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

Rashid Ahmad Bhatti[1] · Byeong-Yong Park[1] · Yun-Taek Im[2] · Seong-Ook Park[1]

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 bysampling a continuous circular Taylor aperture distribution at the 25 dB sidelobe level in both the E and Hplanes. A commercially available electromagnetic (EM) simulation tool has been used to characterize the individual isolated slot and that data hassubsequently 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 onlinear 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 data 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 bothexternal 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 radiatorsin 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-

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[1]Department of Electrical Engineering, Korea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea.

[2]Department of Information & Communications Engineering, Korea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea.

Corresponding Author : Yun-Taek Im (e-mail : imyuntaek@kaist.ac.kr)

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 forthe 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_r) and susceptance (B/G_r), was recorded and was normalized with resonant conductanceof the slot. The design curve thus generated is shown in Fig. 1. The real and imaginary parts of the admittance are denoted as a $h_1(y)$ and $h_2(y)$, respectively, where y is the slot length normalized by the resonant length, l/l_r, which can be written as

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

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, $g(x) = G_r/G_0$, of the isolated shunt slot as a function of offset is plotted

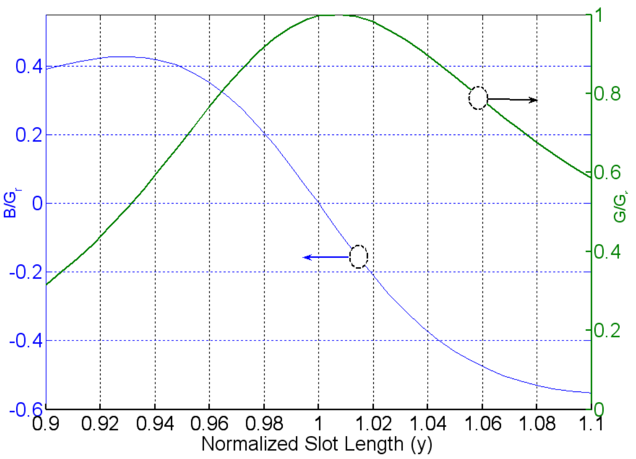

Fig. 1. Normalized admittance of isolated shunt slot versus slot length.

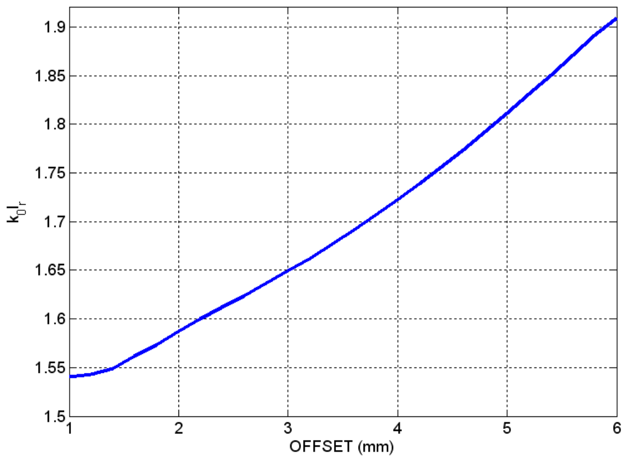

Fig. 2. Resonant length of the broadwall slot versus offset.

in Fig. 3. The curve $g(x)$ can be approximated as follows [7],

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

$$K = 2.09 \frac{a/b}{\beta_{10}/k_0} \cos^2\left(\frac{\pi \beta_{10}/k_0}{2}\right), \tag{1}$$

where a and b are the waveguide width and height respectively, and β_{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 thedistance of the half-guided wavelength from the center of the slot and resistance was recorded at the port located at thedistance of the half-guided wavelength from

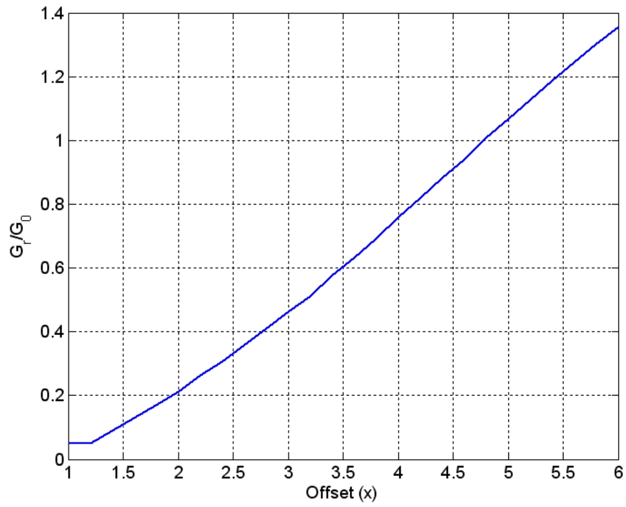

Fig. 3. Normalized conductance of isolated shunt slot as a function of offset.

