Analysis of Aperture-coupled Microstrip Antenna Using Method of Moments

Tomáš FARKAŠ, Peter HAJACH
Dept. of Radio-electronic
Slovak University of Technology
Ilkovičova 3, 812 19 Bratislava
Slovak Republic

Abstract

A microstrip patch antenna that is coupled to a microstripline by an aperture in the intervening ground plane is analyzed by using the method of moments. Integral equation is formulated by considering the exact dyadic Green's function in spectral domain for grounded dielectric slab so that the analysis includes all coupling effects and the radiation and surface wave effects of both substrates. The combination of the reciprocity method analysis and a Galerkin moment method solution seems to be suitable for a number of planar antenna problems, especially when coupling slots in the ground plane are included. Results for antenna input impedance are compared with other authors and verified by experimental results.

Keywords

Microstrip antenna, method of moments, reciprocity theorem

1. Introduction

Microstrip antennas are commonly fed by one of three methods: coaxial probe, microstripline connected directly to the edge of a patch, and stripline coupled to the patch through an aperture in the ground plane [1]. The latter method was first proposed by Pozar in 1985 and it becomes very popular in recent years as a method for feeding planar phased arrays in millimeter wave band. Some of its advantages are as follows: a) the feed network is isolated from the radiating element by ground plane that prevents spurious radiation, b) active devices can be fabricated in a feed substrate with high dielectric constant for size reduction, c) the radiating element can be placed on a thick substrate with low dielectric constant in order to achieve higher bandwidth.

The method of analysis of aperture coupled microstrip patches presented here is basically a full-wave method, which uses exact Green's functions in spectral domain in order to find necessary field components from electric and magnetic currents in the presence of a grounded dielectric slab. Reciprocity theorem [2] is used to derive expressions for the amplitudes of reflected and transmitted waves on the microstrip feeding line, and equivalent circuit representing the slot discontinuity is found.

The combination of the reciprocity analysis and a moment method solution is very versatile technique that should find application in a number of planar antenna problems. This method avoids the more "brute-force" approach of modeling the nonuniform currents on a feed line, as was done in [3].

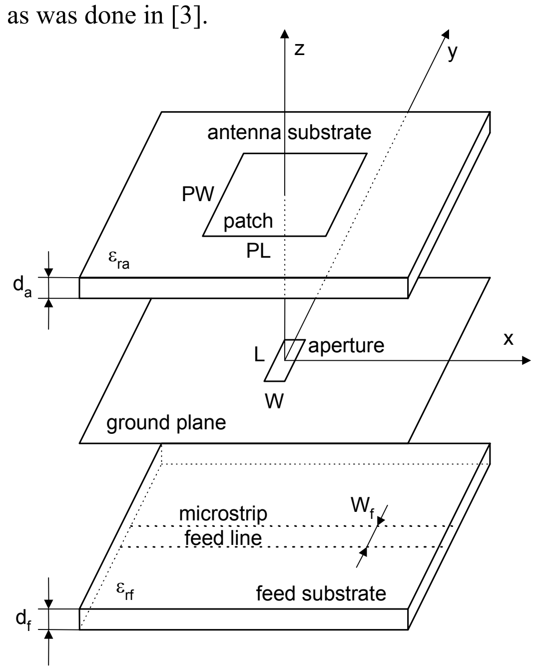

Fig. 1 Geometry of an aperture coupled microstrip patch antenna

2. Theory

The geometry of the aperture coupled microstrip patch antenna is shown in Fig. 1. Assuming a microstrip patch antenna with element size *PL*, *PW* placed on the surface of the antenna substrate and a coupling aperture with size *L, W* in the ground plane located below the centre of the patch. Feed line of width W_f is placed on the surface of the feed substrate on the opposite side of the ground plane, along the axis of the patch.

