Analysis and Design of Coplanar Waveguide-Fed Slot Antenna Array

Tzung-Fang Huang, Shih-Wen Lu, and Powen Hsu, *Senior Member, IEEE*

***Abstract*— A method for analyzing and designing the slot antenna array excited by a coplanar waveguide (CPW) is presented. The slots are etched on the conducting plane of the CPW and placed in the direction perpendicular to the transmission line. Moment-method analysis and matrix-pencil approach are adopted to calculate the scattering parameters and hence the self-admittance of each slot. The mutual admittances between the slots are calculated from the formulas derived for the complementary strip dipoles based on the reciprocity theorem and via Booker's relation. Then the transmission line theory is used to calculate the input impedance of the array, and an iterative process is employed to obtain a matched design for a desired slot-voltage distribution. A four-element slot array is fabricated and measured using this design procedure. Calculated results are in good agreement with measurements.**

***Index Terms*—Coplanar waveguides, slot arrays.**

I. Introduction

DUE to its low radiation losses and easy insertion of the active and passive components, the coplanar waveguide (CPW) has been often used as an alternate to microstripline for feeding the printed antennas, especially when used in the millimeter wave applications. Many investigators have studied extensively many different kinds of CPW-fed printed antennas, such as the patch antennas [1]–[5], the slot antennas [6], [7], and the loop antennas [7]–[9]. All these studies, however, were concentrated on the isolated radiating elements only. As for the antenna array fed by the CPW, only experimental works [10], [11] have been published so far. Since an array is usually required in a specific design of a high-directive antenna, it deserves an analysis effort to facilitate the design. In this paper, we present a method for analyzing and designing the CPW-fed slot antenna array.

In array design, the mutual coupling between the antenna elements cannot be ignored simply, because it can result in mismatch of the individual elements to their feed and the distortion of radiation pattern of the array. In this analysis, a simple and efficient method for calculating the mutual admittances between the slots in the array is described. First, the slot array is transformed into its complementary structure of strip dipole array. Then, by neglecting the surface waves propagating along the dielectric substrate and the higher order modes of magnetic current flowing in the CPW, the mutual impedances between the strip dipoles are calculated using the formula derived from the reciprocity theorem [12]. Next, these dipole mutual impedances are transformed back to the desired slot mutual admittances via Booker's relation [12]. Although this method for calculating the mutual admittance is only approximate, it has the advantages of memory and central processing unit (CPU) time savings over the full wave calculation when a large array is considered.

The self admittance of each slot in the array is calculated using the matrix pencil approach [13], [14] in conjunction with the moment method described in [4], with a slight modification in the numerical analysis that will be described in the section that follows. Once the self and mutual admittances are obtained, the transmission line theory is then employed to connect the adjacent slots for calculating the input impedance of the array. For array pattern synthesis, or designing an array with a desired slot-voltage distribution, an iterative design procedure is introduced to provide suitable slot dimensions for the input match and the required slot-voltage distribution.

To check the accuracy of our analysis, a uniformly excited four-element slot array is designed, fabricated, and tested in the anechoic chamber using HP-8510B network analyzer with thru-reflect-line (TRL) calibration technique. The input impedances and the radiation patterns are measured and compared to the theoretical ones. Good agreement is observed between theory and experiment.

II. Analysis for Self and Mutual Admittances

The geometry of a CPW-fed slot antenna array is shown in Fig. 1(a). The center-fed radiating slots are etched on the conducting plane of the CPW and placed in the direction perpendicular to the transmission line. The transmission line is extended beyond the last slot (marked in reverse order as slot 1) by a short circuited section of length ll, and a quarter-wave transformer section d_{in} of CPW is attached in between the first slot and the connector (or other transmission line). These additional short circuited section and the quarter wave transformer are used to have a better and easier input match of the design. In the following analysis, we assume that the substrate and the conducting plane extend to infinity and that the conductor is of negligible thickness.

