

# Optimized design of antenna systems for radar applications at 37GHz

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## Abstract

This paper describes a rigorous and efficient procedure for designing large circularly-polarized antennas, especially oriented to radar applications. Specifically, it deals with the optimized design of large arrays of resonant shunt slots loaded with parasitic dipoles, by proposing an approach to tune up the standard uniform array design. Such method turns out to be fundamental when external mutual couplings among radiators exhibit strong levels. This is the case treated in this paper, where radiators are etched on a single-layer electrically thick dielectric sheet. For validation purposes, two challenging prototypes are designed, fabricated and measured, reporting here the most representative results.

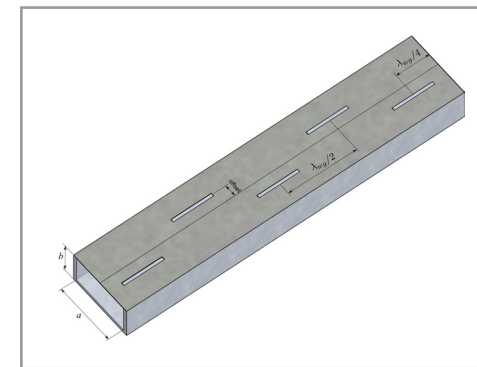
**Keywords.** Slot arrays, waveguide slot array, circular polarization, radar antennas, parasitic dipoles, steerable antennas, monopulse antennas, array design, array optimization.

## 1. Introduction

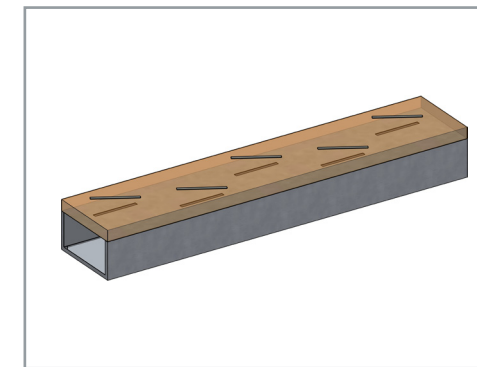
Radar applications in millimeter-wave band call for very efficient and versatile antennas capable to meet all required specifications, such as conformal low profile, robust mechanical structure, low sidelobes, steerable operation or cosecant-shape pattern, among others. One of the most used antenna topology in radar has been a slotted-waveguide array [1], thanks to its numerous advantages as geometric simplicity, high efficiency, polarization purity, good reliability and conformal installation.

The most common array configuration [1] consists of longitudinal shunt slots located on the broad face of a standard short-ended mono-mode rectangular waveguide. Slots are oriented parallel to the waveguide axis and placed alternately on both sides of it. An axial inter-slot spacing of  $\lambda_{wg}/2$  assures in-phase radiation toward broadside direction, being  $\lambda_{wg}$  the wavelength within the rectangular waveguide. Since a short-ended waveguide supports a standing wave along its axial direction, all slots will lay on a standing wave maxima. A diagram of this configuration is depicted in Fig. 1, where slots' offsets relative to the guide axis control their excitation weights and consequently the H-plane radiation pattern. The main drawback exhibited by this configuration is the bandwidth limitation caused by the intrinsic narrowband resonant behavior of the waveguide field. Nevertheless, available bandwidth is usually enough for typical narrowband radar operation.

Obviously, the above described disposition of radiators only is able to radiate pure linear polarization whereas circular polarization is strongly required by radar systems. In previous works, several approaches have been proposed to achieve circular polarization in rectangular waveguide slot arrays. On the one hand, more complicated slot shapes, e.g. crossed slots [2], are capable to radiate circular polarization when they are excited by the waveguide fundamental mode. However, such elements must be spaced  $\lambda_{wg}$  for broadside radiation, thus forcing to resort to tilted leaky-wave designs [3]-[4] so as to avoid high diffraction sidelobes. On the other hand, alternative radiating elements, such as probe-fed microstrip patches [5] may be used. This option,



■ Figure 1. Resonant shunt slot array.



■ Figure 2. Slot-dipole array for circular polarization

however, presents modeling and construction difficulties particularly at millimeter-wave bands where manufacturing tolerances are known to be very critical.