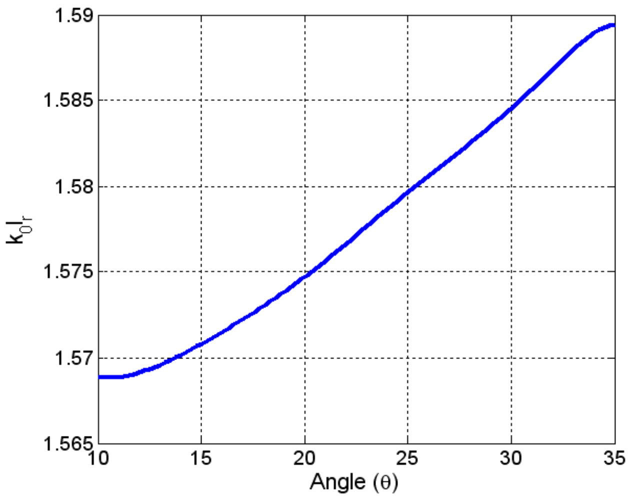

Fig. 4. Resonant length of the series coupling slot as a function of tilt angle.

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].

$$\frac{Y^a_{mn}}{G_0} = K_1 f_{mn} \frac{V^s_{mn}}{V_m} \tag{2}$$

where,

$$K_1 = \frac{1}{j(a/\lambda)} \sqrt{\frac{2(k/k_0)}{\eta G_0(\beta_{10}/k)(ka)(kb)}}$$

and,

$$f_{mn} = -K \frac{(\pi/2kl_{mn})\cos(\beta_{10} l_{mn})}{(\pi/2kl_{mn})^2 - (\beta_{10} k)^2} \sin\left(\frac{\pi x_{mn}}{a}\right).$$

In equation (2), V_m is the mode voltage in the m^{th} branch waveguide and V^s_{mn} is the slot voltage of the mn^{th} slot. Y^a_{mn} is the active admittance of the mn^{th} slot.

$$\frac{Y^a_{mn}}{G_0} = \frac{2 f^2_{mn}(x_{mn}, l_{mn})}{\dfrac{2 f^2_{mn}(x_{mn}, l_{mn})}{\dfrac{Y}{G_0}(x_{mn}, l_{mn})} + MC_{mn}} \tag{3}$$

$$MC_{mn} = K_3 \sum_{r=1}^{R} \sum_{\substack{s=1 \\ rs \neq mn}}^{S} \frac{V^s_{rs}}{V^s_{mn}} g_{mnrs}(x_{mn}, l_{mn}, x_{rs}, l_{rs})$$

$$K_3 = j(\beta_{10}/k)(k_0 b)(a/\lambda)^3$$

$$g_{mnrs} = \int_{-k_0 l_{mn}}^{k_0 l_{mn}} \cos\left(\frac{z'_{mn}}{4l_{mn}/\lambda_0}\right) \bullet \left\{ \frac{1}{4l_{mn}/\lambda_0} \left[\frac{e^{-jk_0 R_1}}{k_0 R_1} + \frac{e^{-jk_0 R2}}{k_0 R_2} \right] + \left[1 - \frac{1}{4l_{rs}/\lambda_0}\right] \bullet \int_{-k_0 l_{rs}}^{k_0 l_{rs}} \cos\left(\frac{z'_{rs}}{4l_{rs}/\lambda_0}\right) \frac{e^{-jk_0 R}}{k_0 R} dz'_{rs} \right\} dz'_{mn},$$

where R is the distance from the point $P_{mn}(0,0,\zeta'_{mn})$ measured in local coordinates at the center of the mn^{th} slot to the point $P_{rs}(0,0,\zeta'_{rs})$ measured in local coordinates at the center of the rs^{th} slot. l_{mn} and l_{rs} represent the slot length of $mn^{th}$and rs^{th} , respectively.

