Slotted Waveguide Frequency-Scanned Slow-Wave Antenna With Reduced Sensitivity of the Closed Stopband at Millimeter-Wave Frequencies

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ABSTRACT In this paper, we describe the design and implementation of a frequency-scanned slotted waveguide antenna with a closed stopband which has a low sensitivity to the fabrication errors. The antenna is implemented as a slotted slow-wave waveguide by loading a standard WR22 waveguide with elliptical posts. A systematic design approach for scanning through broadside and closing the stopband with a reduced sensitivity to fabrication tolerances is discussed in the paper. This approach allows for fabricating the antenna using CNC machining with a tolerance of $\pm 0.13$ mm for the waveguide and a tolerance of $\pm 0.05$ mm for the posts. The antenna beam can be steered from $-38^\circ$ to $+27^\circ$ by changing the frequency from 27 GHz to 34.7 GHz. The effective radiating length of the antenna is 27.6$\lambda$, corresponding to a 1.75$^\circ$ half-power beamwidth while its radiation efficiency changes between 54% and 90% throughout the steering range.

INDEX TERMS Antennas, beam steering, slotted waveguide, frequency control, slow-wave, millimeter-wave, leaky-wave.

I. INTRODUCTION The need for low-cost, efficient, and extendable beam-steerable antennas at millimeter frequencies has been rising due to their expanding applications in communication systems, medical imaging, autonomous vehicles radars, and remote sensing [1]–[7]. Electrical beam-steerable antennas can be classified into fixed-frequency beam-steerable antennas and frequency-scanned antennas. Due to the considerable drop in the efficiency of tunable electronic components at millimeter-wave frequencies, frequency-scanned antennas can provide better efficiencies at this frequency range, especially if a long electrical length is needed for a small half-power beamwidth (HPBW) and a large gain [8]–[10]. Moreover, at millimeter-wave frequencies, the use of metallic waveguides is appealing due to their reduced losses and the high cost of low-loss dielectric substrates. Hence in this work we adopt a waveguide approach.

Frequency-scanned antennas are implemented by increasing the phase dynamics of the guided wave with changing the frequency. This can be achieved by meandering a waveguide which results in an increased phase change between the consecutive slots with changing the frequency. Serpentine waveguides can provide any desired scanning rate by varying the length of the meandered branches [11]. However, due to their large width, they cannot be placed side-by-side to narrow the beam in the non-scanning plane. On the other hand, frequency-scanned antennas can be implemented by periodically loading a waveguide to support a slow-wave. Slow-wave waveguides can be placed side-by-side for beam-shaping in the non-scanning plane as their width can be close to half-wavelength [12]–[14]. Nevertheless, the scanning rate of slow-wave frequency-scanned antennas is limited to how much the guided wavelength can be decreased.

The performance of frequency-scanned antennas at broadside radiation is degraded due to the constructive addition of reflections from each radiating element. One approach to close the stopband at broadside radiation for waveguides is to use an additional post in the unit cell to cancel this reflection [15]. However, in slow-wave waveguide structures, as the guided wave is confined in a smaller volume, which results in a smaller wavelength, closing the stopband becomes more
challenging. This is because there may not be enough space to add a new post. Therefore, one has to perturb the features in the unit cell structure to achieve this goal. Furthermore, due to the decreased feature sizes in a unit cell, the closed stopband becomes more sensitive to any variations in the fabrication due to the limited tolerance. As a result, the scanning rate has a trade-off with the sensitivity of the closed stopband.

Different fabrication processes have different tolerances. For example, implementing a slow-wave frequency-scanned antenna using PCB fabrication provides a typical tolerance of $\pm 0.01$ mm. This is approximately equivalent to a good percentage tolerance of about 1% at millimeter frequencies assuming the smallest feature size of about 1 mm. Nevertheless, PCB substrate-based antennas operating at millimeter-wave frequencies suffer from a low efficiency due to the substrate loss. This drawback manifests itself more for electrically long antennas. On the other hand, silicon micro-machining can allow for a much smaller tolerance of $\pm 2$ $\mu$m [16]. However, this fabrication technique is limited by the size of the antenna as well as the cost of fabrication which makes it more useful for sub-millimeter frequencies. CNC machining is another versatile approach that can be used for fabricating electrically large antennas at a reasonable cost at millimeter frequencies. However, the drawback of CNC machining is its relatively poor tolerance. A typical tolerance in CNC machining is $\pm 0.13$ mm which could be reduced to $\pm 0.05$ mm at an additional cost. Therefore, it is very important to reduce the sensitivity of the closed stopband to the fabrication error when the antenna is implemented using CNC machining. This becomes even more important when a high scanning rate is needed.