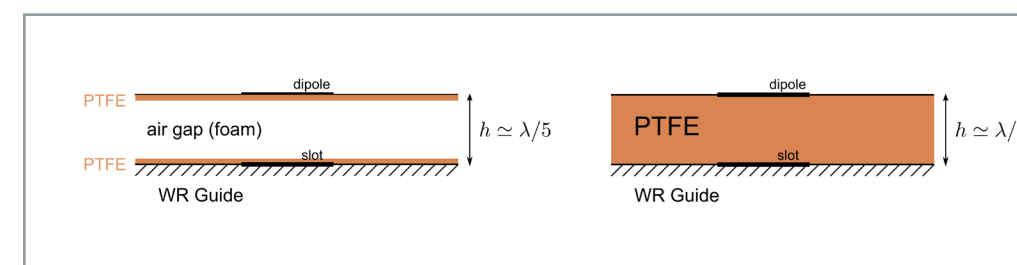
Finally, another approach involves making use of linear-to-circular polarization converters. In a former paper [6], it was demonstrated that circular polarization can be produced by placing a tilted parasitic dipole (Fig. 2) at an approximated distance of  $\lambda/5$  right above a slot. Later in [7], this technique was successfully extended to an array, where parasitic dipoles were etched on a very thin polyethylene sheet placed on top of an expanded polystyrene layer. As reported by the authors, such a low permittivity material (nearly that of vacuum) allowed them to neglect it in designing length and orientation of dipoles to achieve circular polarization. Furthermore, analysis of the whole antenna may be carried out by considering just a few surrounding elements because of their weak mutual couplings. An additional advantage in design is derived from the fact that dipole array showed no major influence on the slot array which needed no modification.

In the case discussed in this paper, the array is designed for operation at 37 GHz (Ka band). A proper choice of materials on which slots and dipoles will be etched plays a fundamental role in electrical and mechanical antenna characteristics. A slot-dipole layout based on 'light' materials like those of [7] is first considered, whose side view scheme is depicted in Fig 3a. Both slots and dipoles are etched on very thin, almost negligible, PTFE sheets. The space left between them is filled by a foam material, with an electrical behavior very similar to air. Although this first op-

tion enjoys all the above mentioned advantages attributed to design in [7], it suffers from a lack of mechanical robustness which becomes crucial in radar applications. In addition to that, etching process and welding to the rectangular guide as well as alignment of such thin sheets make fabrication process very exigent and often unsuccessful. As a consequence, a new slot-dipole layout must be adopted to alleviate fabrication exigencies and enhance antenna resistance.

Alternatively, slots and dipoles can be etched on both sides of a unique PTFE substrate whose thickness, therefore, must be around  $\lambda/5$  (little more than 1 mm at 37 GHz) to produce circular polarization (see Fig. 3b). The vertical distance and horizontal alignment between slots and dipoles are now inherently guaranteed by a standard PCB fabrication process. An obviously enhanced mechanical robustness of this single-piece structure supporting all radiating elements also helps in the subsequent welding to the guide. Such mechanical and fabrication benefits turn into difficulties when electrical performance is evaluated. A  $\lambda/5$  thick dielectric material with a permittivity around 2.2 as the one considered, permits propagation of significant surface-waves, which leads to much stronger mutual couplings among radiators than those of the structure based on light materials. Fortunately, the effect of those augmented interactions may be accurately evaluated by making use of a rigorous analysis code and properly corrected to achieve the desired performance.

The design method of a circularly-polarized slot-dipole array is explained in detail in a previous authors' paper [8]. In few words, this approach



■ Figure 3. Side view of proposed slot-dipole layouts, (a) based on light materials and (b) single-layered

A single-layer fabrication of slots and dipoles is adopted to improve assembly, fabrication, durability and mechanical robustness

applies a modified version of the so-called equivalent self-admittance method [1], so that slot resonant length and dipole tilting and length are determined for a given offset by assuming identical radiators throughout the whole array. As a result, the arrangement of all antenna elements can be defined. However, with this initial disposition, the desired performance of the antenna is not completely achieved. Notice that the design method necessarily considers identical antenna elements, which does not correspond to the actual antenna. This departure has a noticeable impact on antenna parameters due to the huge mutual couplings mentioned above. Hence an iterative tune-up procedure starting from this first design is proposed in this paper.

This work describes the required optimization procedure followed for a slot array antenna loaded with parasitic dipoles for circular polarization, when an electrically thick dielectric substrate is employed. Section 2 summarizes the initial array design procedure described in [8] while section 3 describes the optimization approach itself. Next, section 4 reports some representative results of the proposed technique, whose application to the design, fabrication and measurement of prototypes is illustrated in section 5. Finally, section 6 summarizes the main conclusions of this paper.

