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DESIGN AND SIMULATION OF A SLOTTED WAVEGUIDE ANTENNA FOR COASTAL RADAR

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ABSTRACT

Systems consisting of a slotted waveguide antenna on the narrow face of the waveguide and a reflector are widely used in naval and coastal radars. In order to obtain an antenna for coastal radar applications in the X band, the design of a system based on a slotted waveguide antenna in the horizontal plane and a reflector in the vertical plane is proposed. The design is done on both planes separately. For the design of the slotted waveguide, the traveling wave array design equations are used and a Taylor one-parameter amplitude distribution is assumed. The reflector is designed based on obtaining a half power beamwidth determined by the vertical plane. In both designs the Mathcad software is used. According to the analytical models, the results of the simulation of the final antenna in the HFSS software obtain a half power beamwidth in the vertical and horizontal plane of 26^0 and 0.6^0 respectively and secondary lobe levels of - 25 dB and -30 dB respectively. To reduce the cross polarization level, a grid filter with parallel conductive strips was designed, which reduced the cross polarization level to -41 dB. The final antenna gain is 31.051 dB.

INDEX TERMS: Cross Polarization filter, reflector, slot, slotted waveguide antenna, Taylor one-parameter.

DISEÑO Y SIMULACIÓN DE UNA ANTENA DE GUÍA DE ONDA RANURADA PARA RADAR COSTERO

RESUMEN

Los sistemas que consisten en una antena de guía de onda ranurada en la cara estrecha de la guía de onda y un reflector son ampliamente utilizados en radares navales y costeros. Para obtener una antena para aplicaciones de radar costero en la banda X, se propone el diseño de un sistema basado en una antena de guía de onda ranurada en el plano horizontal y un reflector en el plano vertical. El diseño se realiza en ambos planos por separado. Para el diseño de la guía de onda ranurada, se utilizan las ecuaciones de diseño de arreglos de ondas viajeras y se asume una distribución de amplitud de Taylor de un parámetro. El reflector se diseña basándose en la obtención de un ancho de haz de media potencia determinado por el plano vertical. En ambos diseños se utiliza el software Mathcad. De acuerdo con los modelos analíticos, los resultados de la simulación de la antena final en el software HFSS obtienen un ancho de haz de media potencia en los planos vertical y horizontal de 26° y 0,6°, respectivamente y niveles de lóbulos secundarios de -25 dB y -30 dB, respectivamente. Para reducir el nivel de polarización cruzada, se diseñó un filtro de rejilla con láminas conductoras paralelas, que ha permitido reducir el nivel de polarización cruzada a -41 dB. La ganancia final de la antena es de 31,051 dB.

PALABRAS CLAVES: Filtro de polarización cruzada, reflector, ranura, antena de guía de onda ranurada, Taylor de un parámetro.

1. INTRODUCTION

Their ease of construction, strength and integration into existing designs, as well as their low cost and efficiency, make Slotted Waveguide Antennas (SWA) a very good choice for radar, communications and space navigation applications [1-4]. The designs of these antennas are linear arrays of slots that have a defined amplitude and phase distribution to obtain a given radiation pattern [5-8]. This type of antenna is classified according to the geometric position of the design and the type of feed source. In marine radars, edge wall slots arrays are generally found on the narrow face of the WR-90 waveguide at frequencies in the X-band. The antennas of the Navy-Radar 4000 and Furuno radar are examples of SWA with the slots on the narrow face of the WR-90 waveguide.

On the other hand, aperture antennas are popular for their ability to direct the electromagnetic beam in such a way

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that they concentrate the emission and reception of the radiating system in one direction [9]. There are several types of aperture antennas such as: horn antennas [10], satellite antennas and reflective surfaces in general. A horn antenna consists of a waveguide, in which the area of the section is progressively increased to an open end that behaves like an opening. Its construction is easy and it has a low cost.

For radar applications in the marine environment, such as coastal radars, antennas formed by a slotted waveguide antenna in the horizontal plane and a reflector in the vertical plane are used to obtain far-field directive radiation patterns and low Secondary Lobe Levels (SLL) as can be observed in [1, 2]. Cuba is a country surrounded by the sea, which is why the inspection of territorial waters and exclusive economic zones is essential to guarantee its safety and protection. An Exploration System (ES) is essential to help search and rescue at sea, maritime and port security, navigation (including the operation of lights and buoys at sea), the fight against drug trafficking, control of borders and the fight against marine contamination.

Our country has proposed to advance on the path of technological independence and import substitution, which is why efforts are directed to the design of parts and components of coastal radars compatible with those already existing in the country (Navy-Radar 4000 and Furuno radar). In a previous paper by the authors [8], to adapt the Navy-Radar 4000 to a coastal radar, its antenna was characterized and simulated in the HFSS software. On that occasion, the geometric dimensions of the antenna were obtained by physically measuring it and not through its design. In [11] the amplitude distribution on the array axis is unknown.

Taking into consideration that the antenna design is an essential part for the effectiveness of the ES, the situation occurs that the design of the antenna that can be used in the coastal radars of our country had not been carried out, taking into account criteria like the amplitude distribution, the SLL and the Half Power Beamwidth (HPBW). Taking this as a precedent, the objective of the present investigation is to design and simulate a slotted waveguide antenna with a reflector to be used in coastal radar applications. The antenna design will be carried out in both planes separately. In order to reduce the cross polarization level a filter of this type of polarization will be designed.

2. DESIGN OF THE TRAVELING WAVE SLOTTED WAVEGUIDE ANTENNA OF THE COASTAL RADAR SYSTEM

This section is dedicated to the design of the SWA. The procedure for determining the normalized excitation coefficients of the array elements in a first step is explained. Then the conductance of each slot is calculated and finally the laws of the angle of inclination and penetration for the traveling wave SWA are obtained.

Initial Design Considerations

For linear resonant slot arrays on the narrow face of the WR-90 waveguide, where at the design frequency 9.41GHz the waves propagate in their fundamental mode TE_{10}, the slots are arranged at alternate angles $\pm\ \theta$ at a separation of $\lambda g/2$ (where λg is the waveguide wavelength) where the wave energy reaches its peaks, to guarantee a constant phase distribution. In the case of non-resonant arrays or traveling wave arrays, the slots are arranged at a separation different than $\lambda g/2$. In both cases, the variation of the angle of inclination of the slot on the narrow face of the waveguide is the common technique to control the power radiated by the slot [3].

