A Planar SIW-Based mm-Wave Frequency-Scanning Slot Antenna Array with no Scan Blindness at Normal

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Abstract—Planar antenna reflectors are the modern design trend of both multibeam and frequency-scanning antenna arrays. The planar implementation of reflectors is typically performed using substrate-integrated waveguide (SIW) technology. A reflector’s profile can be different from the canonical one (parabolic, elliptic, hyperbolic, etc.) because of the effect of spatial dispersion of the reflection coefficient of the SIW-based surface. It should be synthesized considering the magnitude and argument of the field reflected from such a surface to maximize the efficiency of the reflection. In this paper, we present a planar mm-wave slot antenna array with SIW-based horn-reflector feeding. We analytically formulate the optimization of the SIW surface dimensions while accounting for the spatial dispersion of the reflection coefficient. We minimize the dimensions of the planar horn-reflector feeding. Finally, we demonstrate that using a dual-slot radiating element, we can avoid the effects of scan blindness along the normal direction. A prototype has been built and a good agreement has been achieved between the measured results and the predicted results based on calculations. The prototype achieved ±17° beam scanning within 16% of the operational frequency range, with no scan blindness along the normal direction.

Index Terms—frequency-scanning antenna array, horn-reflector antenna, scan blindness effect, spatial dispersion of the reflection coefficient, surface integrated waveguide technology

I. INTRODUCTION

Modern mm-wave telecommunication systems have necessitated the miniaturization of antennas and microwave feeding networks. Substrate-integrated waveguide (SIW) technology is becoming a popular choice for substrate-based planar multibeam and frequency-scanning antenna arrays and reflectors [1–7]. Previously, studies have investigated an offset parabolic reflector [1], Gregorian system [2], and pillbox parabolic reflector [3–7]. However, none of these studies have considered the effect of the spatial dispersion of the reflection coefficient when synthesizing the reflector profile. Instead, the profiles were selected to resemble that of a continuous metallic surface. In [8], we demonstrated that the reflection properties of SIW-based surfaces generally depend on the incidence angle of an incoming wave. Therefore, accounting for the effect of spatial dispersion of the reflection coefficient would increase the radiation efficiency of a planar antenna reflector. Because the spatial dispersion effect can significantly influence the reflection properties of offset parabolic reflectors, such reflector designs must be corrected for this effect. Moreover, the slot-based frequency-scanning arrays in previous studies used a single slot at the radiation element, which leads to scan blindness along the normal.

This study extends our analytical work in [8] to the design, fabrication, and characterization of a planar mm-wave slot antenna array having horn–reflector feeding. The antenna array comprises a SIW-based horn–reflector feeding system and radiating slots (Fig. 1). We optimize the parameters of the SIW-based feeding horn–reflector and radiating-slot aperture to maximize the array’s efficiency. We propose a dual-slot radiation element that avoids blindness, along the normal. Finally, we demonstrate the calculated and measured characteristics of a fabricated prototype of the antenna array.

Fig. 1. Structure of the antenna array.

SIW-based surfaces generally depend on the incidence angle of an incoming wave. Therefore, accounting for the effect of spatial dispersion of the reflection coefficient would increase the radiation efficiency of a planar antenna reflector. Because the spatial dispersion effect can significantly influence the reflection properties of offset parabolic reflectors, such reflector designs must be corrected for this effect. Moreover, the slot-based frequency-scanning arrays in previous studies used a single slot at the radiation element, which leads to scan blindness along the normal.

This work was funded by King Abdullah University of Science and Technology (KAUST) and King Saud University (KSU) Collaborative Research Grant.

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II. THE HORN-REFLECTOR FEEDING DESIGN

As the antenna substrate, we selected Rogers 4533 laminate, which has a relative permittivity of \( \varepsilon_r = 3.3 \), a loss tangent of 0.002, and a thickness of 0.762 mm. In the standard excitation configuration, the excitation horn is located at a certain distance from the SIW-based offset parabolic reflector (Fig. 2(a)), but here we design a horn–parabolic reflector as shown in Fig. 2(b).