The self-admittance of each slot in the array is calculated by assuming that it is the same as the isolated slot admittance. For an isolated slot element as shown in Fig. 2(a), since it is excited by the magnetic currents in the slots along CPW, it can be considered as a series impedance presented to the

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The authors are with the Department of Electrical Engineering and Graduate Institute of Communication Engineering, National Taiwan University, Taipei, Taiwan, 10617 R.O.C.

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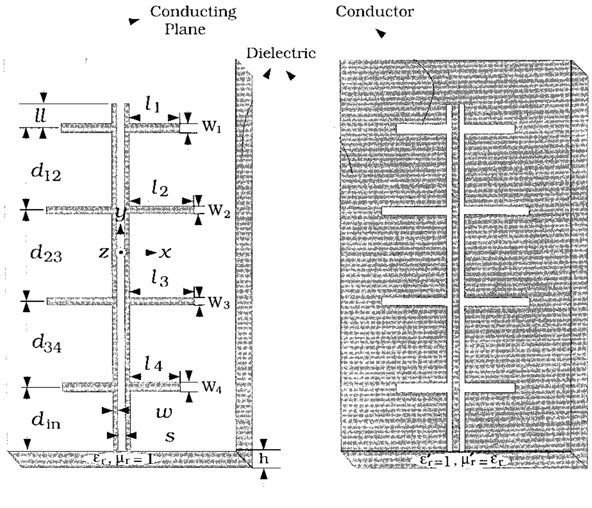
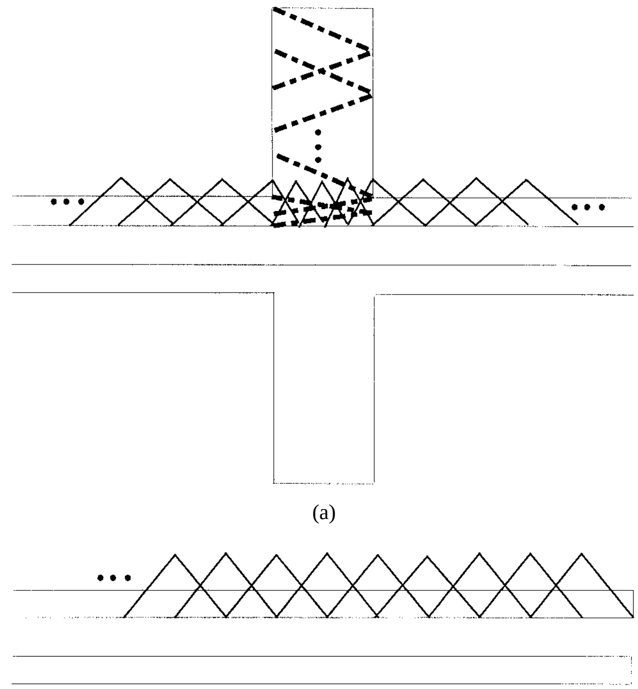


(a) (b)

Fig. 1. (a) Geometry of a CPW-fed slot antenna array. (b) The complementary structure of (a); a strip dipole array fed by coplanar strips.



(a)

(b)

Fig. 2. The rooftop basis functions of several cell divisions in the moment-method analysis. (a) Isolated slot. (b) Shorted end of the CPW line.

transmission line. From two-port network analysis, the self-admittance of the slot $Y_{\text{self}}^{\text{slot}}$ can be defined as

$$Y_{\text{self}}^{\text{slot}} = \frac{S_{12}}{2Z_{\text{CPW}} \cdot S_{11}} \tag{1}$$

where Z_{CPW} is the characteristic impedance of the CPW and S_{11} and S_{12} are scattering parameters which can be calculated from the matrix pencil approach [13], [14] if the magnetic current in the CPW is known. The magnetic current can be obtained by using a Galerkin moment method with rooftop basis functions similar to that described in [4] except there is no patch exists. The traveling wave component used in [4] is abandoned in our analysis and we expand the total magnetic current into a pure combination of rooftop basis functions. In order to have better results, the division regions which support the rooftop functions are used differently inside and outside the radiating slot. A smaller division is used inside and, as shown in Fig. 2(a), an asymmetric rooftop function is employed at the intersection where the CPW meets the radiating slot. Once the magnetic current is calculated, the matrix-pencil approach [13], [14] is then used to do the de-embedding procedure and extract the input and output components for the CPW dominant mode, from which the scattering parameters and, hence, the self admittance of the radiating slot can be solved. This method can also be applied to calculate the load impedance Z_L of the shorted section at the end of the transmission line as shown in Figs. 1(a) and 2(b).