$$R = \sqrt{(x_{mn} - x_{pq})^2 + \left[(z_{mn} - z_{pq}) + \left(\zeta_{mn}/k_0 - \zeta_{pq}/k_0\right)\right]^2}$$

R_2 is the distance from P_{mn} to the left end of the slot $P_{rs}(0,0,-l_{rs})$ and R_1 is the distance from P_{mn} to the right end of the slot, $P_{rs}(0,0,l_{rs})$ and is given as,

$$R_1 = \sqrt{(x_{mn} - x_{pq})^2 + \left[(z_{mn} - z_{pq}) + \left(\zeta_{mn}/k_0 - l_{pq}\right)\right]^2}$$

$$R_2 = \sqrt{(x_{mn} - x_{pq})^2 + \left[(z_{mn} - z_{pq}) + \left(\zeta_{mn}/k_0 + l_{pq}\right)\right]^2}$$

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 witha feed waveguide placed

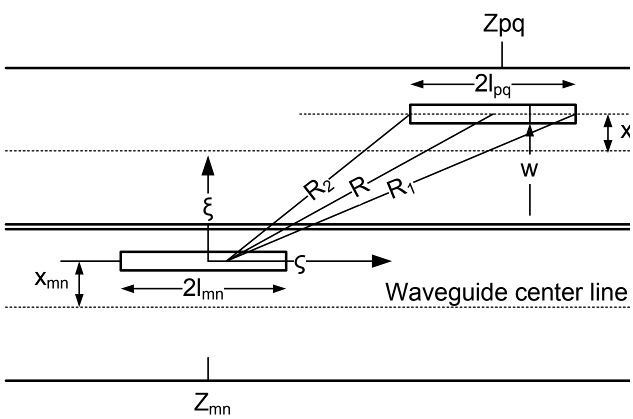

Fig. 5. Two slots in a planar array.

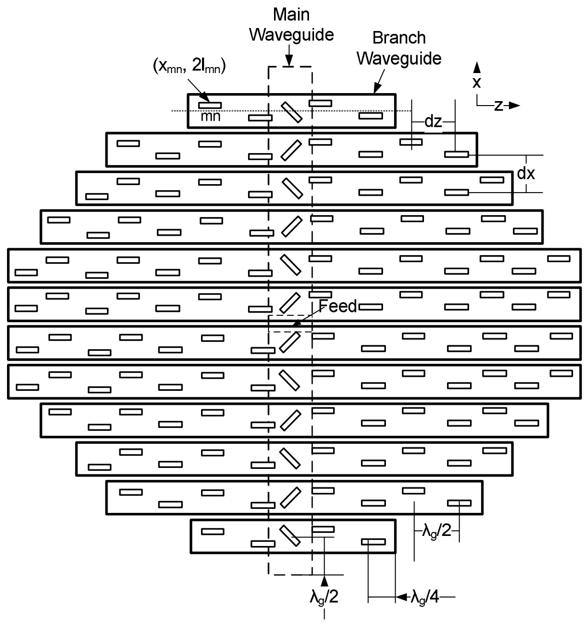

Fig. 6. Planar slotted waveguide array antenna.

behind.

For the mn^{th} slot in the array, the distance from the origin is calculated as [9],

$$\rho_{mn} = \sqrt{\left(\left(\frac{(2m-1)dz}{2}\right)^2 + \left(\frac{(2n-1)dx}{2}\right)^2\right)},$$

where dx and dz are the inter-element spacing along x and z directions.

The aperture distribution is then calculated as follows [9],

$$A_{mn} = g_0(p_{mn}), \tag{4}$$

where

$$p_{mn} = \frac{\pi \rho_{mn}}{a}$$

$$g(p) = \frac{2}{\pi^2} \sum_{0}^{\bar{n}-1} \frac{S_n J_0(\pi p \mu_n)}{J_0^2(\pi \mu_n)}$$

$$S_m = J_0(\pi \mu_m) \frac{\prod_{n=1}^{\bar{n}-1} \left(1 - \frac{\mu_m^2}{u_n^2}\right)}{\prod_{\substack{n=1 \\ m \neq n \\ m>0}}^{\bar{n}-1} \left(1 - \frac{\mu_m^2}{u_n^2}\right)}, \quad S_0 = 1,$$

where μ_m are zeros of $J_1(\pi u)$.

