A Frequency-Scanned Slow-Wave Waveguide Antenna at Millimeter-Wave Frequencies

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ABSTRACT
In this paper we describe the design, theory, implementation, and measurement results of a novel frequency-scanned slow-wave slotted waveguide antenna. The slow-wave antenna is fabricated by periodically loading a standard WR22 waveguide using cylindrical posts. The antenna has a backward radiation and can steer the beam from $-72^\circ$ to $-12^\circ$ by changing the frequency from 32.5 GHz to 37.4 GHz. The gain of the antenna remains within a 3 dB range from the maximum gain of 17 dBi throughout the steering range. The antenna is 29$\lambda$ in length and 0.69$\lambda$ in width. The antenna radiation efficiency is between 74% and 92% throughout its frequency range which allows for further extension of its length in order to achieve a higher gain and a smaller beamwidth. Moreover, the small lateral width of the antenna allows placing several of them side-by-side to either narrow or facilitate scanning the beam in the transverse plane. The antenna has a fast and constant scanning rate of 4.3$^\circ$ over a 1% bandwidth and a return loss of less than $-10$ dB throughout its operating frequency range.

INDEX TERMS
Antenna, beam steering, frequency scanning, slot array, slow-wave, radars.

I. INTRODUCTION
The need for electrical beam-steerable antennas at millimeter and sub-millimeter wave frequencies is growing due to their expanding applications in different fields such as high data-rate communication, high-resolution radars, imaging systems, etc. [1]–[4]. Electrical beam-steering can be accomplished either by using tunable electronic components, such as diodes and variable capacitors, at a fixed frequency or by frequency-scanning [5]–[7]. At millimeter and sub-millimeter frequencies, tunable electronic components become more expensive and they introduce more power dissipation to the system compared to lower frequencies [8], [9]. Therefore, in some applications such as frequency-scanned radars, frequency-scanning is preferred in order to reduce the cost and increase the efficiency in this frequency range [10], [11]. In addition, the phase-shifting mechanism is much simpler in frequency-scanned antennas, as it is inherent in the nature of traveling waves without the need to design a complicated feeding network. Therefore, using frequency-scanned antennas can reduce the complexity of the device.

One important characteristic of antennas is their gain. A higher antenna gain provides a larger signal to noise ratio for communication links. A higher gain is also associated with a smaller beamwidth which results in a higher angular resolution in radar and imaging systems, as well as a longer range. In order to increase the gain of an antenna, one should decrease the beamwidth in both the E and H planes by extending the effective radiating lengths of the antenna in both directions [12]. As frequency-scanned antennas are usually series-fed, extending their length along the direction of the wave propagation increases the loss exponentially. Therefore, it is essential to reduce the loss per unit length of a series-fed frequency-scanned antenna in order to make it possible to extend its length without degrading the efficiency substantially. In particular, substrate-based frequency-scanned antennas are not favorable for applications that require high gain and narrow beamwidth because substrate loss grows exponentially with increasing length [13].

In order to extend the effective radiating length in the direction perpendicular to the direction of wave propagation in the waveguide, one can arrange several frequency-scanned antennas side-by-side. However, if the lateral width of the antenna is much larger than half-wavelength, this method cannot be used as it produces grating lobes. Therefore, serpentine waveguide-based antennas, which is one class of the most frequently used antenna types for frequency scanning, are not suitable for increasing the lateral effective radiating
length [14]. To narrow the beamwidth in the non-scanning plane, Sarabandi et al. developed the design of a serpentine antenna that narrows the beam in the elevation plane by putting a patch array along each meandered branch [15]. This method narrows the beam in the plane perpendicular to the steering plane. However, it is limited by the length of the meandered branches and it is also not suitable for 2D scanning. In [16], Chen et al. introduced a metasurface that narrows the beam of a frequency-scanned slotted-waveguide antenna in the non-scanning plane and hence increasing the gain. This approach leads to a considerable increase in the gain of a serpentine antenna but the presence and cost of an additional PCB structure might not be desirable. Furthermore, it is still not possible to achieve 2D scanning using this method. In [17], Chu et al. have utilized a meandered stripline that is coupled to radiating elements that are placed on one side of it. The radiation is oriented towards the thin side of the substrate which reduces the lateral width of the antenna to the thickness of the substrate. This method makes it possible to place several of them side-by-side to enable 2D beam-steering. The drawback of it however is the resulting low efficiency due to the substrate loss.