The slots penetrate the wide face of the guide at a distance d (Fig. 1) since the height of the narrow face of the WR-90 guide (b=10.16mm) is less than a half of a free space wavelength ($\lambda_0/2$), which is approximately the resonance length of the slot for any angle of inclination. Therefore, to maintain resonance for any angle of inclination, a certain d is required. Penetration d decreases when angle θ increases and vice versa.

Based on the bibliography consulted and the experience of the authors in the study of SWA, for this design a thickness of the WR-90 t = 1.75mm and a width of the slot w_r = 1.75 mm are assumed (Fig. 1). For 9.41GHz, λg=44.42mm and a HPBW of 0,7 $\pm$ 0.1º and a SLL <-30 dB is required for the horizontal plane, with a separation from the center of one slot to the center of the closest slot (d_r) of approximately 0.56λg (24.87 mm) and 86 slots. The final system gain should be > 30 dB.

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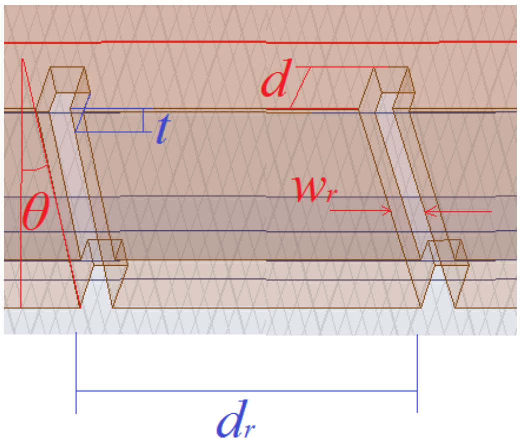

Figure 1: Detail of the slot dimensions in the WR-90 waveguide.

Design Procedure

The design of a resonant and non-resonant linear slot arrays on the narrow face of the guide starts from defining the amplitude distribution and then the excitation coefficients of each slots from the sampling of the distribution function. Subsequently, the normalized conductance corresponding to each slots are determined and then the laws or functions of θ and d respectively [8].

For the present case, the design of an antenna is required for a coastal radar application. The discrete Dolph-Tschebyscheff array design is known to offer radiation patterns with equal intensity secondary lobes for all angles while the Taylor Tschebyscheff-error produces patterns with narrow HPBW and constant first interior lobes at time the other lobes decrees monotonously [5]. However, our case corresponds to that of radar in a marine environment. For some applications such as radars or low noise systems, monotonically decreasing levels of the secondary lobes are desired from the same first inner lobe although the radiation angle of the pattern is sacrificed to some extent at the expense of a low level of the first secondary lobes. An array design that meets these requirements is Taylor one-parameter; therefore, we select this distribution in the present design.

Calculation of Normalized Excitation Coefficients

The excitation coefficients of the 86 elements of the array necessary to obtain the desired far-field radiation pattern are calculated from the equations of the Taylor one-parameter amplitude distribution [1,5,8], programmed in Mathcad. Fig. 2 shows the distribution of the normalized excitation coefficients (a_n) along the right half axis of the array. The amplitude distribution is known to be symmetric with respect to the coordinate origin. These values are summarized in Fig 1. The left half of the Table 1 shows the first 43 slots and the right half the remaining 43.

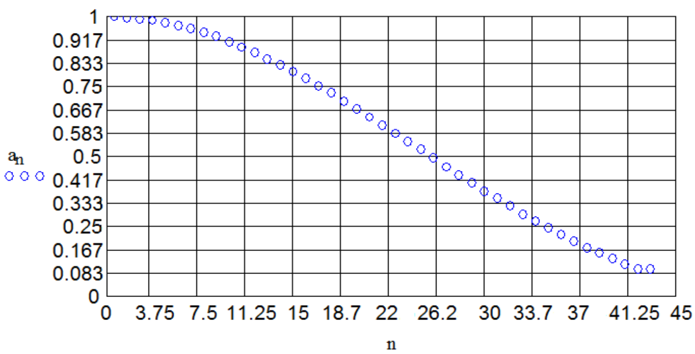

Figure 2: Normalized amplitude excitation coefficients for Taylor one-parameter.

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Calculation of the Conductance of the slots and Determination of the Inclination Angles

The circuital model of the non-resonant slot array on the narrow face of the WR-90 waveguide is shown in Fig. 3. Each slot can be represented by an admittance load normalized to the admittance of the WR-90 waveguide y_n= g_n= jb_n, where g_n and b_n are the normalized conductance and susceptance of each slot respectively. If the slots are in resonance, b_n = 0.

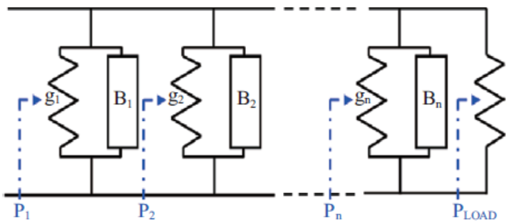

Figure 3: Circuital model of a slotted waveguide on the narrow face.

To determine the individual conductance of each slot, Eq. (1) is used:

$$g_n = \frac{P_n}{1-\sum_{i=1}^{n-1} P_i} \qquad (1)$$

Where $\sum_{i=1}^{n} P_i = 1 - P_L$ and P_L is the normalized power registered in the load, in this case 5% of power consumed in the load was assumed, n is the slot number and $P_n = (a_n)^2$ is the normalized power radiated by each radiator slot [8]. Programming Eq. (1) in Mathcad, and then normalizing the conductance, the results shown in Fig. 4 are obtained and summarized in Table 1. As expected, an asymmetric function with a more pronounced decay in the slots close to the source is observed, since the powers radiated by each slot depend on the given amplitude distribution that is symmetric as shown in Fig. 2. But in the slotted waveguide arrays with load at its end the power that reaches each slot n is the total input power minus the power consumed by the previous slots (n-1). Consequently, to maintain the target radiation level, slot *n* must have a higher conductance and thus a larger inclination angle.

