A Continuous Beam Steering Slotted Waveguide Antenna Using Rotating Dielectric Slabs

Amirhossein Ghasemi and Jean-Jacques Laurin, Senior Member, IEEE

Abstract—The design, simulation, and measurement of a beam steerable slotted waveguide antenna operating in X band are presented. The proposed antenna consists of a standard rectangular waveguide (RWG) section with longitudinal slots in the broad wall. The beam steering in this configuration is achieved by rotating two dielectric slabs inside the waveguide and consequently changing the phase of the slots excitations. In order to confirm the usefulness of this concept, a nonresonant 20-slot waveguide array antenna with an element spacing of \( d = 0.6\lambda \) has been designed, built, and measured. A 14° beam scanning from near broadside \((\theta = 4^\circ)\) toward end-fire \((\theta = 18^\circ)\) direction is observed. The gain varies from 18.33 to 19.11 dBi and the antenna efficiency from 95% to 79%. The sidelobe level is \(-14\) dB at the design frequency of 9.35 GHz. The simulated and measured co-polarized realized gains are in good agreement.

Index Terms—Beam steering, rotating dielectric slabs, slotted waveguide antenna.

I. INTRODUCTION

Many types of beam scanning antennas have been designed and developed in the past. In particular, radar sensors for automotive applications use antennas with agile beams, either by mechanically moving the complete antenna or by electronically scanning the beam of a fixed antenna [1]. Electromechanical beam scanning offers a low-cost solution, but it is less agile compared to electronically steerable antennas. However, it can potentially be used at higher peak power levels because no electronic parts are used [2]. Special attention was paid to the concept of a fixed antenna in which small parts are moved [3]. Many such concepts have been presented in [4]–[8], as candidates for millimeter wave scanning antennas and in [9]–[11] for satellite communications. The Eagle Scanner of World War II was possibly the first electromechanical beam scanning antenna and included the displacement of the narrow wall of a rectangular waveguide (RWG) [12] in order to introduce phase shifting in a series-fed linear array. However, the realization of precise and simultaneous linear motion of the side wall over the total length of the antenna is difficult to achieve.

Recent works on reconfigurable antennas have provided an alternative solution to perform the beam steering with less complexity [13]. Such a reconfigurable antenna which uses a movable dielectric plunger within the feedline of a traveling wave array has been presented in [14]. Although a large scanning range is reported, the measured return loss is low (7.5 dB), and the movable part is bulky.

Frequency scanning antennas based on series-fed arrays or leaky-wave structures may possibly offer a simple and low-cost solution compared with phased arrays. However, frequency variation over a wide bandwidth to scan the main beam is required [15], [16], which is not suitable for the narrow bandwidths generally allocated to certain applications, such as X-band weather radars.

In order to simplify the requirements of the beam scanning mechanism, a new concept is proposed, which consists of a waveguide linear slot array in which beam scanning is achieved by rotating two dielectric slabs inside the RWG. The configuration is inspired by an earlier design of a waveguide phase shifter presented in [17]. In this design, the wavelength of the traveling wave is changed depending on the angle of the slabs relative to the dominant mode field.

Although this concept has been first presented in [6] and worked out in [1], this work exhibits more scan range with lower sidelobes due to symmetric field distribution inside the waveguide. Furthermore, a matched transition from a coaxial port to the waveguide filled by the rotating slabs is presented.

This paper is organized as follows. Section II introduces the design methodology of the slotted waveguide antenna. In Section III, engineering details on the design method are given. Finally, the simulation and experimental results are provided and discussed in Section IV.

II. DESIGN OF THE SLOTTED WAVEGUIDE ANTENNA

A. Rotating Dielectric Slabs Inside the Waveguide

The experimental evidence of the scanning capabilities realized by rotating a single ridge in a metal waveguide was given by Solbach and Demirel [1]. Such configuration allows the propagation of a TEM mode. The longitudinal slots in an RWG are not excited by the TEM mode because the current flow is not perpendicular to the slots. However, at the feeding transition, modes can couple, and therefore, affect the return loss. Also, the power in the TEM mode will be dissipated in the terminating load, resulting in reduced antenna efficiency. In [1], another drawback was the presence of high sidelobes due to the asymmetric field distribution inside the waveguide. The asymmetry of the structure can be observed in Fig. 1(a), showing the nonsymmetric excitation of the slots on the RWG when the position of the single slab is...
oblique. To avoid the asymmetric coupling, we propose using two symmetrically positioned rotating slabs in the waveguide, as shown in Fig. 1(b). Also, the slabs are made of dielectric instead of metal in order to eliminate the TEM mode.