Another way to achieve a narrow beamwidth in both the E and H planes is to make a low-loss, slow-wave structure in a periodic waveguide with a small lateral profile. Using this method, one can increase the dynamics of the propagation constant and achieve a large steering angle within a desired bandwidth by tailoring the dispersion relation. One way to implement such periodic structures is through substrate integrated waveguides (SIW). Due to the flexibility in the fabrication of SIWs, perturbations such as vias can be easily used to close the stop-band of the periodic structure at broadside [18], [19]. However, the dielectric loss in frequency-scanned SIW antennas limits their efficiency, gain, and beamwidth in the scanning plane. This effect becomes more significant at millimeter and sub-millimeter frequencies. Furthermore, substrate-based frequency-scanned antennas are prone to deep dielectric charging in space applications which can lead to electrical breakdown and damage to the substrate. Moreover, high quality microwave substrate cost can add considerably to the overall fabrication cost.

In order to overcome the limitations imposed by the dielectric loss of SIW antennas at high frequencies, one can use loaded air-filled waveguides such as corrugated waveguides. In [20], non-standard size rectangular waveguides with corrugated bottom walls are placed side-by-side. The structure is fabricated using a micromachining technique. Although a narrow beamwidth in both scanning and non-scanning planes is achieved, the corresponding gain fluctuations are large such that imposing the constraint of a maximum 3 dB gain variation limits the steering angle to less than 30°.

In this paper, we propose a novel loaded WR22 waveguide which has a small lateral profile, a high efficiency, and small gain fluctuations throughout its operating frequency range. This is achieved due to the combination of its loading scheme and the slot orientation. An important attribute of the proposed antenna is that its length can be extended to further decrease the beamwidth without significant additional power dissipation thanks to its high radiation efficiency. Moreover, due to its small lateral size, one can place several of them side-by-side to narrow the beam in the H-plane or to implement 2D beam-steering. A prototype of the antenna was fabricated and measured within the frequency range from 32.5 GHz to 37.4 GHz and a good agreement between the measurement and the simulation results was found.

II. DESIGN OF THE SLOW-WAVE WAVEGUIDE

In order to simplify the excitation and termination of the antenna, we chose WR22 standard waveguide as the basis of the periodic structure, as it supports the desired frequency range. Investigating different schemes of loading the waveguide periodically, we decided to use two symmetric cylindrical posts on the top and bottom walls of the waveguide facing each other as depicted in Fig. 1. This scheme of loading has some advantages over other schemes such as corrugated waveguides as described below.

This proposed method of loading confines the current distribution to the middle of the top and bottom walls of the waveguide (see Fig. 2). This current confinement minimizes the loss due to the gap in the middle of the two side-walls that is made by joining the top and bottom parts of the waveguide after assembly. Another advantage of the current confinement is that it maximizes the amount of current captured by transverse slots placed in the middle of the top wall such that they can radiate the desired amount of power with a shorter length. Therefore, we can further reduce the lateral width of the waveguide without decreasing the radiated power or changing the modal configuration inside the waveguide. We optimized the radius of the cylindrical posts and their spacing to achieve a maximum current confinement. As shown in Fig. 2, the current surrounding the posts is more than ten times larger in magnitude compared to the current on the two side-walls for a post radius of 0.3mm and a spacing of 2mm.

The scanning range and its corresponding bandwidth are determined by the dispersion relation in the waveguide. In order to have a large steering angle within a certain bandwidth, we need to increase the rate of change of
the phase-shift between consecutive slots with frequency. In Fig. 3, the dispersion diagram of the proposed periodically loaded waveguide for different heights of the post is plotted. Increasing the height of the post increases the dynamics of the dispersion relation with frequency which in return increases the rate of change of the phase-shift between consecutive slots. As shown in Fig. 3, periodically loading the waveguide decreases the cut-off frequency. However, as we use regular WR22 waveguide excitation and termination for simplification, we choose the frequency band of 32.5 GHz to 37.5 GHz to operate in. According to the dispersion relation, the optimum value for the height of the posts that provides the desired dynamics within our operating frequency range is 1.2mm. We also confirmed that the designed slow-wave waveguide is single-moded throughout the operating frequency range using the HFSS eigenmode solver.