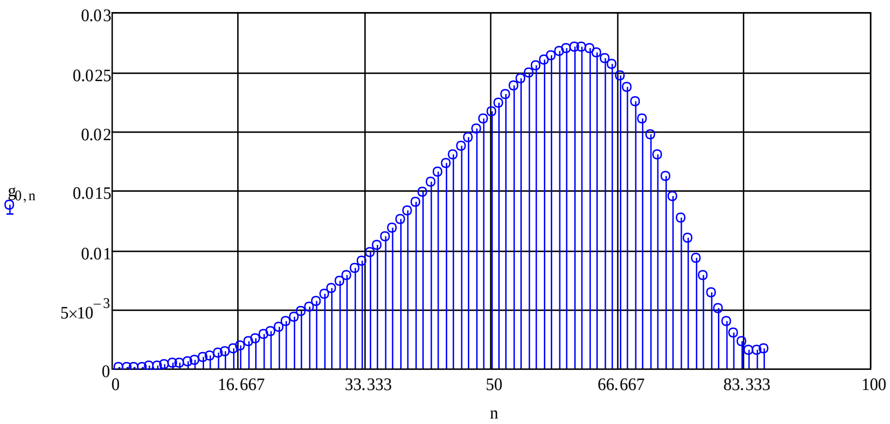

Figure 4: Normalized conductance of the slots.

In 1946, Watson obtained experimental results of the characterization of the slots in waveguides under laboratory conditions [10]. Within them are results for resonant arrays of slots on the narrow face of the WR-90 waveguide at a

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frequency of 9.375 GHz. He defined through experiments the incremental conductance caused by the effect of mutual coupling between identical and equally spaced slots at λg/2. It must be remembered that the curves obtained by Watson are valid for resonant arrays. For traveling waveguides arrays, the relationship between penetration and the angle of inclination can be obtained through Eq. (2) [1] that is obtained from the data of the family of simulations of the angle variation for a set of uniforms arrays of 71 slots modeled in HFSS.

$$\theta = -900.7 g_n^{\ 2} + 237.6 g_n + 2.0 \qquad (2)$$

Programming Eq. (2) in Mathcad, the results shown in Fig. 5 and summarized in Table 1 are obtained.

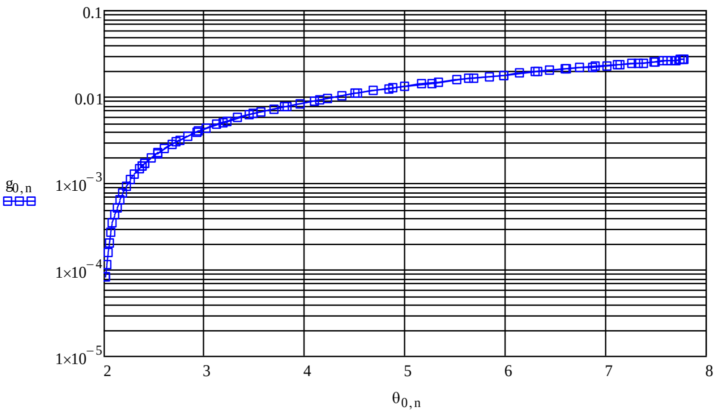

Figure 5: Normalized conductance vs. inclination of slot (θ in degrees).

Resonance Length of the Slot. Calculation of Penetration

In most of the bibliography consulted [1,11, 12] the value of $0.4625\lambda_0$ is assumed as the resonance length of the slot for the WR-90 waveguide at a frequency of 9.375 GHz with good results. This resonance length value was proposed by [11], but in this study the power of the evanescent modes is not taken into account. However, in a study carried out by [12], which incorporates the internal power of the evanescent modes in the waveguide into the impedance formula, a more accurate resonance length value is given. In this study the resonant length is approximately $0.545\lambda_0$. Furthermore, it is argued that the resonance length does vary with respect to the angle, unlike that stated by Stevenson [13] who states that the resonance length is fixed and with a value of $\lambda_0/2$. In the study carried out by [12] it is assumed, in the derivation of the theoretical results for the admittance, that the guide walls are a perfect conductor of negligible width, that the slot is narrow and that only the TE_{10} mode propagates. Penetration into the wide face is neglected and slot is assumed as a rectangle in one plane.

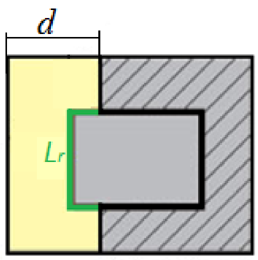

Figure 6: Resonance length.

From the resonance length value of the slot: $L_r = 0.545\ \lambda_0$, from the antenna data and based on Fig. 6, the penetration formula d_n for the nth slot is obtained in Eq. (3):

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$$d_n = \left(\frac{0.545\lambda_0 - \dfrac{b}{\cos(\theta)}}{2} \right) + 1.75mm \qquad (3)$$

The penetration results for each slot as well as the other parameters obtained in the design are summarized in Table 1. The model of the traveling wave SWA is shown in Fig. 7.

Table 1. Design of the SWA 86 elements for Taylor one parameter.