The dimensions of the waveguide and the dielectric slabs have to be chosen to ensure single-mode propagation at the desired frequency of operation and to maximize the variation of the mode propagation constant with the rotation of the slabs. In order to simplify the fabrication, a standard WR-62 waveguide made of aluminum was used. Fig. 2 shows the propagation constant versus frequency for two different dielectric slabs with relative permittivities of 6.15 and 10.2. Since the cutoff frequency of the second mode for the case $\varepsilon_r = 10.2$ is close to the design frequency of 9.35 GHz, the dielectric with $\varepsilon_r = 6.15$, corresponding to RT/Duroid 6006 from Rogers, is preferred. The dimensions are given in Table I. The values of $a_{slab}$ and $d_{rotation}$ were varied in simulations to maximize the variation of $\beta_g$. The curves of Fig. 2 were obtained with the 3-D full-wave solver of Ansys-HFSS, with no slots on the waveguide. The cutoff frequency of the air-filled WR-62 waveguide is 9.49 GHz. In Fig. 2(b) we see that it varies between 7.6 and 8.5 GHz for the fundamental mode, depending on the rotation state of the slabs. It can be noted that at the chosen operation frequency of 9.35 GHz applicable to weather radars, the 2nd mode is well attenuated and the waveguide supports only one mode for all the rotation states of the slabs when $\varepsilon_r = 6.15$.

### Table I

| Dielectric Slab and Waveguide Parameters |
|------------------------------------------|
| $a_{guide}$ | $b_{guide}$ | $a_{slab}$ | $b_{slab}$ | $t$ | $\varepsilon_{slab}$ | $\tan \delta_{slab}$ | $d_{rotation}$ |
| 15.8mm | 7.9mm | 5mm | 2.5mm | 1mm | 6.15 | 0.0019 | 4.45mm |

### B. Nonresonant Array With the Slots Alternatively Displaced

The two basic types of slotted waveguide antennas are the resonant array (standing wave) and the nonresonant array (traveling wave). Since we wish to scan the beam off-broadside, the case of interest is the nonresonant array. For that case, the slot-to-slot spacing differs from $\lambda_g/2$, so the reflections from the different slots do not add up in phase at the input of the waveguide, leading to a small reflection coefficient. Thus, the aperture distribution experiences a phase progression that is uniform, or nearly so, which is why these arrays are also referred to as traveling wave fed arrays [20]. Nonresonant arrays include a matched-load termination, necessary to avoid undesirable reflection causing a back-lobe responsible for the degradation of the antenna pattern. The main advantage of these arrays is a larger bandwidth in terms of sidelobe level (SLL) and input matching, which makes them suitable for performing as beam scanning antennas [18]. In the nonresonant slotted waveguide antenna, the slot spacing can be chosen so that we can produce the main lobe at almost any arbitrary angle $\theta$ relative to the axis of the array. If offsets are alternated on opposite sides of the symmetry plane, then the array factor is given by [19]

$$AF = \sum_{n=1}^{N} a_n e^{jn(\beta_0d\sin\theta - \beta_g d + \pi)}$$

where $a_n$ is the slot excitation amplitude level, $d$ is slot spacing, $\beta_g$ is the guided phase constant, and $\beta_0$ is the propagation constant in free-space. In order to have a good aperture efficiency, it is desired to scan the beam near broadside. Therefore, a nonresonant array with the slot spacing $d$ is different but close to $\lambda_g/2$ is necessary. The element spacing of $dl/\lambda_g(\theta_1 = 45^\circ) = 0.4$ was chosen because array factor calculations based on (1) and using the $\beta_g$ values given in Fig. 2 for the design frequency of 9.35 GHz led to maximum beam deviation and minimum SLL. By rotating the dielectric
slabs from vertical to horizontal position, the wavenumber variation observed from Fig. 2 is

$$\beta_g(\theta_1 = 0^\circ) - \beta_g(\theta_1 = 90^\circ) = 45\text{rad/m}. \quad (2)$$

This gives 14° of beam scanning based on (1). A 20-slot waveguide was chosen in this study based on limitations of the fabrication means available to the authors. A triangular amplitude distribution over a pedestal was used in order to demonstrate the capability to control the SLL.