The metal surfaces are assumed to be made of a perfect conductor. The dielectric losses can be readily accounted by introducing complex permittivity

$$\tilde{\varepsilon}_r = \varepsilon_r (1 - j \tan \delta) .$$

2.1 Reciprocity analysis

The feed microstripline is assumed to be infinitely long and propagating a quasi TEM mode with fields [2]

$$\vec{E}^{\pm} = \vec{e}(y,z)e^{\pm j\beta x}, \tag{1a}$$

$$\vec{H}^{\pm} = \pm\vec{h}(y,z)e^{\pm j\beta x}, \tag{1b}$$

where $\vec{e}$, $\vec{h}$ are normalized transverse modal fields and β is the propagation constant of the line. If the aperture is centred at $x = 0$, total microstrip line fields can be written as

$$\vec{E} = \begin{cases} \vec{E}^{+} + R\vec{E}^{-}, & x < 0 \\ T\vec{E}^{+}, & x > 0 \end{cases} \tag{2}$$

$$\vec{H} = \begin{cases} \vec{H}^{+} + R\vec{H}^{-}, & x < 0 \\ T\vec{H}^{+}, & x > 0 \end{cases} \tag{3}$$

where R and T are the voltage reflection and transmission coefficients on the line, respectively. Unknown aperture electric field is taken as a piecewise sinusoidal (PWS) mode of the form

$$V_0 e_x^a(x,y) = V_0 \frac{\sin k_e\left(L/2 - |y|\right)}{W \sin k_e L/2}, \tag{4}$$

for $|x| < W/2$, $|y| < L/2$. In (4) V_0 is the unknown amplitude of the aperture field, and $k_e = k_0\,[(\varepsilon_r + 1)/2]^{1/2}$ is the effective wavenumber in the substrate. The total power flow through the aperture can be evaluated as

$$\Delta v = \int_{S_a} e_x^a(x,y) h_y(x,y) ds \tag{5}$$

where S_a is the aperture surface. Applying the reciprocity theorem the reflection and transmission coefficients can be found

$$R = 0.5 V_0 \,\Delta v, \; T = 1 - R. \tag{6}$$

Now define a Green's function G_{yy}^{HM} to account for the H_y fields on both sides of the aperture ($z = 0$) due to a y directed magnetic aperture current. An external admittance of the slot seen by the feed line can be written as

$$Y^e = \iint_{S_a S_a} e_x^a(x,y) G_{yy}^{HM}(x,y;x_0,y_0) e_x^a(x_0,y_0) ds\, ds_0. \tag{7}$$

Expressions for the unknowns V_0 and R can be written as

$$V_0 = \frac{2\Delta v}{\Delta v^2 + 2Y^e}, \; R = \frac{\Delta v^2}{\Delta v^2 + 2Y^e} \tag{8), (9}$$

The slot discontinuity appears as a simple series impedance Z to the microstripline, that can be found as

$$Z = Z_c \frac{2R}{1-R} = Z_c \frac{\Delta v^2}{Y^e} \tag{10}$$

where Z_c is characteristic impedance of the microstripline.

2.2 Moment method formulation

Following the standard moment method procedure, the aperture field and surface patch currents are expanded in a set of N PWS modes and matrix equations are solved for unknown complex expansion coefficients [4].

Considering only equivalent admittance of the aperture seen by the microstripline and enforcing continuity of H_y through the aperture one can write the following matrix equation [2]

$$\mathbf{Y}^e \mathbf{V} = (1 - R)\Delta\mathbf{v}, \tag{11}$$

where $\mathbf{V}$ is column vector of aperture field expansion coefficients. From (11) R and consequently Z can be found using (10). Components of matrix $\mathbf{Y}^e$ and vector $\Delta\mathbf{v}$ can be computed in spectral domain as

$$Y_{mn}^e = \frac{1}{4\pi^2} \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \tilde{f}_u^2(k_x, W) \tilde{G}_{yy}^{HM}(k_x, k_y) \tilde{f}_p^2(k_y, L) \times \cos k_y (y_m - y_n) dk_x dk_y, \tag{12}$$

$$\Delta v_n = \frac{1}{2\pi\sqrt{Z_c}} \int_{-\infty}^{\infty} \tilde{f}_u(k_y, W_f) \tilde{G}_{yx}^{HJ}(k_x = -\beta_f, k_y) \tilde{f}_p(k_y, L) \times \cos k_y y_n dk_y, \tag{13}$$

where k_x, k_y are spectral domain variables, y_n is the center point of the nth expansion mode, $f_u^{\sim}$ and $f_p^{\sim}$ are Fourier transforms of the expansion mode (see Appendix) and $G^{\sim}(k_x, k_y)$ are various components of spectral Green's functions which can be found for example in [2]. The wavy line in $G^{\sim}(k_x, k_y)$, $f_u^{\sim}$ and $f_p^{\sim}$ indicates, that these functions are defined in spectral domain variables as opposed to $G(\overline{r}, \overline{r}_0)$, f_u and f_p defined in space variables.