Coefn	g_n	θ(deg)	d(mm)	Coefn	g_n	θ(deg)	d(mm)
0.096	0.00008	2.019	5.354	1	0.017	5.837	5.331
0.114	0.00011	2.027	5.354	0.997	0.018	5.988	5.33
0.133	0.00015	2.037	5.354	0.992	0.019	6.145	5.3
0.153	0.00020	2.05	5.354	0.986	0.02	6.292	5.32
0.174	0.00027	2.064	5.354	0.977	0.02	6.444	5.325
0.197	0.00034	2.082	5.354	0.967	0.021	6.592	5.324
0.22	0.00043	2.103	5.354	0.955	0.022	6.734	5.322
0.244	0.00053	2.126	5.354	0.941	0.022	6.869	5.321
0.269	0.00064	2.154	5.354	0.925	0.023	7.005	5.319
0.295	0.00078	2.185	5.354	0.908	0.024	7.14	5.318
0.322	0.00093	2.221	5.354	0.89	0.024	7.262	5.316
0.349	0.00109	2.26	5.354	0.87	0.025	7.369	5.315
0.376	0.00128	2.304	5.353	0.848	0.026	7.48	5.314
0.377	0.00149	2.354	5.353	0.826	0.026	7.57	5.313
0.406	0.00171	2.406	5.353	0.802	0.026	7.647	5.312
0.434	0.00197	2.466	5.353	0.777	0.027	7.707	5.311
0.464	0.00224	2.528	5.353	0.751	0.027	7.76	5.311
0.493	0.0025	2.598	5.352	0.725	0.027	7.776	5.31
0.523	0.00286	2.672	5.352	0.697	0.027	7.777	5.31
0.553	0.00319	2.75	5.352	0.669	0.027	7.743	5.311
0.582	0.00355	2.833	5.351	0.64	0.027	7.687	5.312
0.611	0.00394	2.922	5.351	0.611	0.026	7.604	5.312
0.640	0.00435	3.018	5.351	0.582	0.026	7.49	5.314
0.669	0.00478	3.117	5.35	0.553	0.025	7.325	5.316
0.697	0.00525	3.224	5.35	0.523	0.024	7.121	5.318
0.725	0.00572	3.331	5.349	0.493	0.023	6.895	5.321
0.751	0.00623	3.446	5.348	0.464	0.021	6.61	5.324
0.777	0.00676	3.566	5.348	0.434	0.02	6.324	5.326
0.802	0.00731	3.691	5.34	0.406	0.018	5.981	5.33
0.826	0.0078	3.817	5.346	0.377	0.016	5.626	5.333
0.848	0.0084	3.952	5.345	0.349	0.015	5.262	5.336
0.870	0.0091	4.09	5.34	0.322	0.013	4.88	5.339
0.890	0.00974	4.229	5.344	0.295	0.011	4.504	5.342
0.908	0.01	4.375	5.343	0.269	0.009	4.142	5.344

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0.925	0.011	4.527	5.342	0.244	0.007	3.801	5.346
0.941	0.012	4.682	5.341	0.22	0.006	3.486	5.348
0.955	0.013	4.84	5.339	0.197	0.005	3.187	5.35
0.967	0.013	5	5.338	0.174	0.003	2.936	5.351
0.977	0.014	5.168	5.337	0.153	0.003	2.718	5.352
0.986	0.015	5.333	5.335	0.133	0.002	2.534	5.353
0.992	0.016	5.508	5.334	0.114	0.001	2.382	5.353
0.997	0.017	5.685	5.332	0.096	0.001	2.384	5.353
1	0.017	5.837	5.331	0.096	0.001	2.402	5.353

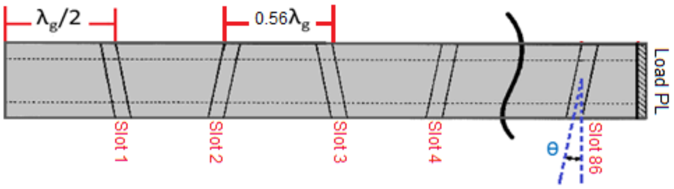

Figure 7: Final design model.

Considering the amplitude distribution of Taylor one-parameter and using the Array Factor (AF) for an even number of discrete elements [5], we obtain the function of the AF normalized in dB for the design of Table 1. The theoretical prediction model of the AF for the presented design is shown in Fig. 8. The SLL is verified well below -30 dB and the narrowness of the main lobe. The simulated model should roughly correspond to that predicted in this section.

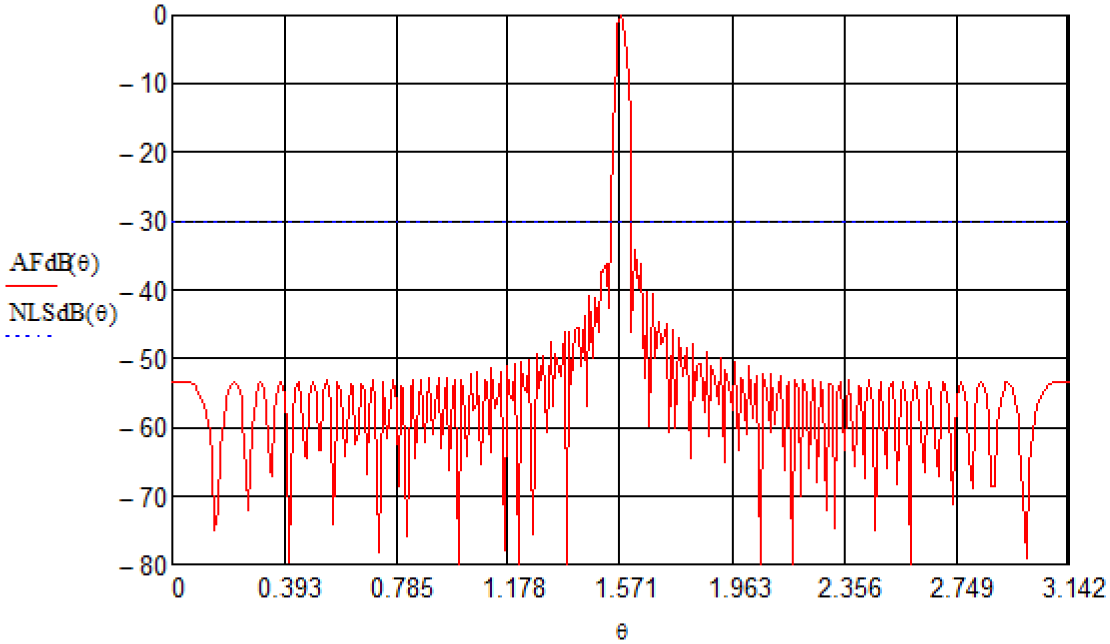

Figure 8: Function of the AF in dB normalized for the Taylor one-parameter distribution with 86 elements.

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3. SIMULATION OF THE COASTAL RADAR ANTENNA IN THE ANSOFT HFSS SOFTWARE

The reflector along the vertical plane to be joined to the SWA designed in the previous section is calculated through the optimal vertical plane horn antenna criterion. As a design requirement, an HPBW must be obtained for this plane of 26^0 and SLL less than -25 dB. The calculation of the dimensions was carried out in Mathcad and the results are summarized in Fig. 9.

