Design of a Planar Slotted Waveguide Array Antenna for X-band Radar Applications

Rashid Ahmad Bhatti¹ · Byeong-Yong Park¹ · Yun-Taek Im² · Seong-Ook Park¹

Abstract

A planar slotted waveguide array antenna has been designed at 9.37 GHz for X-band radar applications. The antenna consists of multiple branchline waveguides with broadwall radiating shunt slots and a main waveguide to feed the branch waveguides through a series of inclined coupling slots. The antenna feed point is located at the center of the main waveguide. Element weights in the array have been calculated by sampling a continuous circular Taylor aperture distribution at the 25 dB sidelobe level in both the E and H planes. A commercially available electromagnetic (EM) simulation tool has been used to characterize the individual isolated slot and that data has subsequently been used to design the planar array. The array is finally analyzed in a CST Microwave studio and the measured and simulated results have been found to be in good agreement.

Key words: Antenna, Slot, Waveguide.

I. Introduction

Slotted waveguide array antennas are widely used for radars and communication systems [1]. They offer significant advantages in terms of weight, volume, and radiation characteristics. These antennas are very attractive due to their planar, compact, and rugged construction. A metallic waveguide is a low loss structure and it can handle high power. The dimensions of the slots in the waveguide walls can be controlled to realize the desired pattern shape. Longitudinal shunt slot arrays are also attractive due to their very low cross-polarization levels. The design of a resonant slotted waveguide array antenna is generally based on the procedure published by Elliot [2]–[4]. In these design procedures, inter-slot distance is one-half the guide wavelength and the waveguide end is short circuited at a distance of a quarter-guide wavelength from the center of the last slot. Linear arrays are placed adjacent to each other to realize planar arrays. A feed waveguide is placed beneath the branch waveguides containing radiating slots and power is coupled from the main waveguide to the branch waveguides through the series of inclined slots. The feed waveguide is terminated in the short circuit located at a distance of one-half of the guide wavelength from the center of the last inclined coupling slot. External and internal mutual coupling effects can be fully incorporated into the design procedure. Other major research works on linear and planar slotted waveguide antennas are [5], [6]. Isolated slot characterization is required in the design of slotted waveguide arrays. Isolated slot admittance can be measured experimentally [2] or computed numerically [7]–[9]. Care must be exercised in calculating the individual slot as minor errors might severely affect the overall array design.

In this paper, the design of a planar slotted waveguide array antenna based on the procedure in [3] is discussed. The design procedure includes both external mutual coupling effects and internal mutual coupling using commercial EM simulators to approach more exact results. The planar antenna has been designed at 9.37 GHz with shunt slots acting as radiators in the broadwall of a reduced height waveguide to suppress higher modes. The planar array consists of twelve linear arrays of shunt slots, which have been placed side-by-side with truncated corner slots to fit the antenna in a circular boundary. A main waveguide is run across the branch waveguides to feed them through the inclined-series coupling slots. Antenna feed is located at the center of the main waveguide.

II. Isolated Slot Characterization

In order to use the design procedure [3], it is im-
portant to generate isolated slot admittance as a function of slot length and the resonant length of the slot as a function of its offset. For feed slots, the resonant length of the slot as a function of its tilt angle is also required for the design of the array. A commercially available EM simulation tool, HFSS, has been used to generate the basic slot data subsequently used in the array design. In the HFSS simulations for the shunt slot, the round ended 2.5 mm wide slot was created in a 1 mm thick broadwall and admittance was recorded at a distance of one-half the guide wavelength from the center of the slot; the waveguide was terminated in a short circuit placed at a quarter-wave distance from the center of the slot. Slot length was varied and at each step of normalized complex admittance, conductance ($G/G_0$) and susceptance ($B/G_0$), was recorded and was normalized with resonant conductance of the slot. The design curve thus generated is shown in Fig. 1. The real and imaginary parts of the admittance are denoted as $h_1(y)$ and $h_2(y)$, respectively, where $y$ is the slot length normalized by the resonant length, $l/l_0$, which can be written as

$$Y(x,l) = \frac{(h_1(y) + jh_2(y))g(x)}{G_0}$$

The waveguide width and height are 21.5 mm and 5 mm, respectively. Next, the slot offset is changed and at each step, the slot resonant length is calculated where the imaginary component of the admittance was zero. The resulting resonant length of the slot as a function of offset is shown in Fig. 2.

We will denote the curve shown in Fig. 2 as $v(x)$, where $x$ is the offset of the slot from the center-line of the waveguide.