The horn–reflector design connects the horn to the parabolic reflector, thus forming a single radiation system that eliminates the possibility of direct reception or radiation of energy by the irradiator. This design is important for preventing the sharp weakening of signals received from the direction opposite the main direction. The horn–parabolic reflector prevents energy loss from the irradiator to the reflector because there are no dissipating metallic elements. Energy is dissipated only at the edges of the reflector aperture. This design provides a broadband operation mode and a low level of side lobes. However, horn–parabolic antennas are rarely mentioned in the literature [9-13]. The few studies present conical and pyramidal horn–reflector antennas and their theoretical and experimental investigations, but the flare angle and other antenna parameters were selected based on engineering intuition without scientific justification. The use of a planar SIW-based horn–parabolic reflector raises certain important issues. The first issue relates to the SIW implementation of the reflector. In general cases, the SIW structure causes spatial dispersion of the reflection coefficient. In particular, the parameters of the SIW structure and the angle of incidence of an incident wave affect the phase of the reflection coefficient. This phase dependence should be considered when designing the feeding horn and the reflector profile. The second issue concerns the design of an optimal feeding horn. To minimize the horn length, the flare angle should be as large as possible. However, the horn mode must be propagated to ensure a completely cylindrical wave front at the horn’s aperture, which limits the maximum flare angle.

![Fig. 2. SIW-based (a) offset reflector fed by a horn, and (b) horn–reflector design.](image)

![Fig. 3. A periodic grating.](image)

A. Reflection from the SIW-based Surface

We consider a 2D periodic grating of circular-shaped conductors (Fig. 3) located in a medium with a permittivity of \( \varepsilon \). The radius of the conductors is \( r \), and the grating period is \( p \).

A plane wave is incident on the grating at angle \( \phi_i \). The electric field intensity vector \( \mathbf{E} \) of the incident wave is parallel to the conductors. The reflection coefficient \( R \) of the plane wave from the grating is determined from the fill factor \( q = 2r/p \) as follows [8]:

\[
R = \frac{1 + ik\sqrt{\varepsilon} \cos \phi_i p l_1}{2 - 2ik\sqrt{\varepsilon} \cos \phi_i p l_1},
\]

where \( k = 2\pi/\lambda \), \( \lambda \) is the radiation wavelength in a vacuum, \( i \) is the imaginary unit, and

\[
\begin{align*}
  l_1 &= \pi q \left[ \left( \frac{1}{2} \right) - \ln \left( \frac{\cosh l_1}{\lambda} \right) / \pi \right], \\
  l_2 &= \pi q \left[ \left( \frac{1}{2} \right) - \ln \left( \frac{\sinh l_1}{\lambda} \right) / \pi \right], \\
  l_3 &= \ln \left[ \sin \left( \frac{1 + v}{2} \right) \pi q / 2 \right] / \ln \left( \frac{1}{\pi q / 2} \right), \\
  l_4 &= \pi q / 2 + \pi q \arctan \left( \tan \left( \pi q / 2 \right) \cot \left( \pi q / 2 \right) \right) / l_3,
\end{align*}
\]

with \( v \) being determined from the following non-linear equation:

\[
\pi q \sin \left( \pi q / 2 \right) = 2 \left( \sin^2 \left( \pi q / 2 \right) + \sin^2 \left( \pi q / 2 \right) \right) l_3.
\]
The parameters $l_1$ to $l_4$ in (2), are used for simplify (1), and it should be noted that they have no actual physical meanings. The dependences of the magnitude and argument of the reflection coefficient on fill factor are plotted for different angles of incidence in Fig. 4. From this figure we can see that the magnitude of reflection coefficient practically is equal to 1 when the fill factor is greater than 0.2 for any angle $\phi$. Therefore, the SIW-based horn-reflector feeding can be implemented through the circular-conductor grating for any $q > 0.2$. The argument of the reflection coefficient largely depends on incident angle for all filling factors except $q = 0.2$ and 0.5. When $q = 0.2$ and 0.5, the argument of $R$ was similar at all incident angles. Therefore, we selected $q = 0.5$ for implementing the horn-reflector feeding because at this fill factor, we can ignore the effect of spatial dispersion of the reflection coefficient. This effect should be considered when selecting the reflector profile because it influences both the efficiency of reflection from the SIW-based reflector [8] and the propagation mode in the horn feeding. Moreover, the condition of the optimal flare angle, considered in the following paragraph, requires the absence of this effect.