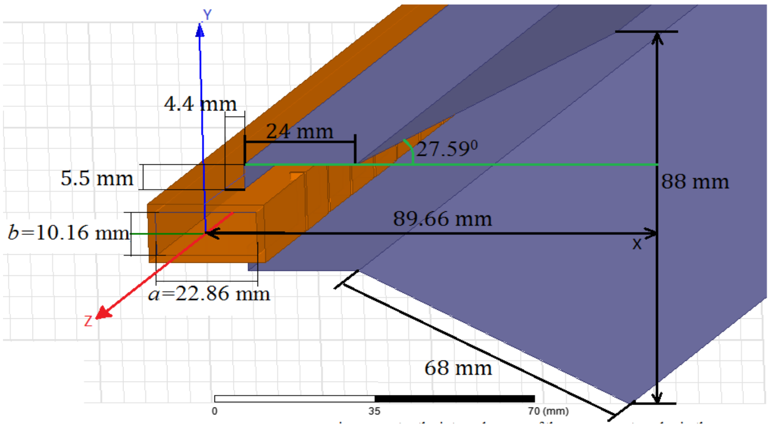

Figure 9: Reflector dimensions.

To obtain the electromagnetic characteristics of the structure of the coastal radar antenna, it is formed by the 86-element SWA and the reflector across the vertical plane. The antenna is simulated and implemented in the Ansoft HFSS Electromagnetic Software (EMS). Two radiation ports are defined in the simulation model on either side of the WR-90 waveguide to obtain the progressive wave condition and the relevant radiation and boundary conditions are assigned.

The model of the structure designed in HFSS is illustrated in Fig. 10. The details of the slots and the reflector as well as the length of the antenna are shown in Fig. 11.

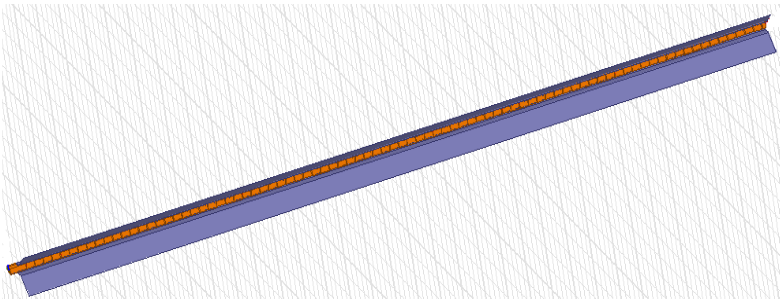

Figure 10: Model of the antenna designed in the HFSS.

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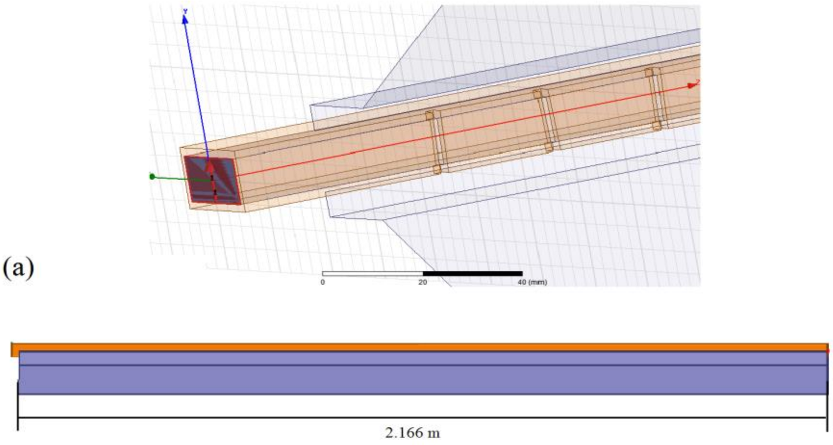

(a)

(b)

Figure 11: Designed antenna. a) Detail of the slot, b) Length of the antenna.

The graph in Fig. 12 shows the simulation results for the rectangular diagram of the far-field radiation pattern along the vertical plane (XY Plane). In this, the HPBW, indicated with blue markers, has a value of approximately 26^0 and corresponds to the design exigencies for this plane. The secondary lobe level is below -25 dB; result that corresponds to the typical characteristics of a radar antenna in a marine environment.

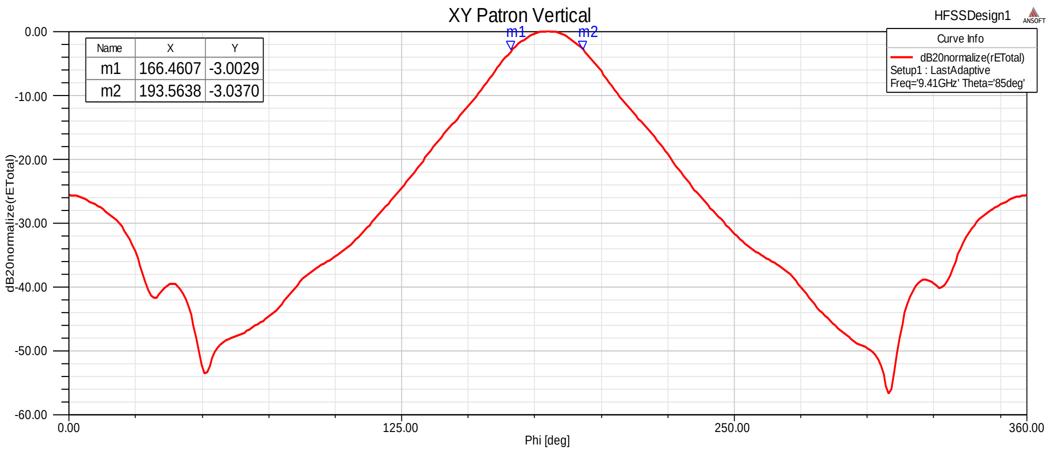

Figure 12: Radiation pattern of the antenna in the vertical plane.

Fig. 13 shows the co-polarized far-field radiation pattern of the antenna designed by the horizontal plane (Z-direction). The HPBW is approximately 0.6^0. This value corresponds to 0.7 ± 0.1^0. The resulting radiation pattern in the horizontal plane corresponds to a Taylor one-parameter amplitude distribution and fits the theoretical prediction model of the AF function, seen in section 2. A -30 dB line is indicated in the figure and the level of the lobes closest to the main indicate than SLL is below -30 dB, in correspondence with the design requirements. The main lobe is approximately 5^0 out of phase from the center because the phase law in the axis of the traveling wave SWA is not uniform.