Next, the patch current is expanded in a series with a column vector of current expansion coefficients $\mathbf{I}$. The patch contribution to the aperture admittance seen by the slot is then

$$\mathbf{Y}^a = \mathbf{V}^{\mathrm{T}} \mathbf{Z}^{-1} \mathbf{V}, \tag{14}$$

where components of moment method impedance matrix $\mathbf{Z}$ and voltage vector $\mathbf{V}$ are given in spectral domain as

$$Z_{mn} = \frac{-1}{4\pi^2} \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \tilde{f}_p^2(k_x, PL) \tilde{G}_{xx}^{EJ}(k_x, k_y) \tilde{f}_u^2(k_y, PW) \times e^{jk_x(x_m - x_n)} dk_x dk_y, \tag{15}$$

$$V_n = \frac{1}{4\pi^2} \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \tilde{f}_p(k_x, PL) \tilde{f}_p(k_y, L) \tilde{G}_{xy}^{EM}(k_x, k_y) \times \tilde{f}_u(k_y, PW) \tilde{f}_u(k_x, W) e^{jk_x x_n} dk_x dk_y, \tag{16}$$

x_n is the center point of the nth expansion mode of the current on the patch, $f_u^{\sim}$ and $f_p^{\sim}$ are Fourier transforms of the current expansion mode (electric current on the patch and magnetic current in the aperture). The equivalent series impedance seen by the microstrip feed line is

$$Z = Z_c \frac{\Delta v^2}{Y^e + Y^a}. \tag{17}$$

The aperture coupled patch antenna is usually tuned with an open-circuited stub of microstrip line, approximately $\lambda_g/4$ long. If stub length is L_s, input impedance seen looking into the microstrip feed line referenced to aperture is

$$Z_{in} = Z - jZ_c \cot \beta L_s. \tag{18}$$

More accurate results can be obtained by adding a length extension to L_s to account for fringing fields at the end of the open stub, the length extension is approximately 0.4 d_f, where d_f is the feed substrate thickness.

When numerically evaluating the infinite integrals in k_x, k_y plane, in (12), (15), (16), it is convenient to transform the rectangular spectral variables k_x, k_y to their polar form α, β and special care must be taken to perform the quadrature integration in the vicinity of surface wave poles, as described for example in [1].

Various components of the spectral Green's functions $\tilde{G}(k_x, k_y)$ needed in computation, can be found for example in [2]. In the case, that the feed and antenna substrates have different parameters, by evaluating Green's function in computation of Δv and Y^e, parameters ε_{rf} and $\tan\delta_f$ should be used and in computation of Y^a parameters ε_{ra} and $\tan\delta_a$ should be used.

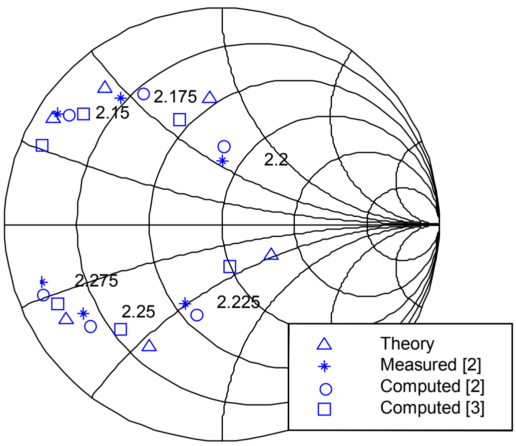

Fig. 2 Smith chart plot of the input impedance of a stub-tuned aperture coupled patch antenna versus frequency in GHz. ε_{ra} = 2.54, d_a = 0.16 cm, PL = 4 cm, PW = 3 cm, ε_{rf} = 2.54, d_f = 0.16 cm, L = 1.12 cm, W = 0.155 cm, W_f = 0.442 cm, L_s = 2 cm.

3. Results

We have implemented the above mentioned moment method in MATLAB and numerically computed input impedance of two configurations of aperture-coupled stub-tuned microstrip antennas and compared our results with experimental and theoretical data published elsewhere.