If the relative excitation level of the nth slot is $a_n$, the power $P_n$ radiated by this slot will be proportional to

$$a_n^2 (P_n = ka_n^2) \quad [19].$$

Let $r$ be the fraction of the incident power to be dissipated in the match load. The equivalent circuit for the array is shown in Fig. 3. Assuming a 1 W power input and a lossless waveguide, we must have $P + \sum_{n=1}^N P_n = 1$. According to the nonresonant array design method presented in [19], we have $g_n = (P_n)/(r + \sum_{i=n}^N P_i) = (P_n)/(1 - \sum_{i=1}^N P_i)$. Fig. 4 shows the values of distribution amplitude ($a_n$), radiated power ($P_n$), and normalized conductance ($g_n$) for a triangular taper excitation of a 20-element array, with $r = 0.15$. This is larger than typical values, but it has been chosen due to the limitation on the antenna length coming from fabrication and test equipment available to the authors. Reducing $r$ for the same antenna length would require larger $g_n$ values. This may cause multiple reflections between slots and make the array design method inaccurate. A larger $r$ value was, therefore, used to illustrate the proposed antenna concept.

C. Evaluation of Offset and Length of the Slot

We compare two different methods to analyze the dielectric-loaded slotted waveguide. The effect of mutual coupling between the slots is taken into account in both methods.

One way to consider the mutual coupling is to simulate an array of several coupled slots and extract their characteristics. For radiating slots, a design curve of resonant length and offset from the waveguide center plane against slot admittance is derived to give the slot resonant offset ($x_r$) for each slot. The input admittance $y_{in}$ is the sum of all the admittances ($y_{in} = \sum y_i$) due to the spacing of $\lambda_g/2$ between the elements. A linear relationship between the admittance ($y_{in}$) and the number of slots is obtained for an array of five identical slots in the simulations.

1) Method 1: Resonant Array: Although we are considering nonresonant scanning arrays, the resonant array case is useful to estimate the admittance of the slots. As Fig. 5 illustrates, the slots of resonant arrays are spaced by $\lambda_g/2$ so that all of them share the same phase excitation necessary to produce a broadside main lobe. Typically, a short circuit is placed at the end of the waveguide, at a distance of $\lambda_g/4$ after the last slot. The input admittance $y_{in}$ is the sum of all the susceptances ($y_{in} = \sum y_i$) due to the spacing of $\lambda_g/2$ between the elements. A linear relationship between the admittance ($y_{in}$) and the number of slots is obtained for an array of five identical slots in the simulations.

2) Method 2: Two-Slot Nonresonant Array Using Periodic Boundary Conditions: In this method, we use periodic boundary conditions to simulate a large array of mutually coupled slots, as described in [20]. This method also assumes a uniform amplitude excitation, but it is still helpful to design the proposed array because the change in the excitation coefficients is gradual, as shown in Fig. 4. The periodic boundary condition is implemented using master–slave boundaries in Ansys-HFSS. Fig. 6 illustrates the two-slot nonresonant array with periodic boundary walls. In this method, the value of $d$ is not necessarily $\lambda_g/2$. We have used $d = 0.4\lambda_g$ because it corresponds to $\theta_1 = 45^\circ$, which is in the middle of the range.

For both methods, we need to find the resonant length ($l_r$) for each given slot offset ($x_i$) for each slot. This was accomplished by running a parameter sweep with HFSS for each $x_i$. The normalized susceptance ($b$) of the slot must be zero at resonance. The parameter sweep returns a set of admittance points that was processed in order to determine $l_r$ (i.e., corresponding to zero susceptance) for a given slot offset.