In this paper, we propose a frequency-scanned slow-wave slotted waveguide antenna at millimeter frequencies using CNC machining. In [10], a related antenna has been proposed. However, it could not scan through broadside resulting in a limited steering range and functionality. The new antenna has a closed stopband with a low sensitivity to the fabrication error. A systematic approach for designing the antenna with a closed stopband and reduced sensitivity is proposed in sections II-V. A prototype of the antenna was fabricated and measured. The simulation and measurement results are compared and a comparison is made with other relevant research works in sections VI-VII.

II. DESIGN OF THE SLOW-WAVE WAVEGUIDE

The proposed slow-wave waveguide antenna is based on periodically loading the bottom wall of a regular WR22 waveguide with cylindrical posts. Fig. 1 shows a picture of the fundamental unit cell of the proposed loaded waveguide. It should be noted that the height of the loaded waveguide is 0.5 mm smaller than that of a standard WR22 due to a fabrication constraint. An analytic approach to derive an approximate dispersion relation for the loaded waveguide is shown in [10]. The derived slow-wave propagating mode in [10] is a superposition of the TE$_{10}$ mode of the regular waveguide as well as evanescent longitudinal-section electric (LSE) modes. The latter are hybrid modes consisting of both longitudinal electric and magnetic fields resulting in a hybrid propagating mode in the slow-wave waveguide. This analysis is done through approximating the posts using strips and diaphragms which simplifies the structure and allows for the derivation of the dispersion relation analytically.

Fig. 2 shows the dispersion relation of the proposed loaded waveguide. Modes 1 and 2 are the perturbed forms of the TE$_{10}$ mode in a regular WR22 waveguide while mode 3 is the perturbed form of its TE$_{20}$ mode. The latter has the minimum electric field magnitude in the middle of the waveguide at the location of the posts. Therefore, this mode is perturbed the least (mode 3 in the slow-wave waveguide). Nevertheless, mode 2 is very similar to mode 3 since the posts act as a wall dividing the waveguide into two identical waveguides with half the original width. As shown, the waveguide is single-moded in the shaded region when operating in its first mode below 40 GHz. This mode will be used for the proposed waveguide which is the hybrid mode that has been calculated analytically in [10].

One important characteristic of the proposed waveguide is the confinement of the fields and the current to the middle of the waveguide. Fig. 3 shows the current confinement in
FIGURE 3. Complex magnitude of the current density in the slow-wave waveguide.

one unit cell of the periodic waveguide. This current confinement is contrary to other classes of frequency-scanned antennas such as serpentine waveguide antennas and corrugated waveguide antennas. In split-block rectangular waveguides fabricated using CNC machining, the E-plane is usually used for splitting the structure in order to reduce the Ohmic loss in the gap after assembly [17]. However, due to the presence of the posts, the proposed slow-wave waveguide should be cut in the H-plane. This current confinement reduces the Ohmic loss due to the gap in the middle of the side walls after assembly resulting in an increased radiation efficiency which proves important for series-fed leaky-wave antennas (LWA). Furthermore, due to the field confinement in the middle of the waveguide, one can further reduce the lateral width of the proposed waveguide without changing its dispersion relation considerably. As a result, by placing several of them side-by-side, one can do beam-steering in the H-plane without the formation of grating lobes when scanning close to the grazing angle.