Normalized resonant conductance, $g(x) = G_1/G_0$, of the isolated shunt slot as a function of offset is plotted in Fig. 3. The curve $g(x)$ can be approximated as follows [7],

$$g(x) = K \sin^2 \left( \frac{\pi x}{a} \right)$$

$$K = 2.09 \frac{a/b}{\beta_{10}/k_0} \cos^2 \left( \frac{\pi \beta_{10}/k_0}{2} \right),$$

where $a$ and $b$ are the waveguide width and height respectively, and $\beta_{10}$ is the propagation constant of TE$_{10}$ mode and $k_0 = 2\pi/\lambda_0$.

The coupling slot was also characterized in HFSS for its resonant resistance as a function of the tilt angle. In the simulation model, waveguide was terminated at the distance of the half-guided wavelength from the center of the slot and resistance was recorded at the port located at the distance of the half-guided wavelength from

![Fig. 1. Normalized admittance of isolated shunt slot versus slot length.](image1)

![Fig. 2. Resonant length of the broadwall slot versus offset.](image2)

![Fig. 3. Normalized conductance of isolated shunt slot as a function of offset.](image3)
the center of the slot. At each tilt angle, the slot length was varied to make the imaginary component of the slot impedance equal to zero. The resulting curve is shown in Fig. 4.

**III. Planar Array Design Procedure**

The antenna design rests on the following design equation [2],

\[
\frac{Y_{na}}{G_0} = K_f f_{mn} \frac{V_{na}^*}{V_n^*} 
\]

where,

\[
K_f = \frac{1}{j(\alpha/\lambda)} \sqrt{\eta G_0 \beta_{mn}/k(\alpha)(k b)}
\]

and,

\[
f_{mn} = -K \left( \frac{\pi}{2 l_{mn}} \right) \cos (\beta_{mn} l_{mn}) \sin \left( \frac{\pi x_{mn}}{a} \right).
\]

In equation (2), \( V_n \) is the mode voltage in the \( mn \)th branch waveguide and \( V_{na}^* \) is the slot voltage of the \( mn \)th slot. \( Y_{na} \) is the active admittance of the \( mn \)th slot.

\[
\frac{Y_{na}}{G_0} = 2 f_{mn} \left( x_{mn}, l_{mn} \right) + MC_{mn}
\]

where,

\[
MC_{mn} = K \sum_{r=1}^{R} \sum_{s=1}^{S} V_{mn}^r g_{mnrs} \left( x_{mn}, l_{mn}, x_r, l_r \right)
\]

The typical configuration of the two slots in an array is given in Fig. 5.

In (3), \( MC_{mn} \) is the mutual coupling term for the \( mn \)th slot, which incorporates interactions from all of the slots in the array.

The aperture distribution has been calculated by sampling the continuous aperture at slot locations for the −25 dB first sidelobe level.

The planar array is depicted in Fig. 6. It consists of a twelve branch waveguide with feed waveguide placed...
behind.

For the \( mn^{th} \) slot in the array, the distance from the origin is calculated as \[9\],
\[
\rho_{mn} = \sqrt{(\frac{(2m-1)dz}{2})^2 + (\frac{(2n-1)dx}{2})^2},
\]
where \( dx \) and \( dz \) are the inter-element spacing along \( x \) and \( z \) directions.

The aperture distribution is then calculated as follows \[9\],
\[
A_{mn} = g_0(p_{mn}),
\]
where
\[
p_{mn} = \frac{\pi \rho_{mn}}{a},
\]
\[
g(p) = \frac{2}{\pi} \sum_{u=0}^{\infty} \frac{S_u J_u(\pi \mu_u)}{u^2} x_l,
\]
\[
S_u = J_u(\pi \mu_u) \prod_{n=0}^{u-1} \frac{1-\mu_n^2}{u_n^2},
\]
where \( \mu_n \) are zeros of \( J_u(\pi u) \).

The calculated aperture distribution is depicted in Fig. 7.

A 3-D array factor plot corresponding to the sampled aperture distribution is shown in Fig. 8.

After calculating the array weights, the following procedure is used to calculate the slot dimensions (offsets and lengths).

A set of initial offsets and lengths is assumed for the radiating slots. Resonant lengths and zero offset are a reasonable choice for starting the design. These offsets and lengths are used to calculate the following mutual coupling term equation (3),
\[
MC_{mn} = K_3 \sum_{r=1}^{\infty} \sum_{r'=m}^{\infty} \frac{V_{rr'}}{V_{mm}} g_{array}(x_{mn}, l_{mn}, x_{r'}, l_{r'})
\]
for every \( mn \) in the array.

