in its third and fifth columns, the respective displacements of the slots with respect to the axis of the guide.

| Ranuras | (mm) | Ranuras | (mm) |
|---------|------|---------|------|
| 1       | 0.8347 | 13      | 1.67798493 | 0.9665 |
| 2       | 0.99926318 | 0.8340 | 14      | 1.6618726 | 0.9566 |
| 3       | 1.00855231 | 0.8422 | 15      | 1.62339987 | 0.9331 |
| 4       | 1.04291921 | 0.8723 | 16      | 1.55533467 | 0.8918 |
| 5       | 1.11241761 | 0.9339 | 17      | 1.45665268 | 0.8324 |
| 6       | 1.21522678 | 1.0263 | 18      | 1.33699436 | 0.7612 |
| 7       | 1.33699436 | 0.7612 | 19      | 1.21522678 | 1.0263 |
| 8       | 1.45665268 | 0.8324 | 20      | 1.11241761 | 0.9339 |
| 9       | 1.55533467 | 0.8918 | 21      | 1.04291921 | 0.8723 |
| 10      | 1.62339987 | 0.9331 | 22      | 1.00855231 | 0.8422 |
| 11      | 1.66187260 | 0.9566 | 23      | 0.9926318  | 0.8340 |
| 12      | 1.67798493 | 0.9665 | 24      | 1         | 0.8347 |

**Table 1.** Normalized excitation coefficients and slots displacements

### 2.6. Theoretical Radiation Diagram

The figure 2 shows the array factor of the electric field radiated by a rectangular flat grouping of 24x24 elements equispaced to $\lambda g/2$, which is given by:

$$ FA(\theta, \phi) = \sum_{m=1}^{M} \sum_{n=1}^{N} a_{mn} e^{jmk \frac{\lambda g}{2} \text{sen} \theta \text{cos} \phi} e^{jnk \frac{\lambda g}{2} \text{sen} \theta \text{sen} \phi} $$

(3)
Where are the excitation coefficients, that for the separable feed \( (a_{mn} = a_m a_n) \), \( k \) is the wave number of the free space and \( \lambda_g \) is the wavelength in the guide.

![Figure 2](image)

**Figure 2.** Array factor of a rectangular flat array of 24x24 elements

In figure 3, it is shown the total electric field radiated by the array, which includes the radiation of the elements, which in this case is a slot of \( \lambda / 2 \).

![Figure 3](image)

**Figure 3.** Total electric field radiated by the array, includes the radiation of the elements \( (\theta=90^\circ) \)

2.7. **Power Distribution in the Array**

The power input to the network of the array is equally distributed to the waveguides of the four sectors of the array (Top right, Top left, Bottom right and bottom left), using a corporate type power network and from each sector is distributed the power using power dividers type T and elbows H and E to each subarray, which are those that contain the radiation slots according to the chosen excitation law, which in our case follows a Taylor distribution.

2.8. **Power Distribution in the Subarray (Parameters S)**

The input power of each subarray can be estimated as the sum of the powers of each linear array and taking into account that the power of each linear array is proportional to the square of the excitation coefficients of the law, in the following way

\[ P_1 \sim b_1^2; \quad P_2 \sim b_2^2; \quad P_3 \sim b_3^2; \ldots; \quad P_n \sim b_n^2 \]

Where \( b_1, b_2, b_3, \ldots, b_n \) are the excitation coefficients. Therefore, the total input power that enters to the three linear arrays is: Total power to the subarray; \( PT = P_1 + P_2 + P_3 = b_1^2 + b_2^2 + b_3^2 \)
a. For the second linear array: If we consider that in the first linear array there is a unit power.