The excitation and the termination of the waveguide is through two tapered sections at the two ends. The tapered sections consist of 10 posts with increasing heights in 10 linear steps as shown in Fig. 4.a. The simulated and measured return loss of a prototype of the proposed waveguide excited through the tapered section is shown in Fig. 4.b. The top wall of the waveguide is loaded by slots with appropriate lengths to achieve the desired radiated power. The slots are placed periodically after a specific number of fundamental unit cells (N). Introducing the slots increases the period of the slow-wave waveguide by a factor of N. Therefore, we first derive the dispersion relation of the slow-wave waveguide without the slots but assuming an increased period by a factor of N. The desired dispersion is composed of N truncated parts of that of the fundamental unit cell as shown in Fig. 5 for 2 different values of N. As shown, N modes are associated with the N truncated parts of the first mode of the fundamental unit cell in each case. In order to have a continuous backward-to-forward scanning, modes 2 and 3 should be used for radiation (as highlighted in Fig. 5). Moreover, the change in the wavenumber by varying the frequency becomes very fast at the ending part of the dispersion diagram for the fundamental unit cell. Therefore, in order to avoid this part of the dispersion diagram which has very high dynamics, the radiating modes should be associated with the middle part of it. As a result, we need to use at least 4 unit cells between consecutive slots to achieve this goal (see Fig. 5). On the other hand, in order to avoid the formation of grating lobes near the grazing angles, slot spacing cannot be more than 4 unit cells. Therefore, we choose 4 unit cells as the slot spacing.

In the next step, we investigate the effect of introducing the slots on the dispersion of the waveguide with the increased periodicity by a factor of 4. Adding the slots after each 4 unit cells introduces a stopband at the 3 boundaries of the truncated sections in the dispersion diagram as shown in Fig. 6. In order to achieve a continuous backward-to-forward
scanning, one needs to close the open stopband between modes 2 and 3. Due to the strong field confinement in the middle of the waveguide and the limited space in this region, it is not possible to close the stopband by adding another post to cancel the reflection from the slots. Therefore, we need to perturb the structure itself to achieve this goal. Nevertheless, due to the small size of the features in the unit cell compared with the fabrication tolerance, the closed stopband becomes very sensitive to any fabrication error. As a result, it is of the utmost importance to reduce the stopband and its sensitivity to the variations simultaneously.

III. REDUCING THE STOPBAND AND ITS SENSITIVITY

In order to reduce the stopband and its sensitivity to the fabrication error simultaneously, we need more degrees of freedom to perturb the unit cell. Therefore, we allow the posts to be elliptical so that we can change their 2 radii along its minor and major axes independently. As shown in Fig. 1, the smallest features of the unit cell are the radii of the posts and the spacing between them. Therefore, we focus on minimizing the stopband by perturbing different features, as well as, reducing its sensitivity to the variations in the radii and the spacing of the elliptical posts. We define the criteria for the maximum acceptable stopband and its sensitivity in the following.

According to Fig. 6, broadside radiation occurs at 32GHz because at this frequency, $4\beta d$ is equal to $2\pi$ where $4d$ is the slot spacing. Therefore, we assume a stopband of less than 0.2% fractional bandwidth at broadside radiation frequency as a closed stopband. Furthermore, we define the acceptable sensitivity of the stopband to be within the limit of 1% of the broadside radiation frequency with $+/−0.05\text{mm}$ change in a feature size. These definitions are equivalent to assuming a stopband of less than 62MHz as a closed stopband and a maximum stopband of 320MHz with $+/−0.05\text{mm}$ variation in the fabrication as an acceptable stopband sensitivity.

In order to make the unit cells with the slots symmetric and to make the optimization process easier, we placed each slot on the top wall in the middle of the consecutive 4 unit cells. In this way, due to the symmetry, the optimized cell should also be symmetric. Therefore, we change the height, spacing, and radii of the two middle posts and the two outer posts simultaneously resulting in 8 degrees of freedom as shown in Fig. 7.

We perturbed and optimized the features of the unit cells with slot lengths of 3.4mm and 3.8mm to achieve the smallest stopband and the lowest stopband sensitivity. The optimized feature sizes are tabulated in Table 1. It should be noted that due to a CNC machining constraint, we had to elevate the spacing between the outer posts by 0.3mm. Comparing the values in Table 1 with the dimensions in Fig. 1 shows that the perturbed values are close to the non-perturbed values. Therefore, a slight change in one of the features can have a huge effect on the closed stopband. As a result, reducing the sensitivity of the stopband to the fabrication errors becomes as important as reducing the stopband itself.