A search is then conducted to find the couplet \((x_{mn}, l_{mn})\) that satisfies the following relation,
\[ \text{Im} \left[ \frac{2 \int f (x_m, l_m) Y}{G_0} \right] = - \text{Im} [MC_m], \]

where \( Y/G_0 \) is calculated from the design curves, as

\[ Y(x_m, l_m) = (h_1(y_m) + h_2(y_m)) g(x_m). \]

Once this is satisfied, then active admittance of the slot become spure real. During this search process, a continuum of couplets that satisfy this relation is found. Similarly, for the \( pq^{th} \) slot, the same search is conducted and the continuum of couplets that make the active admittance pure real is found. But for a given acceptable couplet for the \( mn^{th} \) slot, there will only be one couplet for the \( pq^{th} \) slot that will satisfy the following equation,

\[
\begin{align*}
\frac{Y_a}{G_0} &= f_{pq} (x_{pq}, l_{pq}) V_{pq}, \\
\frac{Y_m}{G_0} &= f_{mn} (x_{mn}, l_{mn}) V_{mn},
\end{align*}
\]

This procedure provides a set of acceptable couplets for each slot. However, which couplet to choose will depend on the required admittance level in each branch-line.

The final acceptable couplets in a branch-line should satisfy the relationship,

\[ \sum_{n=1}^{N_m} \frac{Y_a}{G_0} = C_m, \]

where \( C_m \) is the required admittance level in the \( mn^{th} \) branch waveguide.

The entire procedure needs to be iterated as the new slot data will give an improved mutual coupling calculation and the solution will converge after 3 or 4 iterations.

After calculating the radiating shunt slot data, the series feed slot data is then calculated according to the procedure reported in [9].

The sum of the series of slot resistances is given by,

\[ \sum_{n=1}^{M} r_m = W, \]

where \( r_m = KB^2 (m) \) and \( m=1, 2\cdots M, \)

\( r_m \) is the normalized resonant resistance of the \( m^{th} \) slot,

\( B \) is the slot excitation function,

\( K \) is the excitation normalization constant, and \( W=1 \) for the end-feed or \( W=2 \) for the center feed. \( B_m \) is given in terms of the aperture distribution.

\[ B^2 (m) = \sum_{n=1}^{N_m} A^2 (m,n), \]

where \( N_m \) is the number of shunt slots fed by the \( m^{th} \) series slot.

Using these resistance values, the corresponding slot angles are calculated using the design curves or the following relationship,

\[ R = \frac{0.131 \beta_{mn} \lambda_0^2}{ab} \left[ I_0 \sin \theta + \frac{\lambda_0}{2a} J_0 \cos \theta \right], \]

where \( \theta \) is the tilt angle of the series slot, and

\[ I_0 = \frac{\cos \left( \frac{\pi}{2} K_1 \right)}{1 - K_1^2} \]

\[ J_0 = \frac{\cos \left( \frac{\pi}{2} K_2 \right)}{1 - K_2^2} \]

\[ K_1 = \frac{\beta_{mn} \cos \theta + \lambda_0}{2a}, \]

\[ K_2 = \frac{\beta_{mn} \sin \theta}{2a}. \]

Once slot resistance as a function of the slot tilt angle is known, we can calculate the slot length for different slot angles from the design curve shown in Fig. 4. The applied slot length (l) is 16.5 mm and the offset lengths (x) are 2.1, 1.85, 1.6, 1.0, 0.8, and 0.7 mm after the calculation and CST-simulation.

The feed waveguide is fed at the center through a standard X-band waveguide. The width of the feed slot is optimized in CST to get the desired match at 9.375 GHz.

A 3-D exploded model of the antenna assembly is shown in Fig. 9.

Fig. 9. Exploded view of antenna assembly.
The surface currents on the plane containing the radiating slots are depicted in Fig. 10. It can be observed that the amount of current on the slots decreases with increasing distance from the center of the antenna, that exhibits amplitude tapering.

A simulated 3-D pattern of the antenna is depicted in Fig. 11. A comparison of the measured and simulated patterns is given in Fig. 12. The first sidelobe level of the antenna is 25.6 dB down and the 3-dB beamwidth of the antenna is 8.2 degrees. The measured gain of the antenna is 26 dBi.

A comparison of the measured and simulated reflection coefficients of the antenna is given in Fig. 13. The antenna shows a bandwidth of 200 MHz for the –10 dB input reflection coefficient. The radiation efficiency is calculated as,

\[
\text{Efficiency} = \frac{\text{Gain} \times \lambda^2}{\text{Area} \times 4\pi},
\]

and efficiency of the antenna is 46 %.

**IV. Conclusions**

A planar slotted waveguide antenna has been designed at 9.37 GHz using broadwall radiating slots. The antenna design is based on the commonly used Elliot’s slotted waveguide antenna design procedure, using the commercially available EM simulation tools, HFSS & CST MWS. The antenna has been designed for a –25 dB sidelobe level and a beamwidth of 8.2°. Comparisons of the measured and simulated far-field patterns and input reflection coefficients have been reported, and show a maximum 10 dB error within the measured-bandwidth due to loss or leakage in the waveguide adapter.
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