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.

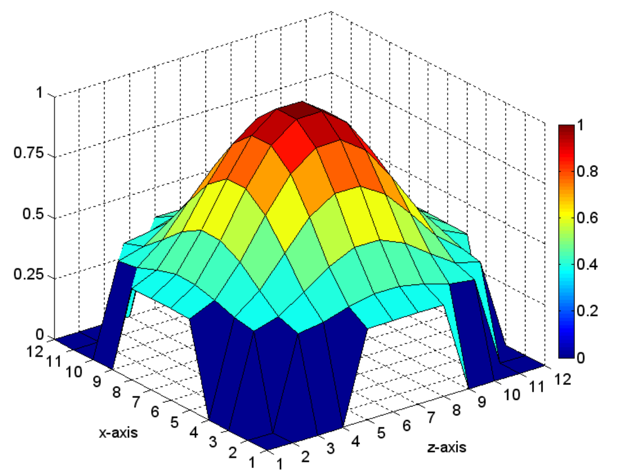

Fig. 7. Aperture distribution for the planar array with truncated corner elements.

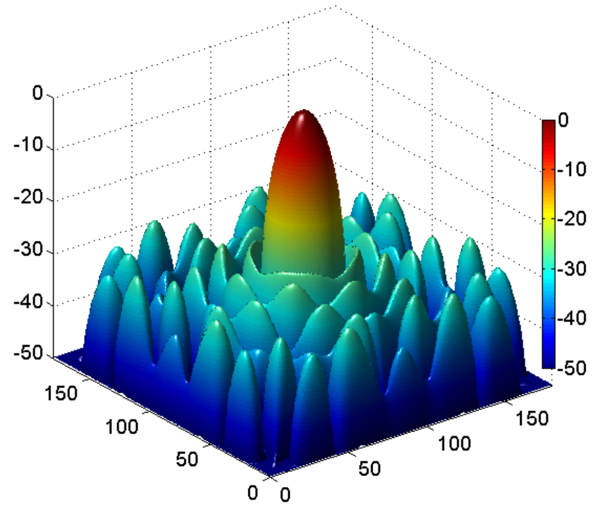

Fig. 8. Array factor corresponding to the calculated aperture distribution.

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 termin equation (3),

$$MC_{mn} = K_3 \sum_{r=1}^{R} \sum_{\substack{s=1 \\ rs \neq mn}}^{S} \frac{V_{rs}^s}{V_{mn}^s} g_{mnrs}(x_{mn}, l_{mn}, x_{rs}, l_{rs})$$

for every mn in the array.

Asearch is then conducted to find the couplet (x_{mn}, l_{mn}) that satisfies the following relation,

$$\mathrm{Im}\left[\frac{2f_{mn}^2(x_{mn},l_{mn})}{\frac{Y}{G_0}(x_{mn},l_{mn})}\right] = -\mathrm{Im}[MC_{mn}], \tag{5}$$

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

$$\frac{Y(x_{mn},l_{mn})}{G_0} = \left(h_1(y_{mn}) + h_2(y_{mn})\right)g(x_{mn}).$$

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,

$$\frac{\frac{Y_{pq}^a}{G_0}}{\frac{Y_{mn}^a}{G_0}} = \frac{f_{pq}^2(x_{pq},l_{pq})V_{pq}^s V_m}{f_{mn}^2(x_{mn},l_{mn})V_{mn}^s V_p}. \tag{6}$$

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,

$$\sum_{n=1}^{N(m)} \frac{Y_{mn}^a}{G_0} = C_m, \tag{7}$$

where C_m is the required admittance level in the m^{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,

$$\sum_{m=1}^{M} r_m = W, \tag{8}$$

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.

$$B^2(m) = \sum_{n=1}^{N_m} A^2(m,n), \tag{9}$$

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,

$$\frac{R}{R_0} = 0.131\frac{\beta_{10}}{k}\frac{\lambda_0^2}{ab}\left[I_\theta \sin\theta + \frac{\lambda_g}{2a}J_\theta\cos\theta\right], \tag{10}$$

where theta is the tilt angle of the series slot, and

$$\left.\begin{matrix} I_\theta \\ J_\theta \end{matrix}\right| = \frac{\cos\left(\frac{\pi}{2}K_1\right)}{1-K_1^2} \pm \frac{\cos\left(\frac{\pi}{2}K_2\right)}{1-K_2^2}$$

$$\left.\begin{matrix} K_1 \\ K_2 \end{matrix}\right| = \frac{\beta_{10}}{k}\cos\theta \pm \frac{\lambda_0}{2a}\sin\theta.$$

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.