The last step of the slow-wave waveguide design is to match it to a regular WR22 waveguide in order to use a 2.92mm coax-to-WR22 waveguide adapter at the two ends for the excitation and termination. We tapered the height of the posts with distance to achieve this goal. Specifically, the height of the posts starts from 0 and is incremented by 0.1mm until it reaches the full height of 1.2mm. Fig. 4.a shows the tapered section structure. Fig. 4.b shows the measured and simulated return loss of a small fabricated prototype of the periodic waveguide without slots when excited by a regular WR22 through the tapered section.

The waveguide was fabricated using CNC machining with a precision of $+/- 0.02\text{mm}$. The upper and lower parts where fabricated separately. We chose the middle of the sidewalls to cut the waveguide for fabrication since unlike the unloaded case, the current density is minimized in that area according to Fig. 2. We used aluminum since it has low loss and it is also suitable for CNC machining.

### III. ANALYSIS OF THE WAVE PROPAGATION IN THE SLOW-WAVE WAVEGUIDE

In order to analytically derive the dispersion relation and the fields modal configuration in the designed slow-wave waveguide, we can simplify the structure by replacing the posts with equivalent PEC strips [21]. Two strips with the same widths as the diameter of the posts and the length equal to 1.35mm provides the best match for the dispersion relation. Since the wave is assumed to be concentrated in the middle of the waveguide cross-section as it propagates, we can further simplify the structure by extending the widths of the strips to reach the side walls of the waveguide to form diaphragms (as shown in Fig. 5), without considerably changing the dispersion relation and the fields modal configuration. Fig. 6 and Fig. 7 compare the simulated dispersion relation and the fields modal configuration of the simplified structures with those of the original designed waveguide with cylindrical posts respectively using the HFSS eigenmode solver. As shown,
These hybrid modes are longitudinal-section electric (LSE) stored energy in the vicinity of the capacitive diaphragms. The electric field in the vicinity of each diaphragm can be written as a superposition of a TE mode and one or more higher order hybrid (LSE) modes. Therefore, the modal configuration of the symmetric diaphragms is given by the following expression:

$$\nabla \times H = k_0^2 \hat{n} + \nabla \times \Pi$$

A suitable form for the nth magnetic vector potential mode that satisfies the boundary conditions of the waveguide is:

$$\Pi = \sin \left( \frac{\pi x}{a} \right) \cos \left( \frac{n \pi y}{b} \right) e^{j \beta n z}$$

Using (3) in (1) and (2), results in the following electric and magnetic fields for the nth mode:

$$E_y = j \omega \mu_0 \beta_n \sin \left( \frac{\pi x}{a} \right) \cos \left( \frac{n \pi y}{b} \right) e^{j \beta n z}$$

$$E_z = j \omega \mu_0 \frac{n \pi}{b} \sin \left( \frac{\pi x}{a} \right) \cos \left( \frac{n \pi y}{b} \right) e^{j \beta n z}$$

$$H_x = \frac{k_0^2 - \left( \frac{n \pi}{b} \right)^2 - \left( \frac{\pi}{a} \right)^2}{\alpha b} \sin \left( \frac{n \pi y}{b} \right) e^{j \beta n z}$$

$$H_y = -\frac{n \pi}{ab} \cos \left( \frac{\pi x}{a} \right) \cos \left( \frac{n \pi y}{b} \right) e^{j \beta n z}$$

$$H_z = \frac{\pi}{a} \beta_n \cos \left( \frac{\pi x}{a} \right) \cos \left( \frac{n \pi y}{b} \right) e^{j \beta n z}$$

The analytic solution of the amplitudes of each excited mode when a TE_{10} wave is incident on the diaphragm can be found using the variational technique as explained in [22].

As we are interested in the periodic structure in which the symmetric diaphragms are placed periodically along a rectangular waveguide, we have to consider the interactions between different modes. If the adjacent diaphragms are placed relatively close such that the excited evanescent modes are not attenuated substantially when they reach the adjacent diaphragms, the incident field on each of them is a combination of the dominant TE_{10} mode and one or more higher order hybrid (LSE) modes. Therefore, the modal configuration of the resulting eigenmode solution of the periodic structure will be a linear summation of the modal configuration of the TE_{10} mode and that of one or more higher order hybrid modes. As a result, the propagating wave in the slow-wave waveguide will be a hybrid mode consisting of $E_y$ and $E_z$.