When we deal with the mutual admittances between any two slots in the array, it is very tedious to calculate them by the full wave analysis as the one we just mentioned for the isolated slot, especially when the array is large. Certain approximations should be made to accelerate the computation. In this analysis, we assume that the mutual coupling effects are mainly caused by the radiation through free-space, the surface waves that propagate along the dielectric substrate can be neglected. This is acceptable as long as the dielectric constant is low and the substrate thickness is small compared with wavelength [15]–[17]. Moreover, we also assume that the mutual coupling does not change the field distribution of the slot [15]–[17] and that the higher order modes of the magnetic current scattered in the CPW can be neglected. The latter is true if the spacing between any two adjacent slots is large enough so that the excited higher order modes are fully decayed before reaching the neighboring slots. Under these assumptions, we can then calculate the mutual admittance of the slot from a simple formula derived using reciprocity theorem for the complementary strip-dipole array and via Booker's relation [12]. The method is briefly described below.

The complementary structure of the CPW-fed slot array, a strip-dipole array fed by coplanar strips, is shown in Fig. 1(b). In this figure, the original substrate of relative permittivity ϵ_r and relative permeability $\mu_r = 1$ is transformed into a substrate of relative permittivity $\epsilon_r' = 1$ and relative permeability $\mu_r' = \epsilon_r$. To simplify the problem, the substrate effects on the slots are replaced by a medium with relative permittivity ϵ_{eff} which surrounds the slots. This ϵ_{eff} is calculated by taking the geometrical mean of all the different effective permittivities on the slots. The effective permittivity on each slot can be calculated from the spectral domain analysis [18] by treating the slot as a slotline. Similarly, in complementary structure, we assume that the strip dipoles are surrounded by a medium with relative permeability $\mu_r' = \epsilon_{\text{eff}}$. To calculate the mutual impedance between the strip dipoles, we further transform the strip dipoles into cylindrical dipoles with radii [12]

$$a_i = \frac{1}{4}(W_i + t), \qquad i = 1, 2, \cdots \tag{2}$$

where W_i's are widths of the strips (or slots) and t is the thickness of the conductor. In this analysis, we assume $t = 0$. The mutual impedance Z_{ij}^{dipole} between cylindrical dipoles i

and j can be calculated from the following formula derived from the reciprocity theorem [12]

$$Z_{ij}^{\text{dipole}} = -\frac{1}{I_i(0)I_j(0)} \int_{-l_j}^{l_j} E(\zeta_j) I_j(\zeta_j) d\zeta_j \tag{3}$$

where E is the electric field set up by dipole i on dipole j and I_i and I_j are electric currents on dipoles i and j, respectively. E can be calculated from [12, eqs. (7.15), (7.16)] and I_i and I_j can be calculated from [12, eq. (7.103)]. After Z_{ij}^{dipole}'s are calculated, the mutual admittances of the original slots and the mutual impedances of the final cylindrical dipoles are connected by the Booker's relation [12], [19], [20]

$$[Y^{\text{slot}}] = \frac{4}{\eta_0^2} \cdot [Z^{\text{dipole}}] \tag{4}$$

where η_0 is the intrinsic impedance of free-space.

III. Design Procedure

Utilizing the above analysis for the self and mutual admittances, we have developed below a procedure for designing the CPW-fed slot array. For simplicity, we will describe our design procedure by using a uniformly excited four-element slot array. This, however, does not limit our method to the small arrays with uniform excitation. In fact, it can be extended to design large arrays with arbitrary slot-voltage distribution.