For that, we used an interpolation routine implemented in MATLAB to generate a contour corresponding to the resonant condition in the length-offset plane for each method. This contour is shown in solid curves in Fig. 7. Then, for each point on this contour, we can determine the corresponding conductance ($g$) at resonance (see Fig. 7). At the end of this process, two interpolation polynomials for each method were derived to give the slot resonant offset ($x_r$) and $l_r$ versus required resonant slot conductance. These interpolation curves are illustrated in Fig. 8.
Fig. 6. Physical form and equivalent circuit model of the nonresonant array with periodic boundary condition.

Fig. 7. Slot admittance ($y$) versus slot length ($l$) and slot offset ($x$), obtained for $\theta_1 = 45^\circ$.

Using the values of conductance from the triangular distribution given in Fig. 4 and the polynomials shown in Fig. 8, we can obtain the resonant parameters of the antenna.

Table II summarizes the final antenna parameters calculated for both methods. As can be seen in Table II, the difference between the resonant lengths obtained with the two methods is very small. The largest difference is 0.32 mm (about 2%), and it occurs for $n = 1$ where amplitude weight ($a_1$) is very small. As for the slot offsets, the largest difference observed at $n = 11$ and $n = 13$ is 0.19 mm (about 13%), where amplitude weights are large. The impact of these differences on the antenna pattern is shown in Section II-D.

D. Array Simulation

In order to validate the design approach, the values of $g_n$ in Table II were used in a simulation of the circuit model shown in Fig. 3 (using ADS® from Keysight). The complex currents ($i_n$) in the shunt conductances representing the slots were used to calculate the array factor with

$$ AF_{cl} = \sum_{n=1}^{N} i_ne^{jn(\beta_d l \sin \theta + \pi)}. $$

This procedure takes into consideration the propagation and the multiple reflections between the loads (i.e., the $g_n$’s) in the transmission line network. The resulting array factors are plotted in Fig. 9 (solid curves), along with the normalized H-plane patterns of the array with the various slab rotation angles simulated with HFSS. Simulations are shown for designs based on the two sets of slot parameters given in Table II. The beam steering of $14^\circ$ predicted by the variation of $\beta_g$ is nearly achieved in both cases. The two full-wave results are very similar in terms of beamwidth and SLL. Looking at the three cases of slab rotation shown in Fig. 9, there is no clear advantage of one method over the other. The fabricated prototype presented in Section IV used the values of $x_n$ and $l_n$ obtained with Method 1. The theoretical and simulated SLLs are different in Fig. 9. This is because the HFSS simulation considers structural scattering (e.g., waveguide edges) and coupling between elements outside the waveguide, whereas array factor assumes ideal noninteracting isotropic elements in free space.

III. DESIGN GUIDELINES

A. Matched Load Design

Nonresonant slotted waveguide arrays require a matched load termination. In this section, the design of a matched
load for the dielectric loaded waveguide used in the antenna is proposed. A standard RWG matched load cannot be used since our waveguide is loaded with dielectric slabs. However, it is quite straightforward to design a matched load by inserting absorber material. At this end, we have chosen the DD-10214 Silicon from ARC Technologies Inc., with \( \varepsilon = (17 - 0.2)i \varepsilon_0 \) and \( \mu = (1.6 - 1.8)i \mu_0 \) at the operation frequency of 9.35 GHz. Absorber strips with a thickness of 0.762 mm were placed on one of the broad walls of the waveguide. Based on this information, a load for WR-62 with rotating dielectric slabs has been designed. As illustrated in Fig. 10, a stand made of Teflon was used to support the slabs, and since it is located after the absorber, it is not causing reflections. Return loss greater than 25 dB was obtained in simulations over a wide frequency range for all the rotation angles of the dielectric slabs (see Fig. 10).

### B. Waveguide Transitions

The cutoff frequency of the air-filled WR-62 waveguide is 9.49 GHz. Since our operation frequency for the RWG loaded with dielectric slabs is 9.35 GHz, it is required to reduce the cutoff frequency of the waveguide feeding the antenna. This was realized with a metallic ridge, as shown in Fig. 11, which reduced the cutoff frequency to 7.8 GHz. A coaxial to ridge waveguide transition was designed. The size of the coaxial probe and its location with respect to the shorting wall were varied to optimize matching. The simulated return loss of more than 20 dB was achieved in the frequency band of interest, as shown in Fig. 11. Then, a second transition, this time from the ridge waveguide to a WR-62 waveguide loaded with dielectric slabs was designed. Wedge-shaped tips on both the ridge and the dielectric slabs led to good simulated return loss for all rotation angles of the slabs, as shown in Fig. 12 for three cases.