In order to find the eigenmode solution of the simplified slow-wave waveguide depicted in Fig. 5.c, we follow the procedure discussed in [22]. We exploit the symmetry of the structure by placing a longitudinal PEC plate in the middle of the waveguide and parallel to the top and bottom walls. In this way, the waveguide is divided into two sections (the lower half and the upper half) both of which are slow-wave waveguides with one diaphragm periodically placed along the waveguide. The lower half section of the waveguide is shown in Fig. 8. We can solve for the fields in this section and use the symmetry to find the fields in the upper half section. Therefore, b in the relations for the lower half should be substituted by b/2 (half the height of the original waveguide). We assume that the incident wave on a diaphragm in the
The periodic structure can be written as (see Fig. 8):

\[
E_y = \begin{cases} 
  a_0 \sin(\frac{\pi x}{a}) e^{-\betaxz} + a_1 \sin(\frac{\pi x}{a}) \cos(\frac{\pi y}{b}) e^{-\betayz} + \\
  a_2 \sin(\frac{\pi x}{a}) \cos(\frac{2\pi y}{b}) e^{-\betayz} + a_3 \sin(\frac{\pi x}{a}) \cos(\frac{3\pi y}{b}) e^{-\betayz} + \\
  \sum_{n=0}^{\infty} a_n \sin(\frac{\pi x}{a}) \cos(\frac{n\pi y}{b}) e^{-\betanz} \\
  \sum_{n=0}^{\infty} c_n \sin(\frac{\pi x}{a}) \cos(\frac{n\pi y}{b}) e^{-\betanz}
\end{cases}
\]

\[
H_x = \begin{cases} 
  k_0^2 a_0 \sin(\frac{\pi x}{a}) e^{-\betaxz} + a_1 \sin(\frac{\pi x}{a}) \cos(\frac{\pi y}{b}) e^{-\betayz} + \\
  a_2 \sin(\frac{\pi x}{a}) \cos(\frac{2\pi y}{b}) e^{-\betayz} + a_3 \sin(\frac{\pi x}{a}) \cos(\frac{3\pi y}{b}) e^{-\betayz} + \\
  \sum_{n=0}^{\infty} b_n \sin(\frac{\pi x}{a}) \cos(\frac{n\pi y}{b}) e^{-\betanz} \\
  \sum_{n=0}^{\infty} c_n \sin(\frac{\pi x}{a}) \cos(\frac{n\pi y}{b}) e^{-\betanz}
\end{cases}
\]

Assuming that the electric field in the aperture above the diaphragm is \( E_a \), and given the fact that the tangential electric field, \( E_y \), is continuous throughout the transverse cross-section of the waveguide at the location of the diaphragm (z = 0), the following equations can be derived by Fourier analysis:

\[
a_0 + b_0 = c_0 = \int_0^b E(y') dy'
\]
\[
a_1 + b_1 = c_1 = \int_0^b E(y') \cos(\frac{\pi y}{b}) dy'
\]
\[
a_2 + b_2 = c_2 = \int_0^b E(y') \cos(\frac{2\pi y}{b}) dy'
\]
\[
a_3 + b_3 = c_3 = \int_0^b E(y') \cos(\frac{3\pi y}{b}) dy'
\]
\[
a_n = c_n = \int_0^b E(y') \cos(\frac{n\pi y}{b}) dy' \quad (12)
\]

The tangential magnetic field is continuous only on the aperture and it is discontinuous on the diaphragm. Therefore, we need to define a set of orthogonal functions on the aperture to apply the continuity of the magnetic field using Fourier analysis. As suggested in [22], we can use the following change of variables for this purpose:

\[
\cos(\frac{\pi y}{b}) = \alpha_1 + \alpha_2 \cos(\theta) \quad (13)
\]
\[
\alpha_2 = \alpha_1 + 1 = \cos^2(\frac{\pi d}{2b}) \quad (14)
\]

We can replace \( E(y')dy' \) in the integrand by \( F(\theta')d\theta' \) and approximate \( F(\theta') \) by the sum of the first 4 terms of the defined orthogonal functions in the aperture:

\[
F(\theta) = A_0 + A_1 \cos(\theta) + A_2 \cos(2\theta) + A_3 \cos(3\theta) \quad (15)
\]

Using equations (12) and (15) and applying the continuity of \( H_x \), the transmission matrix (T) which describes the modal interaction between the 4 modes can be found, (16)–(18), as shown at the bottom of the next page.