Consider a four-element slot array shown in Fig. 1(a). The slots are spaced on centers with equal spacing of one waveguide wavelength λ_{CPW} and series coupled to a CPW, which is supported by a thin substrate. The dielectric substrate used has relative permittivity $\epsilon_r = 2.6$ and thickness $h = 1.6$ mm. The design frequency is $f = 7.0$ GHz and the slots are to be fed with equal and in-phase voltages, i.e., $V_1 = V_2 = V_3 = V_4$. In order to have physically realizable dimensions for all the slots, we choose $s = 2$ mm and $w = 1$ mm, where s and w are strip and slot widths of the CPW, respectively. These dimensions will be kept constants for the whole CPW line and throughout the entire design. Once the CPW dimensions are decided, it can be calculated from the two-dimensional spectral-domain analysis [18] that the characteristic impedance Z_{CPW} and the waveguide wavelength λ_{CPW} of the CPW are 92.22 Ω and 32.91 mm, respectively, and from the matrix-penciled moment method described in the previous section that the shorted end impedance Z_L is $(0.33 + j8.07)$ Ω.

Suppose the antenna is to be connected to a 50-Ω transmission line or connector; then the section d_{in} of CPW before input to the antenna can be considered as a quarter-wave transformer between the antenna and the connector. With this consideration, the length of the quarter-wave section will be $d_{\text{in}} = \lambda_{\text{CPW}}/4 = 8.23$ mm, and the input impedance of the antenna measured at the center of the first slot (marked in reverse order as slot four) will be about 170 Ω. Since the slots are spaced equally with λ_{CPW} apart and are fed by equal and in-phase voltages, the active impedance of each of the four slots can then be assumed to be around 42.5 Ω, which corresponds to an isolated slot with length $l = 16.5$ mm and width $W = 4.0$ mm by the matrix-penciled moment-method calculation described in the previous section. With these initial parameters in hand, a table for the isolated slot admittance with variations in the slot length and width is generated using the matrix-penciled moment-method calculation to facilitate the iteration process in the design procedure to be described subsequently. In generating the table, the slot lengths used are from 10 to 20 mm with steps of 0.5 mm; the slot widths used are from 2 to 4 mm with steps of 0.2 mm. The slot width used cannot be too wide since we assume that the slots are series elements presented to the CPW line, whereas the slot length is chosen such that the slot is close to its second resonance. The isolated slot admittances in the table will be chosen as self-admittances Y_{ii} of the slots in the iteration process that follows. Since the admittance table is quite large, we will not show it here.

If (V_i, I_i) is the voltage and current pair at the terminal of the ith slot as shown in Fig. 1(a), then

$$I_i = \sum_{j=1}^{4} Y_{ij} V_j, \qquad i = 1, 2, 3, 4 \tag{5}$$

where Y_{ij} is the mutual admittance between slots i and j if $i \neq j$. Dividing this equation by V_i, we obtain the expression for the active admittance of each slot

$$Y_i^a = \sum_{j=1}^{4} Y_{ij} \cdot \frac{V_j}{V_i}, \qquad i = 1, 2, 3, 4. \tag{6}$$

For a given slot-voltage distribution, if Y_i^a could be determined beforehand, then a proper set of slot dimensions $(l_i, W_i, i = 1, 2, 3, 4)$ which gives rise to suitable values of Y_{ij} that satisfy (6) would give us what we are looking for in the design. To determine Y_i^a, we call upon the transmission line theory, which, together with the condition for quarter-wave transformer, yields

$$\frac{Z_{\text{CPW}}^2}{Z_{\text{CPW}} \cdot \frac{Z_L + jZ_{\text{CPW}} \tan(\beta_{\text{CPW}} \cdot ll)}{Z_{\text{CPW}} + jZ_L \tan(\beta_{\text{CPW}} \cdot ll)} + \sum_{i=1}^{4} \frac{1}{Y_i^a}} = Z_{\text{in}} \tag{7}$$