Fig. 8 shows the dispersion diagram of the perturbed unit cell with a slot length of 3.8mm. As shown, the stopband between modes 2 and 3 which occurs in the radiation region is closed. However, the other two stopbands, occurring at the mode transitions, become wider. This results in slightly higher dynamics of the dispersion relation at the two ends of the radiation region. The two ends of the radiation region

- **FIGURE 6.** The stopband at the boundaries due to the reflections from the slots with a length of 3.8mm.
- **FIGURE 7.** The unit cell of the slow-wave slotted waveguide. The features used for perturbation are shown.
- **TABLE 1.** The dimensions for the optimized unit cells.
correspond to radiation angles at 0° and 180° resulting in a complete 180° steering range. However, the achieved steering range is limited by the operating frequency-band and the formation of grating lobes which will be discussed in the next section.

Fig. 9 shows the bandwidth of the stopband for different variations in the size of $D_{22}$, $D_{21}$, and $d_2$ for the optimized perturbed unit cell with the slot length 3.8mm. It should be noted that $D_{22}$ and $d_2$ are the smallest features in the unit cell. As shown, the stopband with no fabrication error is 33 MHz which is consistent with the defined criteria for a closed stopband. Furthermore, the stopband remains less than 1% of the center frequency even with a fabrication error of $+/−0.075$mm which is equivalent to a 10% fabrication variance in $D_{22}$. Fig. 9.a also shows that both the stopband and its sensitivity can be optimized considerably better when using elliptical posts instead of cylindrical. As shown, the sensitivity of the stopband to the fabrication error for the diameters of the elliptical posts along their minor and major axes is considerably lower compared to that of the case when using cylindrical posts.

Physically, elliptical posts provide additional capacitance periodically along the waveguide and confine the fields in the middle of it. Therefore, they should be placed in the middle of the magnetic walls. Placing them on the electric walls does not allow achieving these goals. Furthermore, the elliptical shape of the posts provides more degrees of freedom in order to close the stopband with a reduced sensitivity. The longitudinal diameter of the elliptical posts controls the coupling between the evanescent fields in the vicinity of them that constitutes the propagating Floquet mode as discussed in [10]. On the other hand, the transverse diameter of the posts tunes the additional capacitance introduced by them periodically in the waveguide. By changing these two parameters simultaneously, we could effectively close the stopband and reduce the antenna sensitivity to fabrication errors.

It should be noted that the stopband could be reduced even further to a considerably smaller value than 33 MHz. However, the sensitivity of the stopband would be more than the specified limit. Therefore, the optimized unit cell is achieved by simultaneously reducing the stopband and its sensitivity such that both of them remain within their specified limits.

IV. ARRAY DESIGN

The proposed antenna is a 40-element uniformly excited linear array. The ratio of the radiated power over the incident power is shown in Fig. 10 for different slot lengths at 33 GHz. In order to achieve a uniform excitation, we have to increase the slot length as we move away from the feed point. As a result, the designed antenna array is composed of 20 optimized unit cells with a slot length of 3.4mm followed by 20 optimized unit cells with a slot length of 3.8mm. The two ends of the antenna are connected to the tapered section shown in Fig. 4 for excitation and termination.

The directivity of the antenna can be found from equations 1-3 [11]. $N$ is the number of slots, $4d$ is the slot spacing, $\beta$ is the guided wavenumber, and $k_0$ is the wavenumber in vacuum. $EF$ is the directivity of a single slot.

\[
D(\theta) = AF(\theta) \times EF(\theta)
\]

\[
AF(\theta) = \frac{\sin(N \psi/2)}{\sin(\psi/2)}
\]

\[
\psi = k_0 4d \sin(\theta) - \beta 4d
\]
The radiation angle can also be found from equation 4. Using (4) and the dispersion relation shown in Fig. 8, the calculated radiation angle for different frequencies is shown in Fig. 11. As shown, the antenna beam steers from $-33^\circ$ to $+35^\circ$ as the frequency is changed from 27 GHz to 34.7 GHz. The steering range shown in Fig. 11 is limited to $-33^\circ$ from the lower frequencies by the operating frequency-band. For frequencies below 27 GHz, the return loss increases substantially due to the proximity to the cut-off frequency of WR22. On the other hand, the steering range is limited to $+35^\circ$ from the high frequencies by the formation of grating lobe.

V. SIMULATION RESULTS FOR STRUCTURES WITH RANDOM FABRICATION ERRORS

The stopband and a unity reflection coefficient occurs in an ideal infinitely long periodic structure which is simulated using Eigenmode solvers. However, the proposed antenna has a finite length and the fabrication is a stochastic process. Therefore, we need to study the effect of random fabrication errors on the characteristics of the proposed antenna.