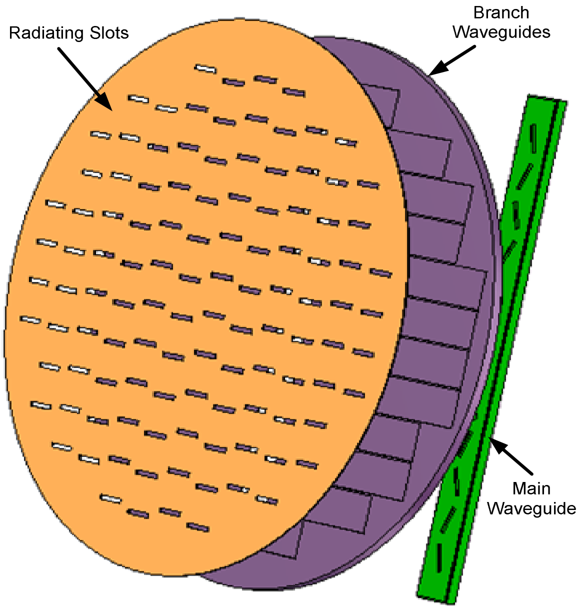

Fig. 9. Exploded view of antenna assembly.

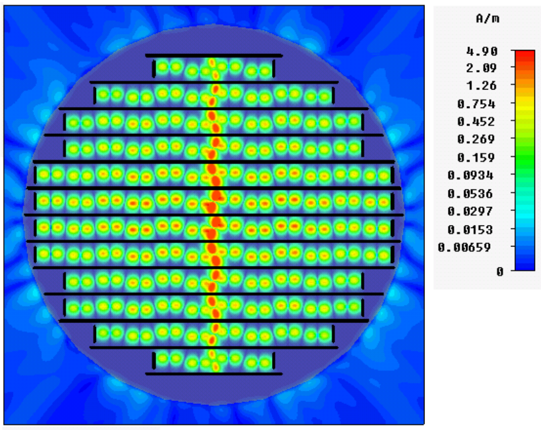

Fig. 10. Surface currents on the aperture.

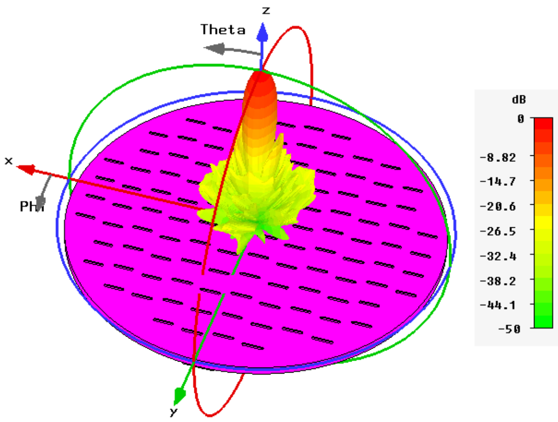

Fig. 11. Simulated 3-D pattern of the antenna.

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,

$$Efficiency = \frac{Gain \times \lambda^2}{Area \times 4\pi}, \tag{11}$$

and efficiency of the antenna is 46 %.

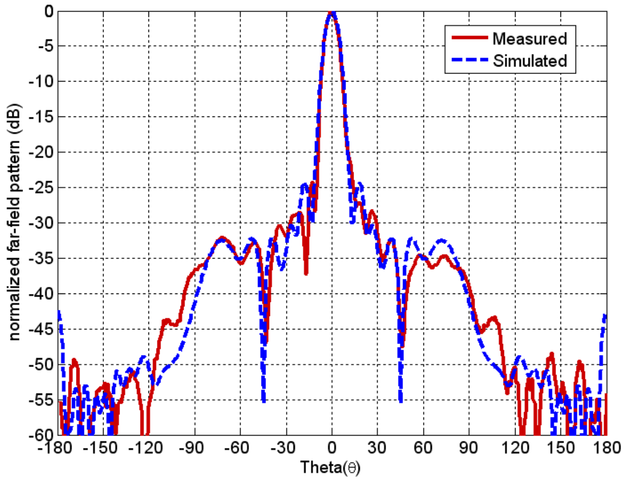

Fig. 12. Comparison of the measured and simulated E-plane patterns.