### C. Directional Flare

In order to improve the gain of the antenna, a low-profile flare section was added on the broad wall radiating surface of the waveguide (Fig. 13). This generates a narrower fan beam and brings the direction of maximum radiation in the scanning plane. The dimensions of the directional flares were varied in HFSS simulations in order to achieve maximum antenna gain. The optimal values of \( H \) and \( \alpha \) are shown in Fig. 13. The realized gain increased by about 5 dB (\( \theta_i = 0^\circ \)) with the addition of the flare, as shown in Fig. 14. This plot uses a linear scale to better illustrate the shift of the maximum gain direction in the \(xz\) plane. It should be noted that, if a much larger value of \( H \) is used, a horn is formed and the gain can be increased.

### IV. Prototype of the Antenna and Measurement

An antenna prototype has been built using the WR62 standard waveguide according to the detailed design given in Sections II and III. The 20 slots have been milled onto the upper broad wall in accordance with the dimensions of the resonant parameters given in Table II. The coaxial to waveguide transition and the matched-load described in Section III were implemented. Fabricated prototype and drawing model of the
antenna are illustrated in Fig. 15. The dielectric slabs can be seen at the end of the waveguide. In our test, they were manually rotated, but adding a controlled motor is an easy task. The top wall is attached to the waveguide narrow walls with one screw at every 10 mm on each side of the slots. In total, 163 screws were used in the presented prototype. It should be noted that although the antenna exhibits continuous beam steering with continuous rotating of the dielectric slabs, measured patterns are only shown here for $\theta_1 = 0^\circ$, $45^\circ$, and $90^\circ$.

A. Impedance Matching

Measurement of the return loss has been made on the antenna coaxial port for three positions of rotating slabs. The results are shown in Fig. 16, together with the simulations. The antenna was designed for a frequency of 9.35 GHz and $\theta_1 = 45^\circ$. We can see that return loss is greater than 10 dB for the three rotation states. A good agreement between measured and simulation $S_{11}$ parameters is obtained for three slab rotation states over the frequency band of 9–10 GHz (BW = 10.5%).

The power lost in the terminating load was not measured due to the nonstandard waveguide at the end of the antenna, but the $S_{21}$ parameters were simulated with HFSS and are also shown in Fig. 16. In the simulations, port 2 replaces the matched termination at the end of the waveguide. The array was designed by assuming dissipation of 15% ($-8.2$ dB) of the input power in the terminating load. We can see in Fig. 16 that...
the simulated $S_{21}$ at the design frequency of 9.35 GHz for the three rotation states is less than $-10$ dB, which is lower than the assumption.

**B. Radiation Patterns**

The radiation properties of the antenna prototype were simulated with HFSS and have been measured in an anechoic chamber. H-plane patterns of the realized gain at the design frequency of 9.35 GHz are presented in Fig. 17. A good agreement is obtained between simulations and measurements for the three scanning states. As expected, the main beam steering for $\theta_1$ varying from $0^\circ$ to $90^\circ$ is around $14^\circ$. The measured and simulated tilt angles and beamwidths are slightly different, but the differences are more visible on the sidelobes, especially at $\theta_1 = 0^\circ$. As seen in Fig. 9, the triangular tapered distribution applied is leading to an SLL of about $-25$ dB in theory. However, both simulated and measured patterns have higher SLLs. The discrepancies between measurements and simulations have multiple causes. This includes the mechanical errors during fabrication and assembly of the antenna structure, inaccurate setting of angle $\theta_1$ in the experiments, inaccuracies of the specified material properties for the dielectric slabs, matched load absorbers, etc. A detailed sensitivity analysis would be required to understand each of these effects, but this has not been considered in this work.