## 2. Design method

### 2.1. Radiating element design

In every antenna design procedure, the accuracy of the electromagnetic analysis tool employed determines the validity of final results. This correspondence is greatly accentuated in a case like the one treated, where mutual couplings and manufacturing tolerances play such a significant role. The present work adopts the modeling technique described in [8], which applies the Method of Moments (MoM) by using specialized Green's functions for both the multilayer external region and the rectangular waveguide internal region. Due to their particular geometry, magnetic currents on slots and electric currents

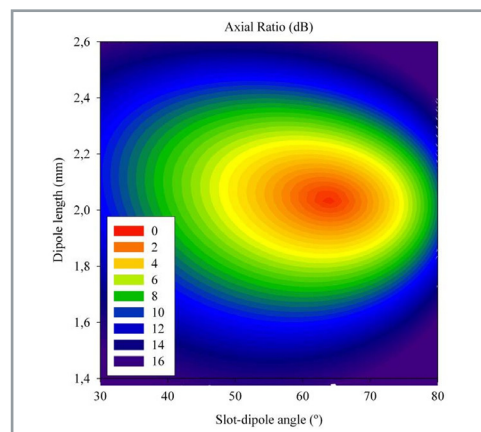


Figure 4. Axial Ratio as function of slot-dipole angle and dipole length.

on dipoles are expanded into a single basis function per slot/dipole, with cosine distribution in the longitudinal dimension and uniform distribution in the transversal dimension. This equivalence in the analysis method (one basis function per element) turns out to be fundamental for a time efficient design and further optimization approach when a large antenna is considered. It is worth noting that such ad-hoc code was successfully applied in the past by these authors [9], confirming its accuracy by measurements.

In order to achieve circular polarization (CP) with the radiating elements of Fig. 2, the electric vertical distance between a slot and its corresponding parasitic dipole must be at least  $\lambda/5$ . Since mutual couplings increase as substrate is thicker, a separation between slots and dipoles as short as possible is recommended. Once a commercial microwave substrate is chosen, parasitic dipole length and its tilt with respect to the slot must be designed in order to confirm the adequateness of chosen material. In Fig. 4, axial ratio for a slot-dipole pair is represented with a substrate thickness of 1.143 mm and relative permittivity of 2.2. Slot length is set to  $0.4\lambda$  and dipole length and tilt are modified. It can be readily verified that the axial ratio remains below 3 dB (typical specification) for a wide range of dipole parameters. A good polarization bandwidth is therefore expected, which becomes essential due to the already mentioned critical manufacturing tolerances.

### 2.2. Array design procedure

In a conventional shunt slot resonant array design, the desired radiation pattern is synthesized by shifting each slot a certain offset with respect to waveguide axis. There exists a correspondence between offset displacement and its excitation weight, and at the same time, the slot resonant length [10] needed to match. An identical procedure is adopted here, but now additional information about length and inclination of the dipole must also be obtained by imposing circular polarization. When this radiating element is part of an array, on the other hand, resonance parameters are modified by the effect of mutual couplings. Hence, an accurate estimation for the slot length, dipole length and dipole tilt must take into account the presence of every array element. Since final arrangement of elements is still not known at this design step, a uniform infinite array is usually considered. This is a reasonable approximation because of a predictable slow variation of physical dimensions for nearby elements. However, this well-known infinite array approximation [11] should not be applied in this case due to stronger mutual couplings, and therefore a uniform finite array model and the method of moments is employed to reach more accurate results. Finite model is certainly slower computationally speaking but it allows to use the actual amplitude and phase distribution of waveguides and radiators and to take into account edge effects.

An iterative approach [8] was proposed by these authors to compute a design tabular data relating each considered offset to slot-dipole resonance parameters. To that end, a slot-dipole pair is placed at the centre of the considered finite array arrangement formed by identical elements. Slot length is varied until resonance (purely real impedance) is achieved. Next, keeping slot unaltered, dipole length and tilt are designed for circularly-polarized radiation. Last variations in dipole parameters in turn affect the slot resonant length, thus forcing to apply an iterative process described in Fig. 5. The algorithm shows a good behavior, and convergence is achieved after few iterations.