Knowing the T matrix, the eigenmode and the characteristic propagation constant can be found by solving the following matrix eigenvalue system [22]:

\[
\begin{bmatrix}
  U \cosh(2T^{-1} - U) & 0 \\
  2(T^{-1} - U) & U
\end{bmatrix} = \begin{bmatrix}
  V \\
  I
\end{bmatrix} = e^{\gamma l} \begin{bmatrix}
  V \\
  I
\end{bmatrix} \quad (19)
\]

where U is a 4 × 4 identity matrix. The \( \Gamma, V, \) and \( I \) matrices are defined as follows (see Fig. 8).

\[
\Gamma = \begin{bmatrix}
  \beta_0 & 0 & 0 & 0 \\
  0 & \beta_1 & 0 & 0 \\
  0 & 0 & \beta_2 & 0 \\
  0 & 0 & 0 & \beta_3
\end{bmatrix}
\]

\[
V = \begin{bmatrix}
  a_0' + b_0' \\
  a_1' + b_1' \\
  a_2' + b_2' \\
  a_3' + b_3'
\end{bmatrix}, \quad I = \begin{bmatrix}
  a_0' - b_0' \\
  a_1' - b_1' \\
  a_2' - b_2' \\
  a_3' - b_3'
\end{bmatrix} \quad (20)
\]
The eigenmode solution is a superposition of the TE\textsubscript{10} mode and the first 3 dominant hybrid modes. We solved the eigenvalue system for the simplified form of the designed slow-wave waveguide with periodic diaphragms as depicted in Fig. 5c for different frequencies. The wave vector corresponding to each frequency calculated by finding the eigenvalue of equation (19) is plotted in Fig. 9. As shown, a very good agreement between the calculated result and the simulation is achieved. It is also shown that the calculated dispersion relation is a very good approximation of that of the slow-wave waveguide with cylindrical posts.

Using the calculated eigenvector of equation (19), the modal electric field configuration is plotted in Fig. 10 in the plane of the diaphragm as well as in the terminal plane (the transverse plane in the middle between two consecutive diaphragms, see Fig. 8) at 32 GHz. As shown, the electric field is concentrated in the middle of the waveguide as expected from the simulation results in Fig. 7. The wave is also more concentrated in the diaphragm plane compared to the terminal plane. As shown in Fig. 10, the y component of the electric field in the diaphragm plane is maximum in the middle of the transverse plane where the aperture is located. On the other hand, the z component of the electric field in this plane is maximum at the edges of the two diaphragms close to the aperture. It is also worth mentioning that the calculated tangential electric field (E\textsubscript{y}) is not exactly zero on the diaphragms in the diaphragm plane. This is because we have included limited number of hybrid modes. A more exact result for the electric field configuration and the dispersion relation can be achieved by including more higher order hybrid modes.

The modal electric field configuration could not be explained without including the hybrid modes which were evanescent in the case of only one diaphragm. However, due to the close spacing of the diaphragms in the designed periodic slow-wave waveguide, the hybrid modes contribute significantly to the propagating modal field configuration and they account for the field concentration in the middle of the waveguide which cannot be explained by considering only the TE\textsubscript{10} mode. Fig. 11 shows the convergence diagram of the modal electric field configuration at 32 GHz on the diaphragm plane as we consider 1, 2, 3, and 4 modes. As shown, the confinement of E\textsubscript{y} in the middle of the aperture between the two diaphragms becomes more clear as we consider more hybrid modes. This is why we considered 4 modes.

\[
\begin{align*}
T &= \begin{bmatrix}
\frac{1}{u} & \frac{2\alpha_1 \beta_0}{\pi u} + \alpha_2^2 & \frac{2\alpha_1 \beta_0}{\pi u} + \alpha_2^2 & \frac{2\alpha_1 \beta_0}{\pi u} + \alpha_2^2 \\
2\alpha_1 u & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 \\
2\alpha_1 u & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 \\
2\alpha_1 u & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 & \frac{\pi u}{2\beta_0^2} + \alpha_2^2 \\
\end{bmatrix}
\end{align*}
\]
FIGURE 11. The modal field configuration convergence diagram as we consider more modes on the diaphragm plane at 32 GHz.

FIGURE 12. The correlation between the modal field configuration in the slow-wave waveguide with diaphragms and that in the designed slow-wave waveguide with cylindrical posts at 32 GHz.

to be more accurate. On the other hand, it is worth mentioning that the calculated dispersion relation shown in Fig. 9 is converging quite rapidly and even if only one mode is retained the results are satisfactory.