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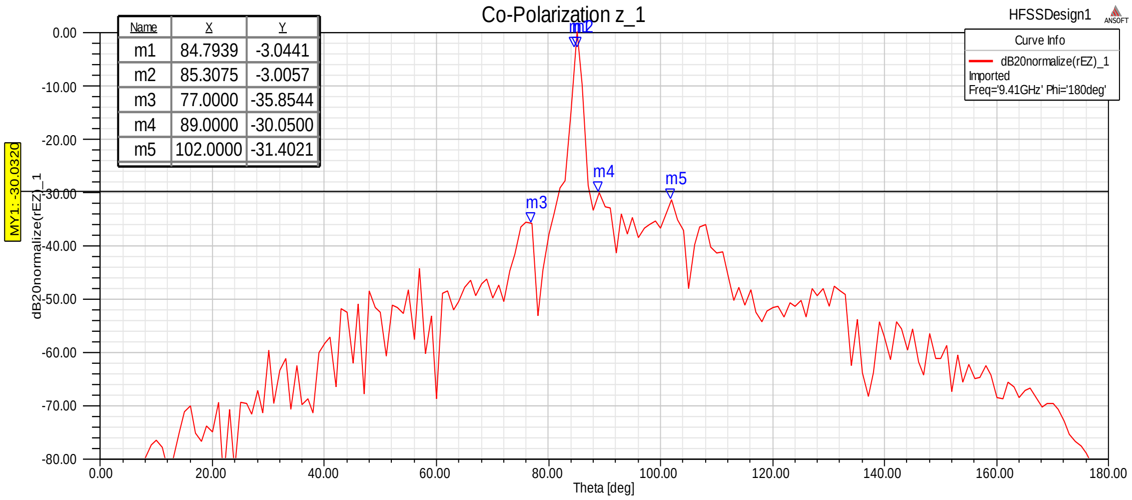

Figure 13: Co-polarized far-field radiation pattern in the horizontal plane.

In Fig. 14 the cross-polarized far-field radiation pattern is shown where the cross-polarized lobes are observed at a distance of approximately ± 45^0 from the main lobe. The particular characteristics of the cross-polarized pattern are the product of the number of slots (electrical length of the array) and the amplitude distribution (the inclination angles of the slots, type Taylor one-parameter).

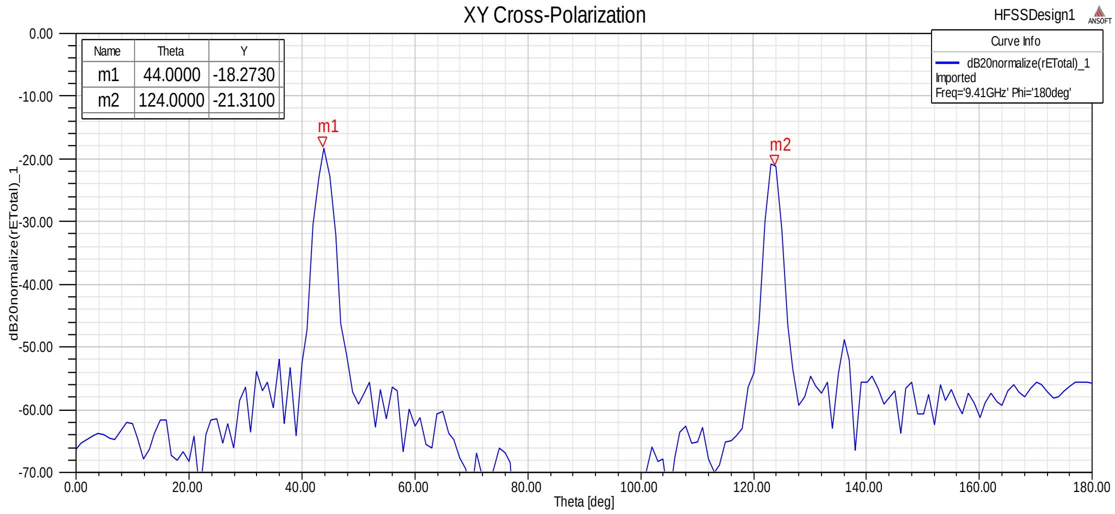

Figure 14: Cross-polarized far-field radiation pattern.

In order to reduce the cross-polarization lobes, a suppressor filter for this type of polarization is included in the design. For this, three variants of filters were designed as shown in Fig. 15: a rectangular waveguide filter (Fig. 15a), a grid with parallel metallization strips (Fig. 15b) and a similar metallic grid made on a layer FR4 laminate (Fig. 15c).

For the design of the grid filter with parallel conductive strips (Fig. 15b), copper strips are used, with a thickness of 0.5 mm. It is assumed for the magnitude of the distance between the strips $g = \lambda g/4 = 11.2mm$, and the width for each strip $2a = 2.2\ mm$. The strips will be placed at a distance of 16.22 mm from the slotted face of the waveguide.

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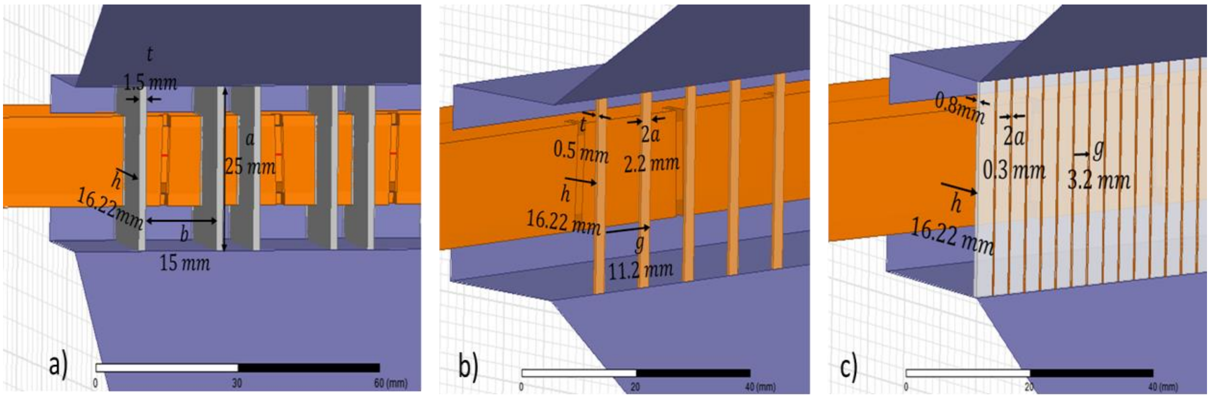

Figure 15: a) Rectangular waveguide filter designed, b) grid filter with parallel conductive strips designed, c) grid filter with parallel conductive strips on FR4 sheet.