Once slot-dipole dimensions to achieve resonance and circular polarization are obtained for each offset, the corresponding equivalent conductance of the resulting radiator is computed. This is the widely known equivalent self admittance method [1], which assumes a radiated power proportional to the element conductance. As a final result of this preliminary step, a tabulated data for each offset is found, with the following parameters: slot length, dipole length, dipole tilt and equivalent conductance, all of which defining a circularly-polarized resonant element.

A one-dimensional shunt slot-dipole array can be replaced by an equivalent circuit where each element is represented by its corresponding equivalent shunt conductance (see Fig. 6). In order to ensure a perfectly matched input port at the working frequency, the sum of all normalized conductances is required to be equal to 1 both for end or centre feed. This last alternative is chosen here to improve overall antenna bandwidth. On the other hand, the desired aperture illumination is attained by the current distribution of conductances, knowing that the amplitude weight of each element is proportional to its equivalent conductance. Mathematically, for a one-dimensional array of  $N$  elements, the above matching and illumination conditions are expressed as follows [1]:

$$\sum_{n=1}^N g_n = W \quad [1]$$

$$g_n = KA^2(n) \quad [2]$$

In the above equations  $g_n$  is the normalized resonant conductance of the  $n$ th slot,  $A(n)$  is the desired aperture voltage distribution,  $K$  is a normalization constant, and  $W$  equals to 1. Solution for  $g_n$  of above system provides the distribution of conductances throughout the array, which can be readily translated into the final physical geometry of elements by means of the previously computed table.

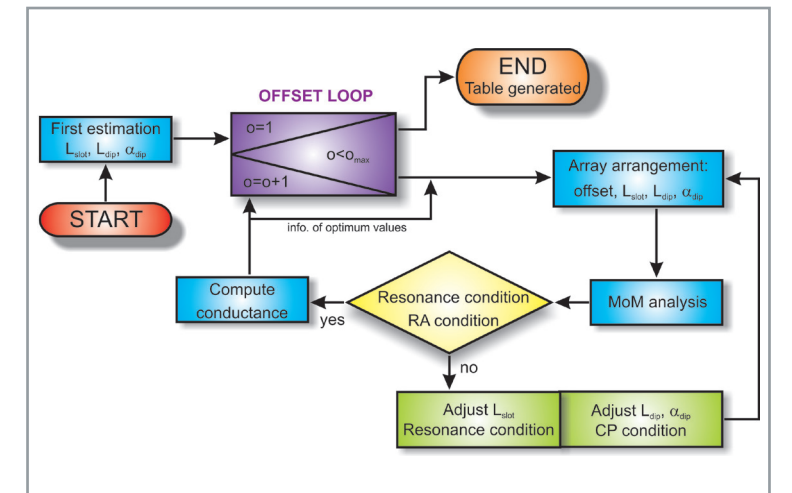


Figure 5. Iterative design process for a shunt slot-dipole array.

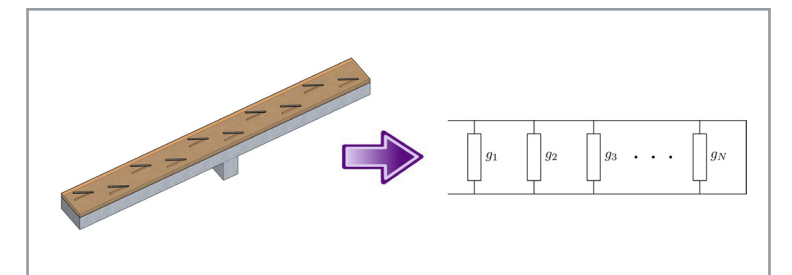


Figure 6. Equivalent circuit of a one-dimensional shunt slot-dipole array.

When designing a two-dimensional array like the one depicted in Fig. 7, (1) and (2) must be solved for every row (guide), which leads to the same set of equivalent conductances since voltage distribution  $A(n)$  is unique. Correspondence between  $g_n$  values and physical parameters of radiators, however, change from one row to another due to the applied finite array analysis. As a consequence, the two-dimensional array does not consist of a simple juxtaposition of identical rows. In fact, mutual couplings between adjacent rows are very strong because of elements' orientation, so edge effects have a notable influence on final array design.