Finally, in order to verify the validity of the simplification used, we need to investigate the correlation between the modal field configuration in the simplified slow-wave waveguide with diaphragms and that in the designed slow-wave waveguide with cylindrical posts. Fig. 12 shows the $y$ and $z$ components of the analytically calculated electric field on the terminal plane in the simplified slow-wave waveguide with diaphragms and the simulated $y$ and $z$ components of the electric field in the terminal plane of the designed slow-wave waveguide with cylindrical posts at 32 GHz. As shown, a strong correlation exists between the corresponding electric field components in the two cases. Naturally, the electric field in the simplified waveguide with diaphragms is slightly stretched along the $x$ axis compared to that in the waveguide with cylindrical posts. Therefore, the analytically calculated eigenmode solution for the simplified structure with diaphragms can be used to explain both the dispersion relation and the modal field configuration with hybrid mode in the designed waveguide with cylindrical posts.

IV. ARRAY DESIGN

The slot spacing needs to be close to half-wavelength of the maximum frequency in order to maximize the gain and prevent any grating lobes. As the periodicity of the slow-wave waveguide is 2 mm, we chose a slot spacing of 4 mm which is half-wavelength at the maximum frequency of operation (37.5 GHz). As the periodicity of the waveguide antenna is changed from 2 mm to 4 mm by introducing the slots, one can determine the dispersion diagram of the waveguide (without the slots) with periodicity of 4 mm by transferring the part of the dispersion diagram in Fig. 3 which is in the 2nd Brillouin zone (BZ), to the 1st Brillouin zone, as shown in Fig. 13.a. Figure 13.b shows the dispersion diagram of the waveguide including the effect of the slots placed periodically 4 mm from each other. As shown, adding the slots introduces a band-gap in the dispersion diagram. The part of the diagram that is used for radiation is highlighted.

Knowing the dispersion relation of the waveguide and the slot spacing, the steering range can be calculated from (22) where $\lambda_0$ and $\lambda_g$ are the wavelengths in vacuum and in the slow-wave waveguide respectively, and $D$ is the slot spacing [12].

The steering range can be found to be from $-75^\circ$ to $0^\circ$ when we change the frequency from 32 GHz to 37.5 GHz which is opposite to the direction of the wave propagation in the waveguide.

$$\theta = 90 - acos\left(\frac{\lambda_0^\circ}{\lambda_g} - \frac{\lambda_0^\circ}{D}\right)$$ (22)

As the steering range is extended from broadside towards angles close to the grazing angle, we developed an array
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This paper discusses the design of a frequency-scanned slow-wave waveguide antenna at millimeter-wave frequencies that minimizes gain variations throughout the steering range. The antenna array directivity can be found from (23) which is the result of multiplying the array factor (AF) by the element factor (EF). For a linear array with a uniform excitation, the array factor can be found from (24) and (25), where $N$ is the number of radiating elements, $k_o$ is the wavenumber in vacuum, $d$ is the slot spacing, $\beta$ is the wavenumber in the waveguide, and $\phi$ is the angle at which the directivity is calculated [12]. As the array factor is fixed for a uniformly excited linear antenna array with a certain number of elements, we developed a favorable element factor in order to minimize the gain fluctuations for different steering angles.

$$D(\phi) = AF(\phi) \times EF(\phi)$$  \hspace{1cm} (23)

$$AF(\phi) = \frac{\sin(N\psi/2)}{\sin(\psi/2)}$$  \hspace{1cm} (24)

$$\psi = k_o.d.\cos(\phi) - \beta.d$$  \hspace{1cm} (25)

Fig. 14 compares the radiation patterns in the E-plane and H-plane of a slot. In the E-plane, the gain of the slot shows little fluctuations throughout our desired steering range. However, in the H-plane, the gain of the slot drops rapidly as we increase the angle from broadside such that at our maximum designed radiation angle ($-75^\circ$), the element factor of the H-plane is 8 dB lower than that of the E-Plane. Therefore, we used transverse slots to exploit their favorable E-plane radiation pattern in the scanning plane. Furthermore, transverse slots also improve the radiation pattern in the non-scanning plane. By using transverse slots, the non-scanning plane of the antenna array is aligned with the H-plane of the slots which has a narrower beamwidth compared to the E-plane. As a result, the antenna array has a narrower beamwidth in the non-scanning plane.