where β_{CPW} is the propagation constant of the CPW and Z_{in} is the impedance input to the quarter-wave transformer d_{in}. We will use $Z_{\text{in}} = 50$ Ω in the following design. Since the slots are equally spaced with λ_{CPW} apart and are series connected to the transmission line, the terminal currents are the same on every slot, therefore, $V_1 Y_1^a = V_2 Y_2^a = V_3 Y_3^a = V_4 Y_4^a$. Thus, the ratio between any two Y_i^a's can be calculated if the slot voltages are known. Substituting the ratios (in terms of one of Y_i^a's) into (7) and giving an initial value for the shorted section ll, then Y_i^a can be determined. Now, choosing the initial value as $ll = 0.07\lambda_{\text{CPW}} = 2.30$ mm, which is more than one half of the maximum value of width W used in the table generated above, the uniform excitation of the slots yields

$$Y_i^a = (21.3 + j6.95) \times 10^{-3}\ \mho, \qquad i = 1, 2, 3, 4. \tag{8}$$

Using (6)–(8), we describe below an iterative procedure which would provide us the desired slot dimensions for the design. To begin with, let us assume that initially $Y_{ij} = 0$ for $i \neq j$. Since $V_i = V_j$ for uniform excitation, therefore, from (6) and (8), $Y_{ii} = Y_i^a = (21.3 + j6.95) \times 10^{-3}\ \mho$. Searching for the closest value to this Y_{ii} in the table generated previously, we find

$$Y_{11} = Y_{22} = Y_{33} = Y_{44} = (21.7 + j6.52) \times 10^{-3}\ \mho$$

which corresponds to

$$l_1 = l_2 = l_3 = l_4 = 16 \text{ mm}$$
$$W_1 = W_2 = W_3 = W_4 = 3.6 \text{ mm}$$

in the table. Using these slot dimensions, we calculate the mutual admittances from the formulas described in the previous section and obtain

$$Y_{12} = Y_{23} = Y_{34} = (6.91 + j1.44) \times 10^{-3}\ \mho$$
$$Y_{13} = Y_{24} = (-0.96 + j4.08) \times 10^{-3}\ \mho$$
$$Y_{14} = (-2.89 - j0.56) \times 10^{-3}\ \mho.$$

Substituting these mutual admittances and the Y_i^a obtained in (8) into (6), we have a new set of self admittances

$$Y_{11} = Y_{44} = (18.2 + j1.99) \times 10^{-3}\ \mho$$
$$Y_{22} = Y_{33} = (8.44 - j0.016) \times 10^{-3}\ \mho.$$

Again, by searching for the closest values in the table for these Y_{ii}, the new self-admittances become

$$Y_{11} = Y_{44} = (21.1 + j4.52) \times 10^{-3}\ \mho$$
$$Y_{22} = Y_{33} = (10.7 - j3.96) \times 10^{-3}\ \mho$$

and their corresponding slot lengths and widths are

$$l_1 = l_4 = 16 \text{ mm}$$
$$l_2 = l_3 = 10.5 \text{ mm}$$
$$W_1 = W_4 = 4 \text{ mm}$$
$$W_2 = W_3 = 2.6 \text{ mm}.$$

Repeat the above procedure from calculating the mutual admittances to searching for the new slot dimensions again and again until we reach the match condition

$$\left|\frac{Z_{\text{in}} - 50}{Z_{\text{in}} + 50}\right| < 0.1 \tag{9}$$

then the design is completed. In (9), Z_{in} is calculated from (7) with Y_i^a calculated from (6) by using the Y_{ij} computed from the latest slot dimensions obtained in the design procedure. It should be noted that in calculating Y_i^a, the ratio V_j/V_i in (6) is obtained from (5) by setting $I_i = I_0 =$ constant for every i, which happens to be the case in this example. If after several iterations the match condition still cannot be reached, then we have to change the selection of the length of the shorted section ll, and sometimes, maybe even the characteristic impedance of the CPW should be changed by changing its dimensions s and w.