We assume a triangular probability density function for the fabrication error of all the features in the proposed unit cells. Fig. 12 shows the probability density functions for three different fabrication tolerances. We constructed different antenna samples composed of 20 optimized unit cells placed side-by-side with a random fabrication error using the given probability density functions. The random fabrication error for each feature in each unit cell is assumed independent of the others. We used three different tolerances of 25 $\mu$m, 50 $\mu$m, and 100 $\mu$m for the fabrication errors as shown in Fig. 12. Finally, we compared the simulation results for their return loss with that of the ideal proposed antenna.

Fig. 13.a shows the return loss of the ideal antenna composed of 20 unit cells with slot length 3.4 mm throughout the closed stopband. As shown, the maximum return loss is 14 dB and the average return loss throughout the closed stopband is 24 dB. On the other hand, Fig. 13.b, Fig. 13.c, and Fig. 13.d show the return loss for the same antenna but with a random fabrication error with a triangular distribution function for 3 different fabrication tolerance values of 25 $\mu$m, 50 $\mu$m, and 100 $\mu$m respectively.

The black dotted line shows the average values of $S_{11}$ calculated over the given frequency band including all the samples. Fig. 13 shows that the random fabrication error has
TABLE 2. Statistical parameters for the randomly generated samples.

| Variance samples | Maximum (dB) | Mean (dB) |
|------------------|--------------|-----------|
| 25 µm            | -10.18       | -9.40     |
| 50 µm            | -5.23        | -5.15     |
| 100 µm           | -1.63        | -1.94     |

VI. SIMULATION AND MEASUREMENT RESULTS

Fig. 14 shows a picture of the fabricated antenna. Two 2.92mm coax-to-WR22 adapters are attached at its two ends. The left adapter is used for excitation while the right adapter is connected to a matched load to absorb all the remaining power and prevent any reflections. As shown, the total length of the antenna is 276mm and its width is 40mm while the width of the waveguide is only 5.69mm. The additional width is to support the flanges.

Fig. 15 shows a picture captured by a laser microscope from a unit cell of the fabricated antenna. Analyzing the data from the microscopic imaging shows that the fabrication error varies from +48 µm to −45 µm with a distribution similar to a Gaussian function but not equal. This is because a Gaussian distribution allows for an infinite range of fabrication error. In order to limit the fabrication error to the given tolerance and to consider the worst-case scenario, we chose a triangular distribution function for the simulations.

The measured normalized radiation pattern for the fabricated antenna is shown in Fig. 16. The beam is steered from −38° to +27° by changing the frequency from 27 GHz to 34.7 GHz. As shown, the antenna half-power beamwidth is 1.75° and the maximum sidelobe level is −10 dBi. Fig. 17 shows the measured and simulated gain of the antenna as well as the radiation angle for different frequencies. The relation between the radiation angle and the frequency is a linear function which is necessary for frequency-scanned frequency-modulated continuous wave (FMCW) radars. The antenna has a maximum gain of 19 dBi at 33.5 GHz. The

FIGURE 14. The fabricated antenna.

FIGURE 15. Laser microscope image from a unit cell in the fabricated antenna.

FIGURE 16. Measured normalized radiation pattern (frequencies are in GHz).

FIGURE 17. The measured gain and radiation angle versus frequency.
gain of the antenna changes between 11.5 dBi and 19 dBi throughout the operating frequency range. The measured gain of the antenna is lower at lower frequencies. This is for two reasons. First, due to the smaller electrical length of the slots at lower frequencies, they radiate less power at this frequency range. Therefore, there is more non-radiated power that is absorbed by the matched load at lower frequencies. This can be redeemed by adding optimized unit cells with a larger slot length. However, another reason for the lower gain at lower frequencies is the lower radiation efficiency at this frequency range. Fig. 18 shows the simulated radiation efficiency of the designed antenna. As shown, the radiation efficiency changes from 54% to 90% and it is lower at lower frequencies. This is because the lower frequency range is closer to the cut-off frequency where the Ohmic loss is higher in the waveguide. This is analogous to the higher conductor loss in rectangular waveguides closer to the cut-off frequency which is due to the increased number of the times the wave is reflected from the walls per unit length of the waveguide. Lastly, it should also be noted that the electrical length of the antenna is shorter at lower frequencies which also contributes to a lower gain.