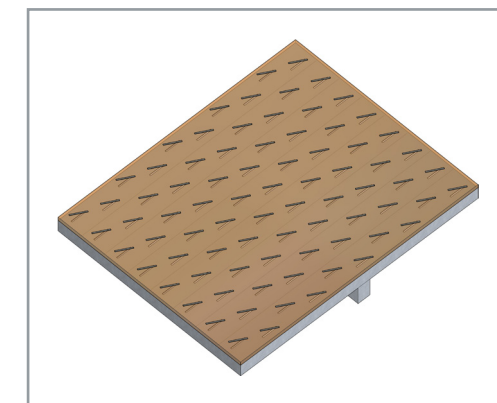


Figure 7. Example of a two-dimensional slot-dipole array formed by 8 rows with 10 elements each.



Two prototypes with different specifications has been designed, manufactured and measured

### 3. Optimization procedure

The implementation of the above design method gives as a result the full arrangement of antenna elements, which is completely defined by offsets and lengths of each slot and orientation and length of each dipole. In short, that method focuses on the following three antenna parameters for design:

- **Aperture illumination:** the desired weight profile of radiators in each row must be achieved by the solution for  $g_n$  of equations (1) and (2). As stated above, such conductances are established by slots' offsets with respect to their corresponding guide axis.
- **Guide matching:** in order to reach matching at the input port of every guide, the sum of equivalent admittances of radiators belonging to each guide must be equal to 1. The real part of this sum is already accomplished if the array illumination satisfies the conductance profile  $g_n$  determined in the above point. However, the imaginary part of the admittance must also be zero for each element, which is mainly controlled by its slot length.
- **Circular polarization:** an axial ratio as close as possible to the ideal 0dB value should be reached. As explained previously, this requirement is satisfied by the proper tuning of dipole length and orientation.

With this initial disposition, however, the desired performance of the antenna is not completely achieved. Notice that the design method necessarily considers identical antenna elements, which does not correspond to the actual antenna. This disagreement affects all mutual couplings in a marked way, which alters antenna parameters. It becomes particularly important for the case treated, where mutual couplings are stronger than usual because of the electrically thick dielectric substrate employed.

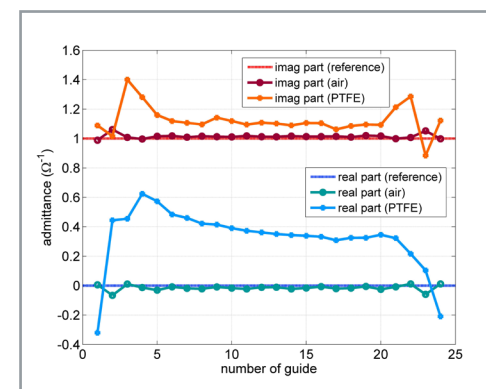
Hence, an iterative tune-up procedure must be carried out starting from this first design, which can be considered a good initial point. In order to meet all above criteria, three corrections are performed successively for each antenna radiator: a slot offset modification to assure the required aperture illumination; a slot length correction to achieve a proper matching for every guide; a dipole adjustment (length and orientation) to improve the axial ratio figure. Notice that each adjustment somehow distorts what is attained by previous corrections. For this reason an iterative procedure is required to accomplish all above antenna parameters. As stopping conditions for this process, a tolerance for the axial ratio and the sum of row admittances may be fixed.

### 4. Results

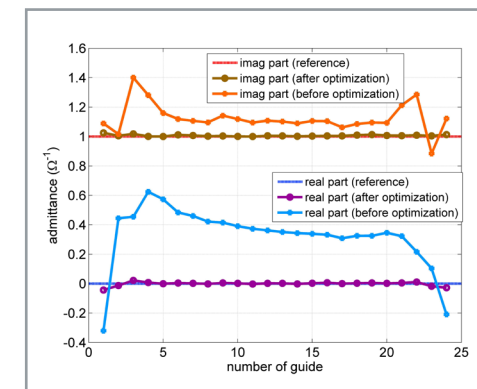
Above described design and optimization approach for two-dimensional slot-dipole shunt arrays is applied to two antennas with different specifications. In both designs, slots and dipoles are etched on a standard PTFE dielectric sheet, with a thickness of 1.143mm and a relative permittivity of  $\epsilon_r = 2.2$ . A 5.1 mm thick gap of air is left between dipoles and a radome made of a 5mm-thick piece of polycarbonate ( $\epsilon_r = 2.7$ ), conceived to give more consistency to the antenna structure. In this section, representative results of the design and optimization approach are shown, leaving fabrication and measurements issues for Section 5.