\[ S_{21}^2 = \frac{P_1}{P_1 + P_2} = \frac{b_1^2}{b_1^2 + b_2^2} \]  \hspace{1cm} (4)

\[ S_{21}^{dB} = 10 \log\left(\frac{P_1}{P_1 + P_2}\right) = 10 \log\left(\frac{b_1^2}{b_1^2 + b_2^2}\right) = 10 \log\left(\frac{0.999^2}{1^2 + 0.999^2}\right) = -3 dB \]

b. For the third linear array:

\[ S_{21}^2 = \frac{P_1 + P_2}{P_1 + P_2 + P_3} = \frac{b_1^2 + b_2^2}{b_1^2 + b_2^2 + b_3^2} \]  \hspace{1cm} (5)

\[ S_{21}^2 = 10 \log\left(\frac{1^2 + 0.999^2}{1^2 + 0.999^2 + 1.009^2}\right) = -1.79 dB \]

Based on these calculations and the use of the CST Microwave Studio software, the displacements of each coupling slot for each subarray are determined.

2.9. Determination of the phases of the radiating grooves and S11

Taking into account that a broadside radiation is required, it is necessary that all the slots have the same phase, therefore with the help of the CST Microwave Studio the phases of the slots of the subarrays are optimized to achieve this objective. The simulation obtained shows the phases obtained in one of the subarrays of the antenna, where it was observed that at the frequency of 30 GHz the phases of the slots are similar. The simulation also showed through the reflection coefficient, that for the S11 obtained in the subarray, it is observed that although they have acceptable values they are not good enough and this is due to the great length of the linear arrays that are part of the subarray.

2.10. Symmetric, Non-Symmetric Power Shunts, and Elbows

The CST software is used for the design of both symmetric and non-symmetrical shunt models.

2.11. Symmetric shunt

As it is known for this type of union, its objective is to divide the signal that is excited at the entrance to each of the outputs with the same amplitude and phase. For this design a discontinuity known as a septum was used and was made using curvatures around a metal plate [24] located exactly in the center of the joint. In addition to the entrance of the transition, an inductive symmetric iris was added, made with two progressive narrowings [25, 26]. In figure 4, the results of the simulations performed for the power and phase parameters for the modules S11, 12 and 13 are shown.
2.12. Non-symmetric shunt
For this design, as was done with the symmetric shunt, curvatures were used around a metal plate [24]. Additionally, two progressive narrowings were added at the entrance of the transition. Figure 10 shows one of the results obtained for a desired theoretical S13 of -3.61 dB and a desired theoretical S12 of -2.48 dB at the frequency of 30 GHz. When performing the simulations of the non-symmetric power derivative, corresponding to the module from S21 and S31 and the phase from S21 and S31, it is observed that acceptable operating values were obtained, for both parameters.

2.13. Elbows H and E
The elbows H and E were designed and simulated, introducing discontinuities at the 90 degree junction and for this a circular curvature was used [27, 28] with the appropriate values to obtain the lowest S11 possible.

In the design it was possible to observe the elbow in H, with a cut in its interior and the results of the simulations are shown where satisfactory values were obtained. With respect to the elbow in E, with a cut in its interior the circular curvature is evident, obtaining also satisfactory values.

2.14. Distribution network of 8 and 16 ports
Figure 5 shows the 8-port network, which includes symmetric shunts, elbows E and H.
With respect to the power network, according to the simulation, the bandwidth is approximately 6.2 GHz (26.2-32.4 GHz) for a SWR = 2 and 2.8 GHz for a SWR = 1.5.

Figure 6 shows the values of the modules of the transmission coefficients, observing acceptable values of equitable distribution of power.

![Figure 6. Cutting the plant view of the power network](image)

Figure 7 shows the 16-port network, which includes non-symmetric shunts.

![Figure 7. S11 of the 16-port power network](image)

The simulation for S11 of the 16-port power network shows that the S11 of the power network has a bandwidth of approximately 3.9 GHz (26.5-30.4 GHz) for a SWR = 2 and 1.21 GHz for a SWR = 1.5.

2.15. Design of the complete Radiation array
After the design of the subarrays, they are joined in the appropriate positions. Figure 8 shows a plan view of the complete antenna and a bottom view where you can see the ports of each subarray, which will allow us to simulate a study of mutual links between subarrays and passive reflection in each of them.