Fig. 19 shows the measured and simulated half-power beamwidth of the designed antenna versus frequency. As shown, the HPBW has its minimum value of 1.5° around the broadside radiation angle and increases as the beam is scanned away from broadside. This is due to the decrease in the effective radiating length of the antenna.

Fig. 20 compares the simulated and measured radiation patterns for 3 sample operating frequencies. As shown, both the maximum gain as well as the radiation angle match very well between the two. The low-level back-lobe at 27 GHz results which was 0.25°. As shown, the HPBW has its minimum value of 1.5° around the broadside radiation angle and increases as the beam is scanned away from broadside. This is due to the decrease in the effective radiating length of the antenna.

Fig. 21 shows the measured $S_{11}$ and $S_{21}$ for the fabricated antenna. As shown, $S_{11}$ is less than $-10$ dB throughout the operating frequency range except in the vicinity of the broadside radiation where it reaches to a maximum value of $-8.5$ dB. This figure also compares the S parameters for the structure with and without screws.
fastening the upper and lower parts after assembly. There are 33 screws located on each side of the waveguide junction in order to minimize the gap between the two parts. The slight change observed shows that the fabricated antenna can operate very well even without any screws which is due to the current confinement in the middle of the waveguide as shown in Fig. 3.

Considering Fig. 18 and Fig. 21, one can derive the detailed power budget of the proposed antenna. According to Fig. 21.a, less than 10% of the input power is lost through reflections at the input port, except in the vicinity of broadside radiation where the return loss accounts for less than 13% of the input power. Furthermore, as shown in Fig. 21.b, less than 10% of the input power is dissipated at the matched load at the end of the antenna. Finally, according to Fig. 18, the power dissipated in the form of Ohmic loss varies by changing the frequency. The Ohmic loss in the proposed antenna ranges from 10% up to 45% of the available power for radiation. As mentioned before, the higher Ohmic loss at lower frequencies is due to the proximity to the cut-off frequency where the Ohmic loss is increased.

VII. COMPARISON AND DISCUSSION

Table 3 compares the characteristics of some of the relevant frequency-scanned antennas with the proposed fabricated antenna. The antennas in [15], [18], [19], and [20] are implemented using substrate-integrated waveguide (SIW) technology while [22], [23], and [10] are metallic waveguides. The advantages of SIW-based antennas are the easier integration with other system components as well as the high precision in the PCB fabrication (+/− 0.01 mm). Furthermore, due to the increased flexibility in the design and fabrication process of substrate-based leaky-wave antennas, they can be designed with higher order modes and operating at more than one bands [21]. However, they suffer from a substantial dielectric loss at high frequencies which does not allow for extending their length for applications that require a narrow beamwidth and long electrical length. This is why a considerable difference can be seen between the electrical lengths of the antennas in the two categories in the table. For leaky-wave antennas, in order to have an acceptable radiation efficiency, the attenuation factor of the field inside the waveguide needs to be mostly due to the field leakage for radiation compared to the loss in the waveguide. Increasing the radiating length requires decreasing the leakage of the wave for radiation. This results in the dielectric loss mechanism to dominate the attenuation factor at much smaller electrical lengths in substrate-based leaky-wave antennas and results in a substantial decrease in their radiation efficiency. Therefore, in general, SIW leaky-wave antennas are not suitable for length extension. However, SIW antennas have a smaller width than half-wavelength in vacuum due to the higher substrate permittivity. This allows placing them side-by-side to narrow the beam or perform beam-steering in the non-scanning plane.

The antennas in [22], [23] are implemented using a micro-machining technique. As they do not use any substrates, they can be designed and fabricated with a large electrical length and a relatively high radiation efficiency. However, due to their large width, they cannot be placed side-by-side to narrow the beam or perform beam-steering in the non-scanning plane. For example in [22], Sarabandi et al. have demonstrated the design and implementation of a frequency-scanned serpentine waveguide. In order to narrow the beamwidth in the non-scanning plane, they have placed a patch array along each meandered branch. Using this method, they have considerably decreased the beamwidth in the non-scanning plane. However, this method is limited by the length of the meandered branches which is fixed by the designed dispersion relation of the waveguide. Furthermore, one cannot perform any beam-steering in this plane using this approach.