Another variant of this same type of filter is a similar grid with metal strips manufactured on a laminated layer of FR4 (Fig. 15c). For this design, strips of copper with a thickness of 0,018 μm on a sheet of FR4 with a height of the substrate of 0.8 mm are proposed. The separation of the strips and the width 2a of each one, are assumed from the design values used and tested in [1] ($g = \lambda/10 \approx 3.2\ mm$ and $2a = \lambda/100 \approx 0.3m$m). The grid will be placed, in the same way as the previous variant, at a distance: h = 16.22 mm from the slotted face of the waveguide.

For the rectangular waveguide filter design, two parallel metal sheets are placed on each side of the waveguide slots as shown in Fig. 15a. These plates, when inserted into the structure that already has the antenna, form with the upper and lower sheets of the existing structure, small rectangular waveguides within which each slot will be found. These small waveguides that are formed with height a, width b, and depth h, must be designed in such a way that they cause all propagation modes except the one corresponding to the main or co-polar polarization to be cut off [14-16].

In this case, the cutoff of the TE_{01} mode must be ensured to reduce the vertical electric field component (cross-polarization), for this it must be fulfilled that $b < \lambda/2 = 15.94mm$, so a value of 15mm is chosen for b. To guarantee the propagation of the TE_{10} mode, which corresponds to the distribution of the horizontal component of the electric field (co-polarization) of each slot in the small rectangular waveguides formed by the filter, it must be satisfied that $a > \lambda/2 = 15.94mm$. In accordance with this requirement, a value for a of 25 mm is chosen, placing the plates in the area before the flared structure of the antenna as shown in Fig. 15. The thickness (t) of the metal plates of the filter is assumed equal to 1.5mm and h = 16.22mm is taken.

By simulating the antenna with the three proposed filter variants and by comparing the cross-polarization patterns obtained, several aspects can be concluded: 1) the filter that best reduces the level of cross polarization, approximately 33dB, is the rectangular waveguide filter and 2) the parallel conductive strips filter was the one that least reduce the level of cross-polarization lobes and this reduction is lower than -30 dB (initial requirement of design). However, with the rectangular waveguide filter, the maximum gain reached by the antenna, in the maximum direction of radiation, decreases below 30 dB and the level of side lobes of the co-polarization radiation pattern increases above -30dB; also this last design requirement is also affected by the use of the parallel conductive strip filter on the FR4 sheet. Therefore, in order to reduce the level of cross polarization but maintaining the other design requirements, the filter that produces the best results is made up of parallel conductive strips without the FR4 sheet. Taking into account the results obtained, the filter with parallel conductive strips is chosen to include it in the final design of the antenna, but a parameterization of the width of the strips is carried out, to choose the most optimal value for the filter design that achieves the reduction of the cross polarization level below -30 dB.

When performing a parametrization of the width of the strips 2a (1.2mm-10.2mm, with a step of 1mm) in the HFSS, it is obtained that the behavior of the cross polarization level decreases with the increase in the width of the filter strips. Its obtained from a strip width of 4.2mm cross polarization levels lower than -30dB and with much better

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values than those reached by the filter made on the FR4 sheet. When the width of the strips of the filter is 7.2mm it is also observed that the level of co-polarization side lobes is not affected. Therefore, this value will be chosen for the final design of the filter. With the dimensions chosen for the final design of the filter, the cross polarization pattern shown in the Fig. 16 is obtained and it is observed that a maximum cross polarization level of -41.4678 dB is reached (<-30 dB, requirement of design). The maximum level obtained with the antenna without the filter is reduced by 23dB.

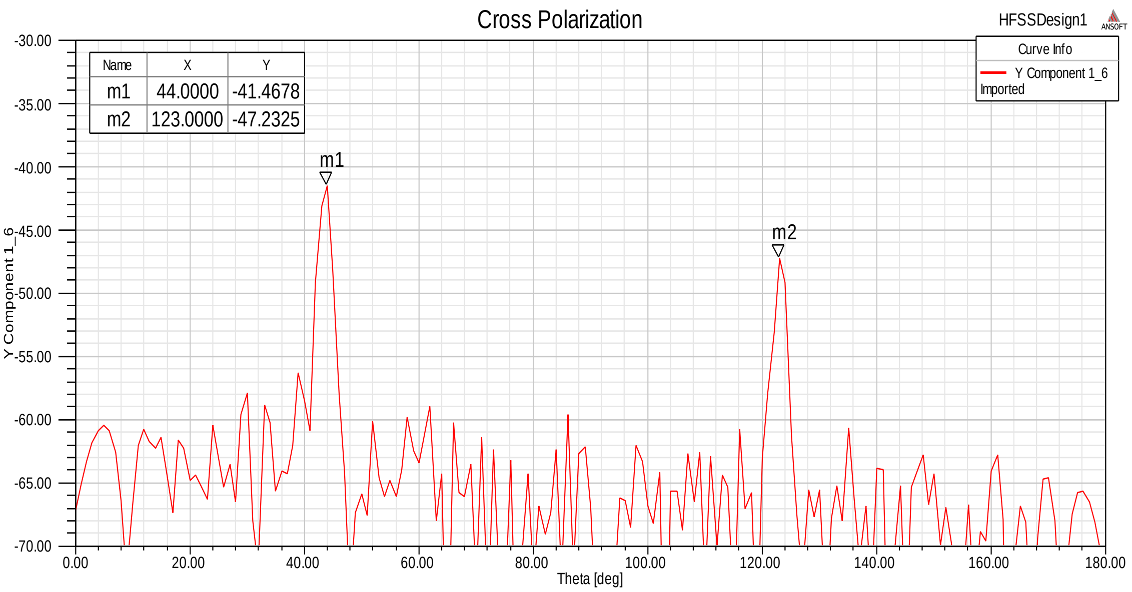

Figure 16: Cross-polarized far-field radiation pattern obtained with the designed filter.