Note that these results are valid not only for the case of a linear grating and a plane wave incidence but also for a curved wave front when its radius of curvature is much larger than the grating period $p$. If an incident field is different from a plane wave but can be represented through a spectrum of plane waves, the obtained results are also valid.

B. Optimal Parameters of the Horn and Reflector

Let us simulate the horn-reflector feeding as the point source of a cylindrical wave (Fig. 5). The cylindrical wave front is incident on the reflector and transforms into a plane wave front after reflection. As the fill factor was selected to obtain the same argument of the reflection coefficient at different incident angles, the reflector has a canonical parabolic profile with focal distance $F$ (Fig. 1) determined from the radius vector $r(\psi) = 2F(1 + \cos \psi)$ for $\psi \in [\pi/2 - \gamma_1, \pi/2 + \gamma_1]$. To develop the cylindrical wave front at the aperture of the horn feeding, the dimensions of the waveguide section $a$ and the flare angle of the horn $\gamma$ should meet the condition of first-mode propagation only. The other dimensions $h$ and $F$ are selected from the required beam width of the antenna array. However, to miniaturize the feeding network dimension $h$, the flare angle should be as wide as possible. We set both the width and length of the waveguide section to $a = 0.9\lambda_g$ (where $\lambda_g = \lambda / \sqrt{\epsilon}$ is the guided wavelength), which allows only first-mode propagation in the section. The flare angle $\gamma$ should be selected such that the wave front at the aperture of the horn is purely cylindrical within the required operational bandwidth. To obtain a purely cylindrical wave front, the congruence of Brillouin wave rays of the feed waveguide section must be transformed to a congruence of the rays of the divergent horn mode having negligible excitation coefficients of other horn modes and reflected waveguide modes entering the waveguide. The condition of conversion of the first propagating waveguide mode to the horn mode with an almost purely cylindrical wave front was obtained in [14, 15] in the geometric optics approximation. Such conversion occurs when the horn’s flare angle satisfies the following condition:

$$\gamma \leq \sin \gamma_B \cos \gamma_B,$$

(3)

where $\gamma_B = \arcsin(\lambda g / (2a))$ is the Brillouin angle of the propagating waveguide mode. We choose to maximize the flare angle as it helps shrink the length of the horn structure, and reduce the propagation losses in the horn as well.

III. DESIGN OF THE RADIATING SLOTS

Because the considered array is a leaky wave antenna, the geometry of the radiating elements determines the amplitude distribution on the radiating aperture. Leaky wave antennas have two major amplitude distributions on the radiating aperture: exponential and uniform. The exponential amplitude distribution is developed when the coupling between a feeding network and each radiating element is equal, whereas a uniform amplitude distribution typically requires variable coupling to some extent. Exponential amplitude distribution does not ensure the optimum aperture efficiency of an antenna as compared to a uniform amplitude distribution. The gain loss of an antenna with the exponential amplitude distribution is approximately 0.9 dB as compared to the uniform one [16]. However, with the increase in the gain for the uniform distribution, the operational bandwidth decreases. In this paper, we use the exponential distribution because our first goal is to make the antenna wideband. The decrease in gain can be compensated by increasing the number of radiating elements $N_r$ along the $y$-axis. Moreover, the operating bandwidth depends on the number of radiating elements $N_r$ along the $x$-axis. Increasing $N_r$ narrows the bandwidth of the antenna.