Without loss of generality, a 24 waveguide array with 40 radiating elements per waveguide is considered in this section. A uniform weight profile is applied in the elevation plane (transverse to the waveguides). On the other hand, a cosine on a pedestal distribution is implemented in the azimuth plane, which should be attained by the slot-dipole design method described here. For illustration purposes, a preliminary result is shown to realize how surface waves alter the natural performance of the standard design method described in Section 2. Fig. 8 represents the imaginary and real part of the sum of equivalent admittances in each waveguide, whose value must be close to zero and two respectively for proper matching, when space between slots and dipoles is filled with both the current PTFE material and air. It appears evident that for this last case, which is free of surface waves, the design method provides optimum results while PTFE case needs an optimization approach to correct admittance values.

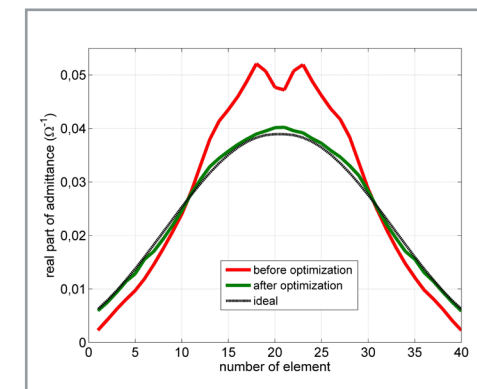
Therefore, the optimization approach described in section 3 is carried out on this antenna, converging after few iterations. The sum of admittance values of the final antenna (after optimization) compared with those of the first design (before optimization) are shown in Fig. 9. All deviations caused by unusually strong mutual couplings are clearly corrected and final admittanc-



■ **Figure 8.** Sum of equivalent admittances for each guide with air and PTFE.



■ **Figure 9.** Sum of equivalent admittances for each guide before and after optimization approach.



■ **Figure 10.** Equivalent conductance of each element along the central guide before and after optimization approach.

es attain nearly ideal values. Additionally, the amplitude distribution along each guide must also be similar to the desired cosine over pedestal function. For example, Fig. 10 represents the real part of equivalent admittances (conductances) for each element along the central guide. Once more, optimization approach successfully restores the conductance distribution. All exhibited results make evident the significant benefit of the optimization procedure on conductances, which will lead to a matching improvement of every guide and a radiation pattern as ideal as theoretically possible. Yet a last requirement must be accomplished, i.e. a good value of axial ratio. In this case, original axial ratio is 0.95dB, which translates into 0.09dB once optimization approach is carried out.

## 5. Prototype description

### 5.1. Prototype 1: Cosecant-pattern antenna

A usual specification on radar applications calls for a cosecant-shape radiation pattern in the elevation plane, transverse to the waveguides, which is realized here by a proper design of the feeding network. Furthermore, a monopulse operation on both planes is recommendable for tracking performance. An elegant design of

waveguides feeding network is therefore demanded to accomplish both requirements. Note that a cosecant-shape radiation pattern requires a non-uniform phase distribution which is designed by applying the pattern synthesis algorithm described in [12]. Such weight profile is realized by an intricate multilevel waveguide network, designed and tested for that purpose. In particular, an arrangement of 24 waveguides with 36 radiating elements each is adopted.

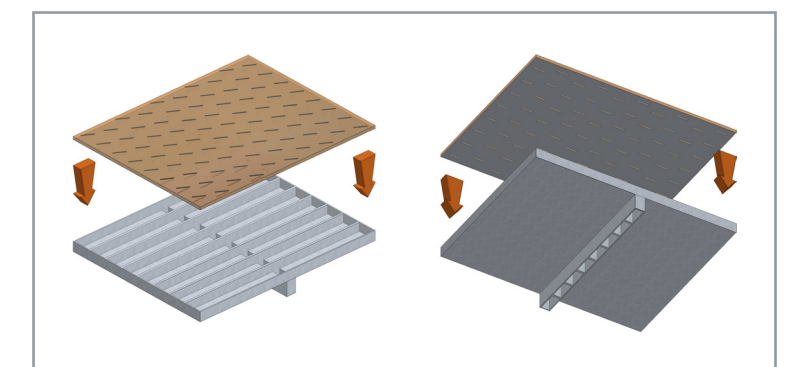
Conversely, a cosine on a pedestal amplitude distribution is synthesized in azimuth plane in order to reduce secondary lobes. Once slots and dipoles are fully designed and optimized, they are fabricated on a single dielectric laminate, and welded to the aluminum structure supporting the short-ended waveguides. A scheme of the setup procedure is depicted in Fig. 11. Waveguide feeding network mentioned above will be connected to the inputs of the short-ended waveguides.