The final design of the above example is as follows:

$$ll = 0.18\lambda_{\text{CPW}} = 5.92 \text{ mm}$$
$$l_1 = l_2 = l_3 = l_4 = 13 \text{ mm}$$
$$W_1 = W_4 = 2.8 \text{ mm}$$
$$W_2 = W_3 = 4 \text{ mm}$$

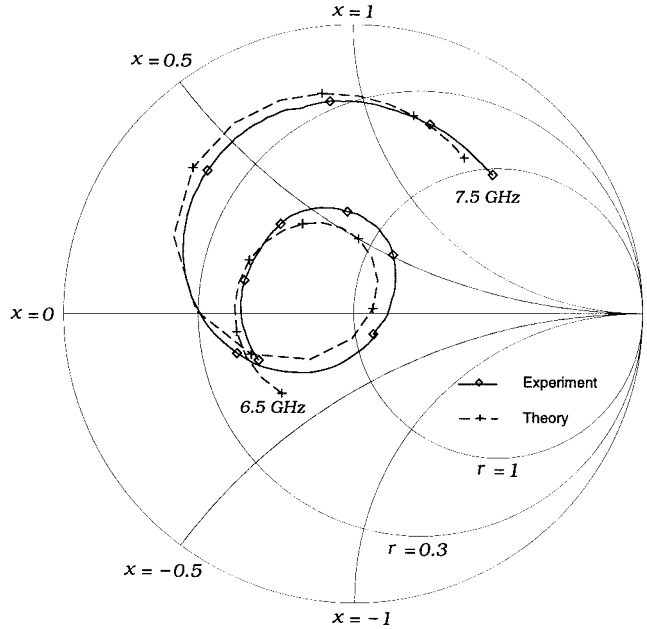


Fig. 3. Normalized input impedance of the CPW-fed slot antenna array. Frequency increases clockwise with steps of 0.1 GHz. $s = 2$ mm, $w = 1$ mm, $\epsilon_r = 2.6$, $h = 1.6$ mm, $ll = 5.92$ mm, $d_{12} = d_{23} = d_{34} = 32.91$ mm, $d_{\text{in}} = 8.23$ mm, $l_1 = l_2 = l_3 = l_4 = 13$ mm, $W_1 = W_4 = 2.8$ mm, and $W_2 = W_3 = 4$ mm.

$$Y_{11} = Y_{44} = (8.71 + j9.16) \times 10^{-3}\ \mho$$
$$Y_{22} = Y_{33} = (9.05 + j6.55) \times 10^{-3}\ \mho$$
$$Y_{12} = Y_{34} = (-0.37 + j3.29) \times 10^{-3}\ \mho$$
$$Y_{13} = Y_{24} = (-1.84 - j0.26) \times 10^{-3}\ \mho$$
$$Y_{14} = (0.18 - j1.30) \times 10^{-3}\ \mho$$
$$Y_{23} = (-0.19 + j3.20) \times 10^{-3}\ \mho$$
$$V_1 : V_2 : V_3 : V_4 = 1 : (0.93 - j0.13) : (0.93 - j0.13) : 1$$
$$Z_{\text{in}} = (57.2 + j1.89)\ \Omega.$$

It is seen that the voltage distribution is close to uniform and that Z_{in} satisfies the match condition.

IV. Experimental Verification and Discussion

To verify the validity of our design and analysis, a test piece was carefully fabricated using above parameters and measured by the HP8510B network analyzer with TRL calibration technigue. The measured and calculated normalized input impedances (input to the quarter-wave transformer d_{in}) are shown in Fig. 3. Their corresponding reflection coefficients R are plotted in Fig. 4. It can be seen from these two figures that the agreement between theory and experiment is excellent. In Fig. 4, the resonant frequencies defined at minimum reflection for theory and experiment are 7.012 GHz and 6.995 GHz, respectively. The discrepancy between theory and experiment may be attributed to a number of sources such as the finite size of the substrate, the dimensional tolerances exist in the test piece, the assumption of uniform transverse field distribution in the coplanar waveguide and in the slot disregarding the singular edge behavior, and the simplified model for calculating the mutual admittances. Nevertheless, the agreement between theory and experiment proves that our analysis is good for