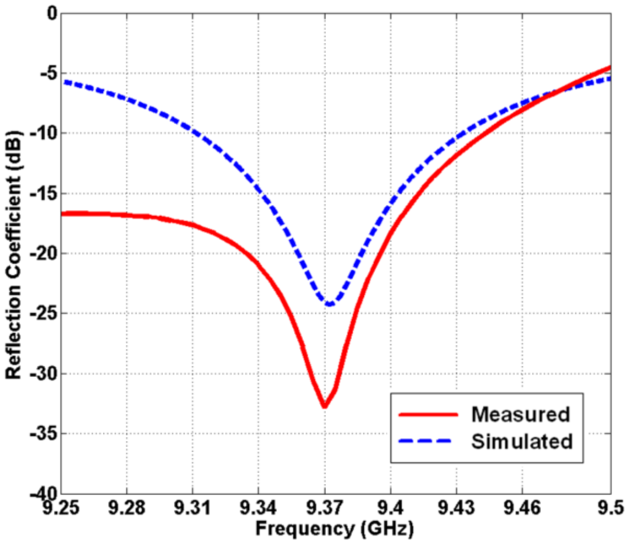

Fig. 13. Comparison of the measured and simulated input reflection coefficients.

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 amaximum 10 dB error within the measured-bandwidthdue to loss or leakage in thewaveguide adapter.

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Rashid Ahmad Bhatti

was born in Bahawal-Nagar, Pakistan, on July 14, 1975. He received the B.Sc. degree in electrical engineering from University of Engineering and Technology, Taxila, in 1999 and the M.Sc. degree in electrical engineering from National University of sciences and Technology (NUST), Rawalpindi, Pakistan, in August, 2005. Currentlyhe is working toward the Ph.D. degree in electrical engineering at the Korea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea. From September 1999 to December 2005, he worked as an RF and Antenna Research Engineer at Advanced Engineering Research Organization (AERO), Pakistan. Here, he has been involved in the EMI/EMC testing and design of slotted waveguide array antennas for trackingradars. His current research interests include the design of MIMO antennaarrays, multiband antennas for portable communication devices, slotted waveguideantennas and numerical methods in electromagnetics.

Byeong-Yong Park

was born in Daejeon, Korea. He received the B.S. degree in division of electrical engineering from ChungNam National University, Korea, in 2008, the M.S. degree in electrical engineering from Korea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea, in 2010, and is currently working toward the Ph.D. degree in electrical engineering atKorea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea. His current research interest is the design of MIMO antenna arrays for portable communication devices.

Yun-Taek Im

was born in ChungNam, Korea, on June 7th, 1978. He received the B.S. degreein division of electrical engineering and computer science from Hanyang University, Korea, in 2005, the M.S. degree in electronic and electrical engineering from the Pohang University Science and Technology at Pohang (POSTECH), in 2007, and is currently workingtoward the Ph.D. degree in electrical engineering at the Korea Advanced Institute of Science and Technology (KAIST), Daejeon, Korea. He was engaged in development of RFID antennas and millimeter-wave beam-forming systems from 2005 to 2007. His current research interests are MIMO antenna systems and radar systems including antenna design.

Seong-Ook Park

(M'05) was born in KyungPook, Korea, in December 1964. He received the B.S. degree from KyungPook National university, in 1987, the M.S. degree from Korea Advanced Institute of Science and Technology, Seoul, in 1989, and the Ph.D. degree from Arizona State University, Tempe, in 1997, all in electrical engineering. From March 1989 to August 1993, he was aResearch Engineer with Korea Telecom, Daejeon, working with microwave systems and networks. He later joined the Telecommunication Research Center, Arizona State University, until September 1997. Since October 1997, he has been with the Information and Communications University, Daejeon, first as an Associate Professor, and currently as a Professor at the Korea Advanced Institute of Science and Technology, IT convergence Campus. His research interests include mobile handset antenna, and analytical and numerical techniques in the area of electromagnetics. Dr. Park is a member of Phi Kappa Phi.