2.16. Mutual coupling between elements
It is considered as a mutual coupling between elements such as the signal level, in module and phase, which is coupled between the i-th and j-th subarrays. Since the simulation presented here has a port for each radiant subarray, this parameter will correspond, in module and phase, with 4.5. The results show
the mutual coupling, taking into account that there are 8 subarrays, in figure 9 the worst and best mutual coupling that each subarray receives from its neighboring subarrays is shown [29].

\[ s_{ij} \mid i \neq j, \forall j \]  

(6)

Figure 8. Plan view of the complete array and bottom view of the array for the study of couplings

Figure 9. Better and worse mutual coupling between subarrays

As expected, the mutual coupling between sub-arrays at t

The operating frequency of 30 GHz does not affect the radiation pattern.

2.17. Passive reflection parameter
This parameter informs about the reflection that occurs at the entrance of each of the radiating elements of the array when the rest of the elements are present, but not fed. We want to emphasize that this parameter is not the same as the reflection parameter of the isolated element, given that the fact that the rest of the elements of the array are present, even though they are not being fed, modifies the impedance of the element (in a non-active way). It corresponds in simulation for the i-th subarray with parameter 4.6. The results show the values for the passive reflection of each subarray of the array [29].

\[ S_i, \forall i \]  

(7)

In the simulation and analysis of the passive reflection for each subarray of the array, it is observed that each of the ports presents an adequate reflection value.

2.18. Active reflection parameter
It informs about the reflection that occurs at the input of each of the elements of the array, but this time taking into account that the rest of the elements of the array are being fed as well [29]. Figure 10 shows this parameter for each of the sub-arrays of the antenna.

The small difference between the active and the passive values is observed.
2.19. Obtaining the Radiation diagram and the S11 of the Array

Figure 11 shows the bottom profile view of the complete array, where the power network with its input port is observed.

2.20. Radiation diagram

Figure 12 shows the 3D radiation diagram of the complete array, for a frequency of 30 GHz, where it is observed that it complies with specifications such as the gain and beam width at the frequency of 30 GHz, but does not comply with the desired NLPS of 20 dB and this is due to the fact that the mutual coupling between slots was not taken into account.
Figures 13 and 14 show the diagram in cartesian coordinates, where the polar and cross-polar radiation is observed for the planes $\phi = 0^\circ$ and $\phi = 90^\circ$ respectively. Finding that the target of an NLPS of 20 dB is not met and this may be possible due to the fact that the effects of the mutual coupling between slots were not contemplated.

\[ g = 2.09 \frac{\lambda_g}{\lambda} \frac{a}{b} \cos^2 \frac{\pi \lambda}{2}\lambda_g \sin^2 \frac{\pi \lambda}{a} \]

Where $\lambda$ is the wavelength in the free space, $\lambda_g$ is the wavelength in the guide, $x_1$ is the displacement from the axis of the guide and $a$ and $b$ are the width and height of the rectangular waveguide, as described in [10, 11], where the effects of the mutual coupling between slots are not contemplated, but which gives satisfactory results in linear arrays but not in flat arrays, additionally and taking into
account that the subarrays were designed of 12x6 slots, presenting arrays very long linear (12 slots), this influenced the reduction of bandwidth.

3. Conclusions
With the realization of this design and its corresponding simulation, it is confirmed that in the construction of slot antennas on waveguide, one of the most important problems is the construction of the guide and the precision of the slots. A Slot Waveguide Array Antenna (SWAA) was designed and simulated for satellite communications, which was performed using the CST Microwave Studio software. It was found that the parameters of the antenna comply with some of the specifications of the initially proposed design, such as the gain and beam width, but it does not comply with the bandwidth and the NLPS and this is because the subarrays very long, 12x6 slots were made, reducing the bandwidth, however it is verified that the subarray method is appropriate for increasing the bandwidth in this type of arrays. It was found that the Taylor design criterion is appropriate for this type of flat arrays because it allows a more equitable distribution of power in each subarray allowing for easier design. It is necessary for future work to perform the modeling of elements in a more precise and detailed manner, an important aspect for the improvement of the power network of the flat array and the joints and bends in both modeled planes.

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