In [23] a gold-plated micro-machined antenna is proposed. The antenna is composed of a parallel-plate waveguide loaded with periodic slots forming a LWA. This antenna can achieve a high radiation efficiency using modified thin-film fabrication techniques that can reduce the tolerances. Nevertheless, the achieved high radiation efficiency is also partially due to the gold metalization used in this work. The gold conductivity is 4.52 × 107 S/m while that of the 6061 aluminum alloy used for the fabrication of the proposed antenna is 2.86 × 107 S/m which is about 40% less. Moreover, the antenna cannot scan through broadside which results in an increase in the HPBW as the beam is scanned close to the grazing angles.
TABLE 3. Comparison with other frequency scanning antenna arrays.

| Ref. | Antenna type | Bandwidth | Scanning range | Maximum gain | Efficiency | Antenna length | Antenna width | Broadside radiation | Suitable for 2D scanning | Suitable for length extension |
|------|--------------|------------|----------------|--------------|------------|----------------|---------------|-------------------|---------------------------|-----------------------------|
| [15] | SIW          | 16.1% (13.2-15.6 GHz) | 103° (-61° − +42°) | 13 dBi | 30%-60% | 11.1λ | 0.37λ | Yes | Yes | No |
| [18] | SIW          | 36.6% (7.6-11 GHz) | 119° (-74° − +45°) | 13.5 dBi | 70%-90% | 7.9λ | 0.42λ | Yes | Yes | No |
| [19] | SIW          | 5.7% (34-36 GHz) | 41° (-17° +58°) | 13 dBi | 20%-50% | 8.2λ | 0.3λ | No | Yes | No |
| [20] | SIW          | 16.7% (55-65 GHz) | 120° (-72° − +48°) | 14.5 dBi | 50%-71% | 5.6λ | 0.4λ | Yes | Yes | No |
| [22] | Micromachined Waveguide | 6.3% (230-245 GHz) | 50° (-25° − +25°) | 30.5 dBi | 55%-65% | 35.6λ | 6.7λ | Yes | No | Yes |
| [23] | Micromachined Waveguide | 30% (220-300 GHz) | 45° (-75° − +30°) | 28.5 dBi | 80%-89% | 18λ | 18λ | No | No | Yes |
| [10] | Waveguide    | 14% (32.5-37.4 GHz) | 60° (-72° − +12°) | 17 dBi | 74%-92% | 29λ | 0.69λ | Yes | Yes | No |
| This work | Waveguide | 25% (27-34.7 GHz) | 65° (-38° − +27°) | 19 dBi | 54%-90% | 27.6λ | 0.57λ | Yes | Yes | Yes |

In [10] a slow-wave antenna is presented using cylindrical posts on the top and bottom walls facing each other. The antenna is fabricated using CNC machining. This antenna has 60° steering range and its radiation efficiency changes between 74% and 92%. However, it is not able to scan through broadside resulting in a larger HPBW close to the grazing angles.

VIII. CONCLUSION

In this paper we discussed the design and implementation of a new frequency-scanned leaky-wave antenna operating from 27 GHz to 34.7 GHz. The antenna can steer its beam from $-38^\circ$ to $+27^\circ$ including at broadside. The designed antenna has a radiating length of 27.6λ resulting in a 1.75° HPBW at broadside. The closed stopband has a low sensitivity to the fabrication error for the smallest feature sizes. This allows for fabricating the antenna using CNC machining with a +/-0.05 mm fabrication tolerance for the elliptical posts and a regular +/-0.13 mm fabrication tolerance for all other features. The antenna simulated radiation efficiency changes from 54% to 90% throughout the steering range. One important characteristic of the designed antenna is the fact that the fields and the current are confined in the middle of the designed slow-wave waveguide. This reduces the Ohmic loss due to the gap in the middle of the side walls after assembly. Furthermore, due to this confinement, the width of the proposed waveguide can be further reduced and several of them can be placed side-by-side to perform beam-steering in two dimensions (e.g. frequency scanning along the longitudinal direction and electronic scanning along the transverse direction).

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A. Ohadi, G. V. Eleftheriades: Slotted Waveguide Frequency-Scanned Slow-Wave Antenna With Reduced Sensitivity

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