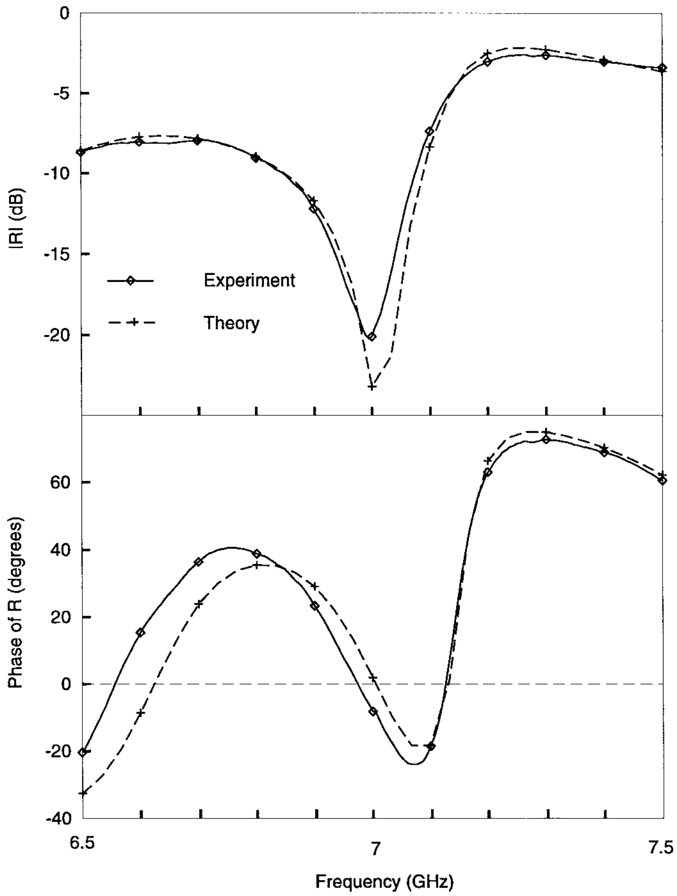


Fig. 4. Reflection coefficients corresponding to results shown in Fig. 3.

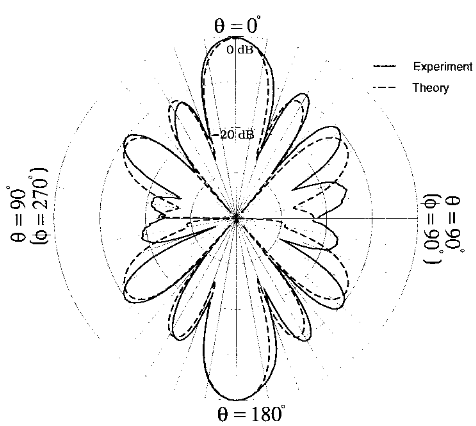


Fig. 5. E-plane radiation pattern of the slot array at $f = 7$ GHz.

the design purpose. Although theoretical calculations may be improved by extending the matrix-penciled moment-method analysis for the isolated slot to the entire array, i.e., the full-wave analysis of the array, our simplified analysis is much faster and more realistic if the array is large (ten or more elements) and serves as an adequate tool for the design.

The far-field radiation patterns of both E- and H-planes are measured and compared to the theoretical values in Figs. 5 and 6, respectively. The measurements are performed in the anechoic chamber and the calculations are made using only the isolated slot magnetic currents. The latter is reasonable

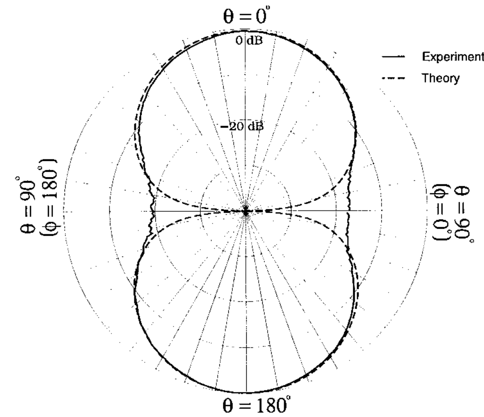


Fig. 6. H-plane radiation pattern of the slot array at $f = 7$ GHz.