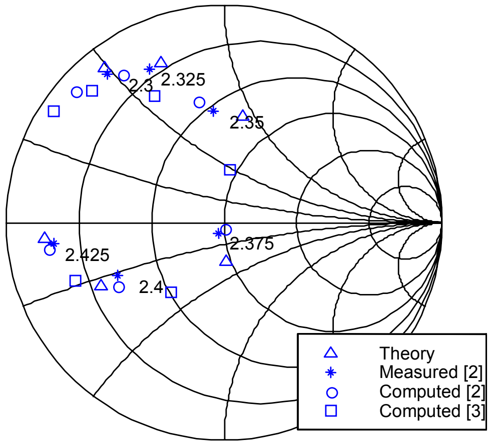

Fig. 3 Smith chart plot of the input impedance of a stub-tuned aperture coupled patch antenna versus frequency in GHz. ε_{ra} = 2.22, d_a = 0.16 cm, PL = 4 cm, PW = 3 cm, ε_{rf} = 10.2, d_f = 0.127 cm, L = 1 cm, W = 0.11 cm, W_f = 0.116 cm, L_s = 1.1 cm.

The case in Fig. 2 has the same substrate parameters for both the feed and antenna substrates. Our theory shows a small shift in the resonant frequency compared to experiment, which may be caused by neglecting the effect of nonuniform current distribution along the feed microstrip-line on the low dielectric constant substrate and by neglecting the actual dielectric losses in the substrate.

Fig. 3 shows results for a low dielectric constant (ε_{ra}=2.22) substrate for the antenna and a high dielectric constant (ε_{rf}=10.2) substrate for the feed line. This configuration simulates the monolithic phased array application, where the feed substrate would be Gallium Arsenide for phase shifters and other active circuitry. Results fit considerably well with both experimental and other authors data.

4. Conclusion

Calculation of input impedance of aperture coupled microstrip antenna using moment method in spectral domain have been presented and compared with experimental and theoretical results published elsewhere. The method is based on the reciprocity theorem and uses exact Green's functions for dielectric slab. The agreement between measured and computed results supports validity of method.

5. Appendix

1D Fourier transforms of following expansion modes are required:

a) pulse function

$$f_u(x,W) = \begin{cases} 1/W, & |x| < W/2 \\ 0, & |x| > W/2 \end{cases}, \tag{19}$$

b) piecewise sinusoidal (PWS) mode

$$f_p(x,L) = \begin{cases} \dfrac{\sin k_e(L/2 - |x|)}{\sin k_e L/2}, & |x| < L/2 \\ 0, & |x| > L/2 \end{cases} \tag{20}$$

The Fourier transform is defined as

$$\tilde{f}(k_x, W) = \int_{-\infty}^{\infty} f(x,W) e^{jk_x x} dx\,, \tag{21}$$

and the transforms of the above functions are

$$\tilde{f}_u(k_x, W) = \frac{\sin k_x W/2}{k_x W/2}\,, \tag{22}$$

$$\tilde{f}_p(k_x, L) = \frac{2k_e[\cos k_x L/2 - \cos k_e L/2]}{(k_e^2 - k_x^2)\sin k_e L/2}\,. \tag{23}$$

References

[1] KAY FONG LEE, WEI CHEN Advances in microstrip and printed antennas. New York: John Wiley & Sons, 1997.

[2] POZAR, D. M. A reciprocity method of analysis for printed slot and slot-coupled microstrip antennas. IEEE Transactions on Antennas and Propagation. 1986, vol. 34, no. 12, p. 1439 – 1446.

[3] SULLIVAN, P. L., SCHAUBERT, D. H. Analysis of an aperture coupled microstrip antenna. IEEE Transactions on Antennas and Propagation. 1986, vol. 34, no. 8, p. 977 – 984.

[4] HARRINGTON, R. F. Field computation by moment methods, IEEE, New York, 1993.

About authors...

Tomáš FARKAŠ (Ing.) was born in Bratislava, Slovakia, in 1971. He received the Ing (MSc) degree in radio-electronics from the Slovak Technical University of Bratislava, in 1995. Currently, he is with the Department of Radio-electronic of the Slovak Technical University in Bratislava. His research interests include microstrip antennas and numerical methods for electromagnetic field modeling.

Peter HAJACH (doc., Ing., Ph.D.) was born in Bratislava, Slovak Republic, on June 1946. He received the Ing. Degree in electrical engineering from the Slovak Technical University of Bratislava, in 1969, and the CSc. (Ph.D.) degree in radio-electronics from the STU Bratislava, in 1984. He is currently the Associate professor at the Radio-electronic department of the STU Bratislava. His research interests include microwave antennas.

RADIOENGINEERING REVIEWERS

December 2001, Volume 10, Number 4