Using Fig. 15, we derived the lengths of the 45 slots for a uniform excitation. The lengths of the slots are shown in Fig. 16. The gradual increase in the lengths of the slots can be seen from the first slot to the 31st slot. From slot number 32, the length of a single slot will be so large that the high reflection from it will disturb the phase distribution of the previous slots. Therefore, we improved the matching to decrease the reflection in this part of the antenna by designing 2 slots together. With this method, there is more flexibility in achieving higher radiated power with less reflection. The two slots are designed simultaneously such that their reflections add destructively in order to minimize their overall reflection while maximizing their combined radiated power. The best result was achieved by alternating the size of the slots between 3.3mm and 3.6mm. On the other hand, the widths of all the slots are the same and equal to 0.3mm.

**V. SIMULATION AND MEASUREMENT RESULTS**

Fig. 17 shows a picture of the fabricated antenna. It can be seen that both ends are connected to a 2.92mm coax-to-WR22 adapter. The adapter at the left side of the picture...
is used for excitation and the one at the right side is used for attaching a matched load at the end of the antenna to absorb all the remaining power and prevent any reflections. The physical width of the fabricated antenna is 38mm while the width of the slow-wave waveguide is only 5.69mm. The extra width is just for providing mechanical support.

Fig. 18 shows the simulated and measured gain of the fabricated antenna in the E-plane. 0° is broadside, negative and positive angles are for backward and forward radiation respectively.

The physical width of the fabricated antenna is 38mm while the width of the slow-wave waveguide is only 5.69mm. The extra width is just for providing mechanical support.

The measured maximum radiation angle changes from −72° to −12° by changing the frequency from 32.5 GHz to 37.4 GHz. The gain of the antenna remains within the 3 dB limit of the maximum gain of 17 dBi at 33.2 GHz. The half-power beamwidth of the antenna is 4° and it can be decreased further by extending the length of the waveguide.

Fig. 19 also demonstrates the linear change of the scan angle versus frequency. As shown, the scanning happens at a constant rate of 12°/GHz which is equivalent to a scanning rate of 4.3° per 1%-bandwidth throughout the operating frequency range. This figure also compares the measured results with the simulated results from the eigenmode analysis of a unit cell in HFSS which shows a very good agreement between the two.

The measured gain, directivity, and radiation efficiency of the slow-wave waveguide antenna are plotted in Fig. 20. As shown, the radiation efficiency varies between 74% and 92% throughout the operating frequency range with its peak at 34 GHz. The lower efficiency at the lower frequencies can be attributed to the remaining power at the end of the waveguide.
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The measured and simulated $S_{11}$ and $S_{21}$ of the antenna are shown in Fig. 21. Fig. 21.a shows that the return loss is below $-10$ dB throughout the frequency range. Fig. 21.b shows that less power is received and dissipated at the second port as the frequency is increased which is equivalent to more radiated power through the slots.

VI. COMPARISON AND DISCUSSION

Table 1 compares different characteristics of relevant recent works on frequency-scanned antenna arrays. The frequency-scanned antennas presented in [17]–[19], and [23] are fabricated using substrate-integrated waveguides. Due to the flexibility of planar fabrication, they can achieve a large steering range by closing the stop-band using perturbations in the waveguide. In this way, they can include both the forward and backward radiation in their scanning range. However, the substrate loss limits their efficiency such that [17] and [19] cannot achieve an efficiency higher than 60% with an antenna length of around $10\lambda$. In [18] and [23], the shorter antenna length and the lower frequency of operation are the reasons for their relatively higher efficiency which would be lower otherwise. The leaky wave antenna presented in [24] has half the width of the antenna designed in [23] but with similar specifications which can be used more easily for 2D scanning. However, due to the drop in efficiency, it also cannot be used for length extension. As the substrate loss increases exponentially with increasing the length, these works are not suitable for length extension in order to further increase the gain and decrease the scanning beamwidth. Specifically, the negative effect of substrate loss manifests itself more at higher frequencies, preventing us from achieving high efficiencies with SIW antennas.

The proposed antennas in [15], [16], and [20] are fabricated using air-filled waveguides. Therefore, they are not prone to substrate loss and are suitable for further length extension without degrading the radiation efficiency substantially. However, none of them are useful for 2D beam steering. The antenna width in [15] and [16] is much larger than half-wavelength which prevents placing them side-by-side to allow 2D beam-steering due to the formation of undesired grating lobes. In [20], the large gain variation of more than 10 dB throughout the steering range also makes this antenna not suitable for 2D beam-steering.