Previous studies on planar antennas [1-6] used a single radiating slot as the radiating element to simplify the planar implementation. However, a single-slot antenna is disadvantaged by scan blindness when the array’s beam is
normal to the radiating aperture. This effect causes an extremely high reflection coefficient because of simultaneous reflection from each radiating slot. To avoid this effect, we adopted a dual-slot radiating element (Fig. 6a) in our antenna design.

Let us optimize the dimensions of the slots to minimize the reflection from the slots within the 27–33 GHz range. For this purpose, we consider a radiating array having an infinite number of elements along the y-axis (N_y→∞) and N_x = 25 elements along the x-axis. We assume that the width of each slot is considerably smaller than the guided wavelength, i.e., w_1,2<<λ. The radiating array is excited by the TEM mode of a parallel-plate waveguide propagating along the x-axis. The infinitely thin planes of the waveguide at x = 0 and x = z (Fig. 6b) are perfectly electrically conducting (PEC) planes.

To identify the S-parameter matrix of the radiating array, we applied a numerical-analytical procedure based on the method of integral equations. For this purpose, we considered a model of uninterrupted PEC planes located at z = 0 and z = -h with magnetic currents at the slots’ locations. We then have two areas separated by the PEC plane at z = 0: the waveguide in the region -h<z<0 and a half-space in the region z>0. The magnetic currents flowing on both sides of the PEC plane at z = ±h are similar and opposite to satisfy electric-field continuity in the slots.

For narrow slots, we assume that E_y≫E_z and the component E_z vanishes. Therefore, only a y-component of the magnetic currents exists inside the slots. Let us now ensure continuity of the x- and y-components of the magnetic field intensity vector H_{x,y} in the slots. For narrow slots, we assume that H_z≪H_y and the component H_z vanishes. We write a system of integral equations describing the excitation of the structure using the magnetic currents I_{1,2}(x, y), where 1 or 2 is the number of the slot in the radiating cell (Fig. 6a) and n is the number of a radiating row along the x-axis (n = 1, 2, ..., N_x). For this purpose, we use some known expressions of the Green function for the half-space and the plane waveguide, which were obtained for one cell periodic along the y-axis. We obtain the following system of integral equations after satisfying the condition of continuity of H_z in the slots:

\[
\sum_{m=1}^{N_x} \int_{S_1} I_{1m}(x', y') G(y, y', x + n P_x, x' + m P_x) dx' dy' + \\
+ \sum_{m=1}^{N_x} \int_{S_2} I_{2m}(x', y') G(y, y', x + n P_x, x' + m P_x) dx' dy' = \\
= H_{yo}(x + n P_x), \quad x, y \in S_1, \quad n = 1, 2, \ldots, N_x,
\]

(4)

Here,

\[
S_{(2)}: \quad |x ± a| < w_{(2)}/2 \quad \& |y| < L_{(2)}/2,
\]

\[
G = G^{\text{ho}} + G^{\text{pw}},
\]

\[
G^{\text{ho}}(y, y', x_n, x'_n) = \sum_{p,q} \int_{\gamma_{p,q}} \coth(\xi_{p,q} h) \left( k^2 - \kappa^2_{p,q} \right) e^{-i\kappa_{p,q}(y - y')} dx',
\]

\[
G^{\text{pw}}(y, y', x_n, x'_n) = \sum_{p,q} \int_{\gamma_{p,q}} \left( k^2 - \kappa^2_{p,q} \right) e^{-i\kappa_{p,q}(y - y')} dx',
\]

where \(Z_0\) is the wave impedance in a vacuum. Equation (4) is then summed over the indices \(p\) and \(q\) from minus to plus infinity, and the integration of \(\kappa\) is considered over \(-\pi/P_x<\kappa<\pi/P_x\). The system of integral equations (4) is then solved using the Galerkin method. The magnetic currents are then decomposed into the following series:

\[
I_{1(2)n}(s) = \sum_{m=1}^{N_x} C_{1(2)m,n} F_{1(2)m}(s).
\]