The 3D Gain pattern is illustrated in Fig. 17. It is seen how, in the direction of maximum radiation, is reached 31.051 dB of gain (maximum value). The final objective of achieving a gain > 30 dB is thus met.

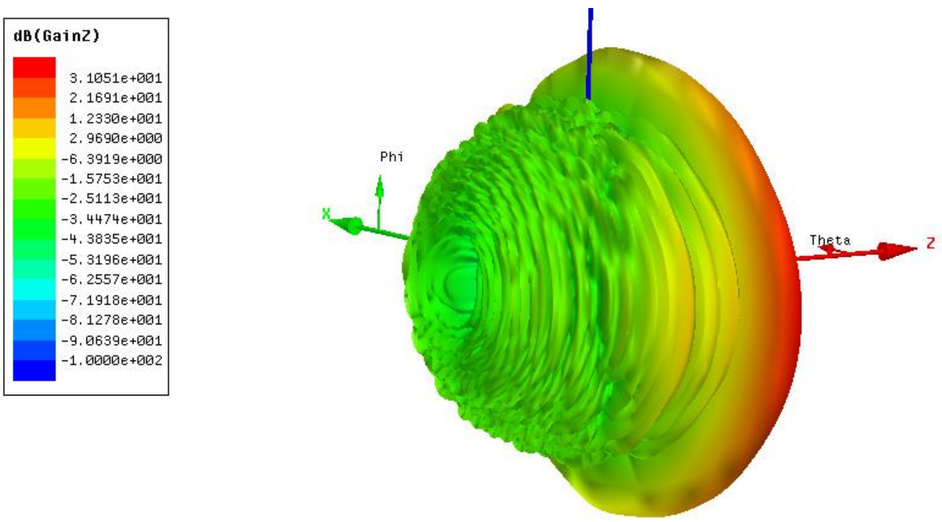

Figure 17: Antenna gain 3D radiation pattern.

In Fig. 18 the parameter S_{11} of the designed antenna is shown. The typical characteristic of a traveling wave SWA is observed. At 9.41 GHz, S_{11} = -42 dB and in a 2 GHz interval around the center frequency S_{11} remains <-23dB.

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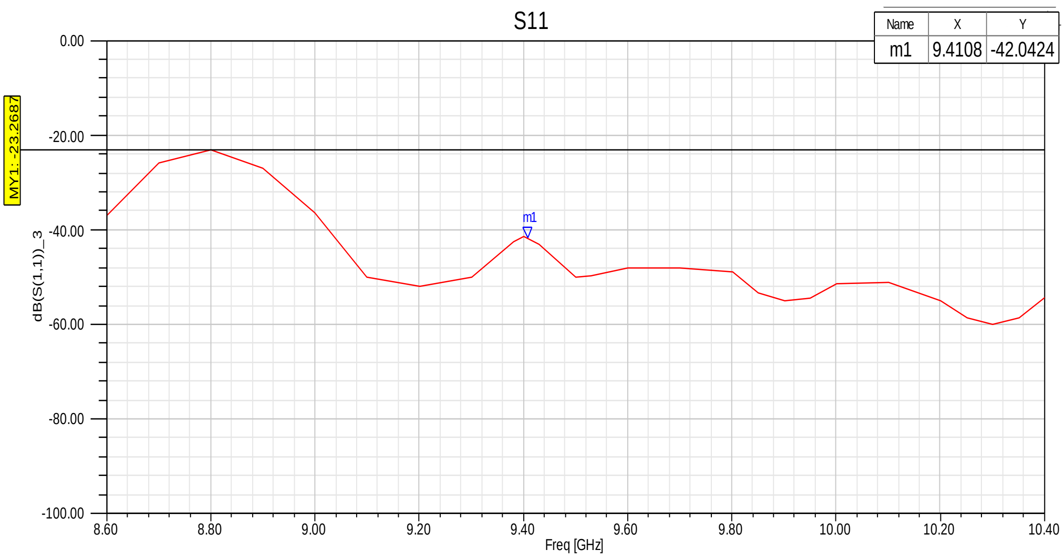

Figure 18: S_{11} parameter of the designed antenna.

4. CONCLUSIONS

A methodology for designing traveling wave slotted waveguide antenna with slots on the narrow face of the WR-90 waveguide is provided. The design, separately in both planes, of a coastal radar antenna consisting of a system formed by a traveling wave SWA in the horizontal plane and a reflector in the vertical plane was provided. The design of the slotted waveguide is carried out from a Taylor one-parameter amplitude distribution along the axis of the array with a parameter of -30 dB of SLL. A formula is proposed to determine the penetration of the slots based on the criterion that the resonance length of the slots is 0.545 λ_0. The design of the reflector by the vertical plane is carried out according to the criterion of obtaining the radiation pattern desired by that plane. In order to reduce the cross-polarization lobes, a grid filter with parallel conductive strips is included in the design.

The results obtained in the simulation of the final antenna of the coastal radar correspond to the analytical models. A value of 26^0 of half power beamwidth are obtained from the vertical plane, while from the horizontal, 0.6^0 are obtained, in this way initials requirements are satisfied. Secondary lobe levels on the horizontal plane are below -30 dB and on the vertical plane are less than -25 dB, coinciding with radar data for naval applications. The cross polarization level reached is -41.48 dB. The final system gain is 31.051 dB. The S_{11} parameter in the band of interest is below -23 dB. The results are classified as satisfactory.

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CONFLICT OF INTERESTS

None of the authors stated the existence of possible conflicts interest that should be declared in relation to this paper.

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Melissa Vega Arias: Conceptualization, data Curation, formal analysis, research, methodology, project management, resources, software, validation-verification, visualization, drafting-original and reviewing and editing.

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Pedro Arzola Morris: Conceptualization, data curation, formal analysis, fund acquisition, research, methodology, project administration, supervision.

José R. Sandianes Galvez: Conceptualization, data curation, formal analysis, research, methodology, supervision, suggestions and formation of the final writing.

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