In this work, thanks to the proposed scheme of loading, the small lateral profile and the high radiation efficiency make the proposed slow-wave slotted waveguide antenna suitable for both 2D beam-steering and length extension simultaneously. Therefore, it can serve as an advantageous alternative for SIW and serpentine antennas when 2D beam-steering and narrow beamwidth are needed at the same time. Due to the fabrication constraints and the space limitation inside the waveguide, it was not possible to close the stop-band using perturbations in order to include the forward radiation in the scanning range. However, the antenna can provide a large steering range of 60° thanks to the combination of the slot

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**TABLE 1. Comparison with other frequency scanning antenna arrays.**

| Reference | Antenna type | Bandwidth | Scanning range | Measured gain variation | Efficiency | Antenna length | Antenna width | Suitable for 2D scanning | Suitable for length extension |
|-----------|--------------|-----------|----------------|------------------------|------------|----------------|---------------|-----------------------------|------------------------------|
| [15]      | Waveguide    | 6.3% (230-245 GHz) | 50° (-25° – +25°) | 29-30.5 dBi | 55%-65%* | 35.6λ | 6.7λ | No | Yes |
| [16]      | Waveguide with meta-surface | 5.2% (33.5-35.3 GHz) | 26° (-26° – 0°) | 21-25.3 dBi | Not reported | 40λ | 15λ | No | Yes |
| [17]      | SIW          | 14.3% (13-15 GHz) | 76° (-40° – +36°) | 14.3-15.6 dBi | 53%-57% | 9.4λ | 0.1λ | Yes | No |
| [18]      | SIW          | 36.6% (7.6-11 GHz) | 119° (-74° – +45°) | 9.1-13.5 dBi | 70%-90%** | 7.9λ | 0.42λ | Yes | No |
| [19]      | SIW          | 16.1% (13.2-15.6 GHz) | 103° (-61° – +42°) | 8-13 dBi | 30%-60% | 11.1λ | 0.37λ | Yes | No |
| [23]      | SIW          | 44.7% (8.25-13 GHz) | 126° (-60° – +60°) | 10-15 dBi | 75%*** | 5.4λ | 0.79λ | Yes | No |
| [20]      | Waveguide    | 18.2% (150-180 GHz) | 60° (-40° – +20°) | 19-31 dBi | Not reported | 27.5λ | 27.5λ | No | Yes |
| This work | Waveguide    | 14% (32.5-37.4 GHz) | 60° (-72° – -12°) | 14-17 dBi | 74%-92% | 29λ | 0.69λ | Yes | Yes |

* Simulated radiation efficiency.
** Extrapolating the antenna length to 29λ results in an efficiency of 27%-68%.
*** Extrapolating the antenna length to 29λ results in an efficiency of 22%.
orientation and the loading scheme which makes radiation close to the grazing angle possible without considerable drop in the gain.

VII. CONCLUSION
In this paper we discussed the design and measurement results of a novel traveling slow-wave frequency-scanned antenna. The antenna can be used for applications that need 2D beam-steering and narrow beamwidth in the E and H planes owing to its small lateral profile and high efficiency respectively. The antenna can steer the beam from $-72^\circ$ to $-12^\circ$ when the frequency is changed from 32.5 GHz to 37.4 GHz. Orienting the slots transverse to the waveguide direction allows radiating very close to grazing angles without any considerable decrease in the gain. It also makes the beamwidth narrower in the non-scanning plane compared to longitudinal slots. The gain remains within a 3 dB limit of the maximum gain of 17 dBi over the entire steering range. The antenna radiation efficiency is between 74% and 92% and it has a half-power beamwidth of $4^\circ$. The HPBW can be further reduced by extending the length of the antenna without any significant drop in the efficiency. Due to its special structure, it can provide flexibility in the dispersion required for beam-steering while requiring much smaller width compared to serpentine waveguides. Furthermore, the antenna is not susceptible to substrate loss and can provide higher radiation efficiency compared to SIW antennas. This antenna is well suited for applications such as frequency-scanned radars with high angular resolution.

As a future direction to this research work, we will consider closing the stop-band thus allowing the inclusion of broadside and forward radiation in the steering range of the antenna. This can be achieved by tuning the post heights individually. This however is beyond the scope of the present work and could be included in a forthcoming report.

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