since we have assumed that the mutual coupling does not effect the slot currents. Also, since the two equivalent magnetic currents flowing in the slots along CPW will radiate almost out of phase, we assume they contribute negligibly to the cross-polarized radiation components. In Fig. 5, the E-plane patterns exhibit 3-dB beamwidths of about 15.7° and 17.6°, respectively, for theory and experiment in $z > 0$ plane. In $z < 0$ plane, they are about 16.7° and 18.2°, respectively. In Fig. 6, the 3-dB beamwidths of the H-plane patterns are quite large. Their calculated and measured values are 56.5° and 52°, respectively, in $z > 0$ plane, while in $z < 0$ plane, they are 60.5° and 57.7°, respectively. The difference in the radiation power between $z > 0$ half-space (free-space) and $z < 0$ half-space (with thin dielectric substrate) is very small in this design. The calculated and measured front-to-back peak power ratios are 0.38 dB and 0.43 dB, respectively. All these data reveal that our assumptions are adequate.

V. Conclusion

A method for analyzing and designing the CPW-fed slot antenna array has been presented. The self-admittance of the slot is calculated using the matrix-penciled moment-method with rooftop basis functions, while the mutual admittance between the slots is calculated approximately from the formulas derived for the complementary strip dipoles using reciprocity theorem and via Booker's relation. Combining these self and mutual admittances, an efficient interative procedure based on the transmission-line theory is developed for designing arrays with input match and arbitrary slot-voltage distribution. A uniformly excited four-element slot array is designed and fabricated using this design method. Comparisons between calculated and measured input impedances and radiation patterns show good agreement between them. This design method is fast and accurate and is particularly useful as compared to the full wave analysis when a large array is to be designed.

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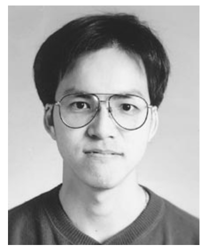
Tzung-Fang Huang was born in Chiayi, Taiwan, R.O.C., in 1968. He received the B.S., M.S., and Ph.D. degrees in electrical engineering from National Taiwan University, Taipei, Taiwan, in 1991, 1993, and 1999, respectively.

In June 1999, he joined the Acer NeWeb Corporation, Hsinchu, Taiwan, working on antenna design required for wireless communications. His interests and specialties include the design of slot and microstrip antennas and computational methods in electromagnetics.



Shih-Wen Lu was born in Taipei, Taiwan, R.O.C., in 1970. He received the B.S.E.E. degree from National Taiwan University, Taipei, Taiwan, in 1992. He is currently working toward the Ph.D. degree at the Department of Electrical Engineering, National Taiwan University.

His current topics of interest are the design and analysis of slot antennas and microstrip antennas.

Powen Hsu (M'86–SM'98) was born in Taipei, Taiwan, R.O.C., in 1950. He received the B.S. degree in physics from National Tsing-Hua University, Hsinchu, Taiwan, in 1972, the M.S. degree in physics from the University of Maryland, College Park, in 1976, and the M.S. and Ph.D. degrees in electrical engineering from the University of Southern California, Los Angeles, in 1978 and 1982, respectively.

From 1982 to 1984, he was with ITT Gilfillan, Van Nuys, CA, where he engaged in research and development pertaining to radar antenna systems. In 1984 he joined the faculty of National Taiwan University, Taipei, Taiwan, where he is currently a Professor in the Electrical Engineering Department. He was the Department Chairperson there from 1992 to 1995. Since August 1997, he has been the Dean of the newly established College of Electrical Engineering at National Taiwan University. His current research interests include the design and analysis of waveguide slot arrays, microstrip antennas, and microwave and millimeter-wave integrated circuits.