A complete measurement process confirmed a good electrical performance concerning all antenna characteristics. In first place, Fig. 12 represents measured radiation patterns for both elevation and azimuth planes. A cosecant-shape profile is fully achieved in elevation plane within the specified thresholds while a low-sidelobe high directive beam is attained in azimuth plane.

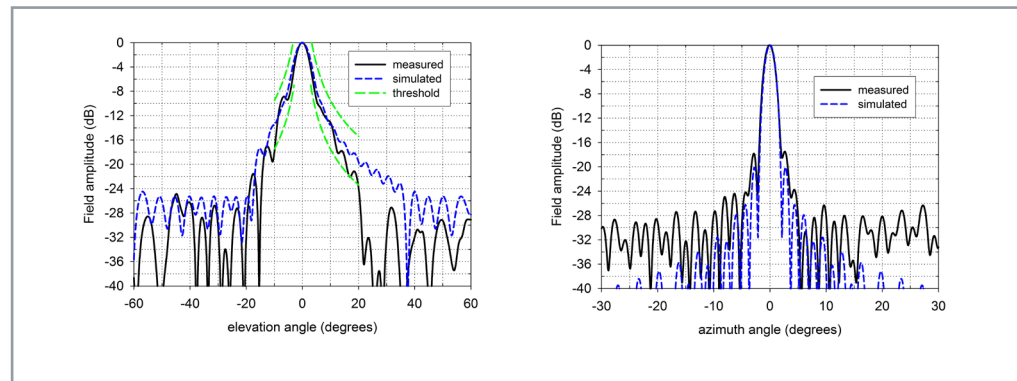
With regard to the circular polarization specification, an axial ratio of 1.9 dB is achieved, which is well above the simulated value, but better than the 3 dB requirement. Maximum directivity reaches 34.1 dBi, while maximum gain is 32 dBi. This 2 dB difference may be attributed to losses in conductors and dielectrics and leaking fields caused by an imperfect assembly of antenna pieces, almost unavoidable at these frequencies. Finally, matching requisite is fully accomplished thanks to an input reflection coefficient below -20dB in the whole working frequency band.

### 5.2. Prototype 2: Electronically Steerable Antenna

One important value-added feature on radar antennas addresses the possibility of electronically modifying the main beam direction. Such char-

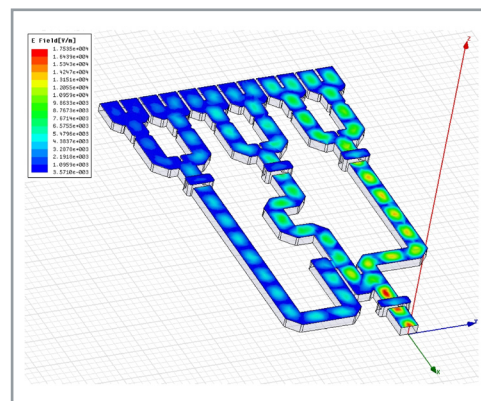


■ **Figure 11.** Top and bottom view of the antenna assembly.

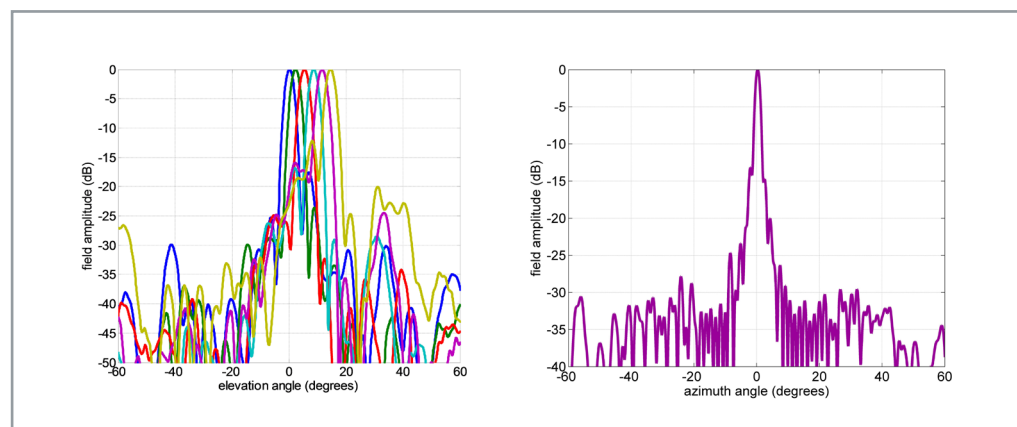


■ **Figure 12.** Radiation patterns in (a) elevation and (b) azimuth plane for prototype 1.