Here, the basis functions \(F_{1(2)m}(s)\) have the following form:

\[
F_{1(2)m}(s) = \sin \left( \frac{\pi r(y + 1/2)}{L_{(2)}} \sqrt{\frac{w_{(2)}^2}{2} - (x ± a)^2} \right).
\]

In accordance with the Galerkin method, the test functions coincide with the basis functions. Projecting system (4) to
system (5), we have the following system of linear equations with unknown coefficients $C_{1(2)r,m}$:

$$
\sum_{r=1}^{N_r} \sum_{m=1}^{N_m} C_{1r,m} Z_{11r,m} + \sum_{r=1}^{N_r} \sum_{m=1}^{N_m} C_{2r,m} Z_{12r,m} = Q_{011,0},
$$

$$
\sum_{r=1}^{N_r} \sum_{m=1}^{N_m} C_{1r,m} Z_{21r,m} + \sum_{r=1}^{N_r} \sum_{m=1}^{N_m} C_{2r,m} Z_{22r,m} = Q_{022,0},
$$

$n = 1,2,... N_r$, $s = 1,2,... N_m$,

$$
Z_{ii,j,r,m,n,m} = \int_5^5 \int_5^5 \int_5^5 \int_5^5 F_{bii,a} (s_n) G(s_n, s_m') F_{bji,r} (s_m') dsndsds',
$$

$$
Q_{0ii,v,a} = \int_5 F_{bii,a} (s_n) H_{r0} (s_n) ds.
$$

The unknown coefficients $C_{1(2)r,m}$ and hence the magnetic currents are determined by solving the system of linear algebraic equations (7). The $S$-parameter matrix of the radiating array is then determined from the fields of magnetic currents.

Using the above procedure, we optimized the dimensions of the radiating element. During the first stage of the optimization process, we considered only one element along the $x$-axis: $N_x = 1$. The dimensions of the slots and the distance between them are chosen thus to minimize the reflection coefficient at the frequency band of 27–33 GHz. Thus, the period $P_s$ was selected to provide a primary beam normal to the radiating aperture at 30 GHz. At this frequency, a couple of slots are excited with the same mode similar to a co-phase situation, which enhances the matching of the reflection coefficients of the radiating array. The matching enhancement can be improved by optimizing the dimensions of the radiating element. During the second stage of the optimization process, the dimensions of the radiating elements were optimized on an array with multiple elements: $N_x = 25$. The optimal slot dimensions were determined as $P_s = 5$, $P_l = 4.7$, $w_1 = w_2 = 0.33$, $L_1 = 2.82$, $L_2 = 2.76$, and $d = 0.78$ mm. The calculated reflection-coefficient profile of the radiating array with the optimized dimensions is presented in Fig. 7. Both cases, single and double slots, have been simulated and it has been found that there is a difference of more than 5 dBs between the gains of the two at broadside direction. This confirms that the double slot approach resolves the problem of scan blindness at normal direction.

**IV. EXPERIMENTAL RESULTS**

The antenna array was designed as shown in Fig. 8. The array was implemented on a (170 × 230) mm$^2$ Rogers 4533 substrate. The dimensions of the horn–reflector feeding were as follows: $h = 60$, $a = 5.1$, $F = 42.8$ mm. The flare angle was set to $\gamma = 20^\circ$ under condition (3). The metalized vias creating the horn–reflector were 0.8 mm in diameter. The antenna was fed by a microstrip-to-waveguide transition shown in Fig. 8. To create the transition, some upper part of the copper film (close to feeding) was removed to build the microstrip to SIW transition. The numbers of cells along the $x$- and $y$-axis were $N_x = 25$ and $N_y = 16$, respectively.