The cross-polarization level in the H-plane has also been measured and plotted in Fig. 17. According to the measured patterns shown in Fig. 17, the cross-polarization gain X-pol is approximately 25 dB below the co-polarized (Co-pol) gain in the main beam direction. In the simulations (not shown), this Co-pol to X-pol ratio was 35 dB. The difference is possibly due to limited fabrication tolerance and misalignment errors during measurements.

**C. Antenna Gain**

The realized gain of the antenna for three rotating angles has been measured by the gain comparison method in the anechoic chamber. This gain is shown in Fig. 18(a). The antenna efficiency based on simulated realized gain and directivity is shown in Fig. 18(b). Higher efficiency is obtained for the larger slab angles. Simulations revealed that the slab-loaded waveguides have attenuation constants of 1.4, 1.2, and 0.86 dB/meter for slab angles of $\theta_1 = 0^\circ$, $45^\circ$, and $90^\circ$, respectively, at the design frequency. The maximum in gain and efficiency are close to the target frequency of 9.35 GHz for $\theta_1 = 45^\circ$. Fig. 19 illustrates simulated realized gain and main beam direction over a continuous range of rotation angles of $\theta_1$. 
Compared to the concept presented in [14], which is based on a dielectric loaded parallel plate waveguide, the proposed antenna offers less scanning range. However, the moving part realizing beam scanning is much lighter, and the measured return loss performance is better. Both antennas have similar SLL performance. The scanning is in the E-plane in [14], as opposed to H-plane scanning in this paper.

The concept presented in [1] uses an RWG and is quite similar to our proposed antenna. However, a single metallic rod is rotated. This rod introduced asymmetry in the slot excitation, leading to strong grating lobes 9 dB below the main lobe level. This problem is avoided in our proposed design by using two symmetrically rotated slabs.

The antenna proposed in this paper is not limited in power handling by the presence of nonlinear active components. Mechanical scanning speed is, of course, limited but remains acceptable for slowly moving targets. The mechanical stability of the slabs and the effect of slabs bending on the radiation pattern deserve further studies. This has to be considered in future work. However, stands [see Fig. 15(b)] made of low permittivity materials in the slotted region of the waveguide minimize slab deformation.

A good agreement between simulations and measurements also confirms the reliability and accuracy of the slot characterization approach. Since no optimization of the complete model in a full-wave simulator has been done, there is still a possibility for improvement in the antenna performance.

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[15] Amirhossein Ghasemi received the M.Eng. degree in embedded electronics and communication systems from the University of Paris-X, Paris, France, in 2012 and the M.Sc. and the Ph.D. degrees in electrical engineering from the Polytechnique de Montreal, Montreal, QC, Canada, in 2015 and 2018, respectively.

[16] From 2012 to 2014, he was a Research Engineer with the Institute d’Électronique Fondamentale, Paris, where he was involved in the design and implementation of passive and active beam scanning in metamaterial-based leaky-wave antennas. He is currently a Post-Doctoral Fellow with the Poly-Grames Research Center, Polytechnique de Montreal. His current research interests include slotted waveguide array antennas, high-power mechanical beam scanning antennas, and beamforming arrays for millimeter-wave applications.

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[23] From 2012 to 2014, he was a Research Engineer with the Institute d’Électronique Fondamentale, Paris, where he was involved in the design and implementation of passive and active beam scanning in metamaterial-based leaky-wave antennas. He is currently a Post-Doctoral Fellow with the Poly-Grames Research Center, Polytechnique de Montreal. His current research interests include slotted waveguide array antennas, high-power mechanical beam scanning antennas, and beamforming arrays for millimeter-wave applications.

[24] Jean-Jacques Laurin (S’87–M’91–SM’98) received the B.Eng. degree in engineering physics from the Ecole Polytechnique de Montreal, Montreal, QC, Canada, in 1983 and the M.A.Sc. and Ph.D. degrees in electrical engineering from the University of Toronto, Toronto, ON, Canada, in 1986 and 1991, respectively.

[25] In 1991, he joined the Poly-Grames Research Center, Ecole Polytechnique de Montreal, where he is currently a Professor. His current research interests include reconfigurable antennas, antenna design and modeling, wave processing surfaces, near-field antenna measurement techniques, and electromagnetic compatibility.