acteristic is usually answered by the incorporation of phase-shifters into the feeding antenna network, making its implementation even more complicated for this case. It should be noted that an efficient integration of multi-state phase-shifters into rectangular waveguides has not been reported to date. In the case treated here, the feeding network synthesizes a low-sidelobe (25dB SLL) Chebyshev illumination in elevation plane. Furthermore, a monopulse operation in both planes is also demanded. Fig. 13 represents a part of the designed feeding network with simulated fields. An array of 24 waveguides with 40 radiating elements each are adopted for this design. On the other hand, design and optimization of



■ **Figure 13.** Part of the Tschebysheff waveguide feeding network for prototype 2.



■ **Figure 14.** Radiation patterns in (a) elevation and (b) azimuth plane for prototype 2.

slots and dipoles are carried out to achieve a cosine on a pedestal amplitude distribution in azimuth plane, similar to that of the first prototype. Apart from the phase-shifters circuitry, an identical fabrication process assembles the final antenna structure.

An exhaustive measurement process was also carried out on this antenna by testing all phase-shifters states and verifying a correct steerable performance. Only a sample of those results is included in this paper for the sake of brevity. In Fig. 14a measured radiation patterns in elevation plane for different steering angles are represented. As can be observed, well-shaped radiation patterns are achieved, though they gradually deteriorate as scanning angle deviates from broadside direction. This effect is owing to the fact that radiators weights are optimized for broadside direction and phase-shifters are mainly calibrated for an optimum performance on zero-state. Nevertheless, scanning performance works in a reasonable way for cited application in pointing the beam to the desired direction without an important decrease in directivity. Regarding azimuth plane radiation pattern represented in Fig. 14b, a good low-sidelobe directive beam is attained, which corroborates the effectiveness of the slot-dipole optimization.

A gradual deterioration for higher scanning angles is also observed in axial ratio figures, even though the 3 dB criterion is satisfied for an ac-

ceptable range of scanning angles. A measured maximum directivity slightly above 35 dBi for the valid range of steering angles confirms the validity of the design. Measured gain, however, drops to 24 dBi because of the already predicted huge losses of phase-shifters. Note that current frequency band lacks efficient designs of phase-shifters in rectangular waveguide technology, leading to resort to monolithic devices with inherent high losses. Nevertheless, angle discrimination is preserved as a result of low beamwidth values, and therefore the reduction of gain will only affect to a shorter distance range of the radar system. Lastly, it is worth noting that matching measurements also fulfills return loss specification for all frequencies in the valid range of scanning angles.

## 6. Conclusions

This paper deals with the design of circularly-polarized slotted rectangular waveguides arrays loaded with parasitic dipoles, mainly conceived for radar applications. A single-layer fabrication of slots and dipoles is adopted with the aim of facilitating the assembly and fabrication of the antenna as well as improving its mechanical robustness and durability. Application of a standard design method to such an electrically thick substrate fails due to strong surface waves propagation. An optimization procedure is proposed here to fine-tune matching and radiation performance given by a conventional design based on a uniform array model. The proposed procedure has revealed to be essential when strong mutual couplings are presented.

A successful antenna development lays both on a well-designed waveguide feeding network and an optimum slot-dipole arrangement. Two prototypes with different starting specifications has been designed, manufactured and measured. A monopulse cosecant-pattern antenna was first considered. Measurements confirm the validity of the proposed optimization method, since radiation patterns in elevation and azimuth plane correspond to a cosecant and a low-sidelobe directive beam respectively. Axial ratio and input reflection coefficient also meet antenna requirements.

The second antenna required a low-sidelobe directive beam in azimuth plane and an electronically steerable directive beam in elevation plane, also able to work as monopulse operation. A full integration of phase-shifters into a Tschebysheff-profile waveguide feeding network allows accomplishing the desired steering capability. Such functionality is demonstrated by measurements at different scanning states. Optimization approach provides high-directive radiation patterns in both planes, good matching behavior, and axial ratio values below 3 dB. However, a gradual deterioration of radiation characteristics is observed as scanning angle increases.

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## Biographies



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