Fig. 9 is an image of the fabricated antenna array, which was fed by a sub-miniature version A connector. Figs. 10–12 show the calculated and measured RF characteristics of the antenna array. As shown in the reflection-coefficient profile (Fig. 10), the antenna array had a 16% impedance bandwidth at the -10 dB level within the range of 25.7–31.6 GHz. Moreover, the profile was rugged. Such behavior occurs because the reflection coefficient results from the superposition of many reflected waves generated by irregularities located at large electric distances from each other. A large number of waves scattered from separate sources irregularly located in space (such as the junction and edges of the horn–reflector feeding and each slot of the radiating aperture) causes a stochastic (noise-like) shape of the total reflection coefficient.
Fig. 10. Measured reflection-coefficient.

Fig. 11. Measured radiation pattern at 29.3 GHz.

Fig. 12. Measured radiation patterns of the antenna array in the scanning plane.

Fig. 12 shows the measured radiation patterns in the two orthogonal E and H-planes at 29.3 GHz when the beam was normal to the radiating aperture. As seen in this figure, the realized gain was 20.5 dBi with a decent cross-polarization level (>25 dB). The measured half-power beamwidths at 29.3 GHz were $\theta_E = 10^\circ$ and $\theta_H = 12^\circ$ in the scanning plane (E-plane) and orthogonal plane (H-plane), respectively.

The measured half-power beamwidths at 29.3 GHz were $\theta_E = 10^\circ$ and $\theta_H = 12^\circ$ in the scanning plane (E-plane) and orthogonal plane (H-plane), respectively.

Fig. 12 shows the frequency-dependent beam-scanning properties. The peak gain was 16.9 dBi at 25.7 GHz with the main beam radiation at $\theta = 17^\circ$. As the frequency increased, the beam gradually shifted to the broadside direction (at 29.3 GHz, the realized gain was 20.5 dBi; Fig. 11) and the primary beam shifted in the negative $\theta$ direction. The peak gain decreased slightly to 18.6 dBi at 31.6 GHz, when the peak radiation was directed at $\theta = 17^\circ$. The scanning range within the 10-dB impedance bandwidth was $34^\circ$ ($17^\circ$ to $-17^\circ$) and the peak gain variation was $\sim 3.7$ dBi, with no scan blindness at the normal direction.

Compared Table I presents the operating frequency ranges and scanning ranges for various planar frequency-scanning slot arrays with SIW-based quasi-optical reflectors. From this table, we can see that the proposed antenna design provides the widest scanning range among the other planar SIW-based antenna arrays as well as the second highest impedance bandwidth. In addition, it is the only design which solves the problem of scanning blindness at normal direction.

The antenna efficiency was then calculated using the following Elliot’s formula for directivity [17]:

$$D = \frac{32400}{\theta_E \theta_H}$$

Here, the calculated value of directivity is expressed through the measured half-power beamwidth. We calculate the efficiency as the measured peak gain divided by the directivity calculated using (8). The calculated directivity was 24.3 dBi at 29.3 GHz, thus yielding a calculated efficiency of 42% within the operational bandwidth. The corresponding gain loss is $\sim 3.8$ dB. The gain loss was caused by nonuniform amplitude distribution on the radiating aperture, active loss in the dielectric, and matching loss. Note that this gain loss was assumed to be caused by the exponential amplitude distribution on the radiating aperture along the x-axis, as mentioned in Section III. The amplitude distribution on the radiating aperture along the y-axis shows a cosine shape with nulls at the ends. Such an amplitude distribution causes a gain loss of $\sim 1$ dB [18]. Accordingly, the active loss in the feeding structure combined with the matching loss should be $\sim 1.8$ dB.

### V. CONCLUSIONS

In this paper, we present the optimization procedure for parameters of the planar SIW-based frequency-scanning antenna array. We demonstrated that a dual-slot radiating element can avoid scan blindness along the normal direction. The fabricated antenna prototype achieved $\pm 17^\circ$ beam scanning within 16% of the operational frequency range and an efficiency of 42%.

### ACKNOWLEDGEMENT

This work was supported by King Abdullah University of Science and Technology and King Saud University Collaborative Research Grant.
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