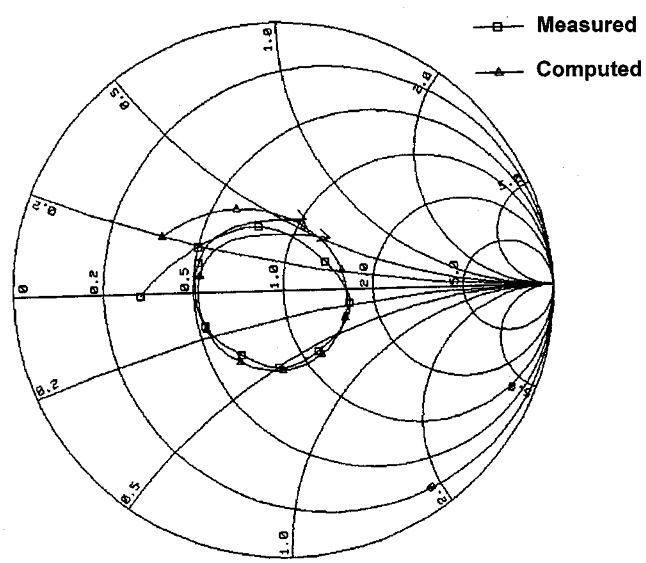

Fig. 2. Comparison of measured and computed input impedance of aperture-coupled antenna of three-coupled strips ($d_{\rm air} = 12.56$ mm, $d_{\rm ant} = 1.575$ mm, $\varepsilon_{\rm ant} = 2.2$, $d_{\rm feed} = 1.575$ mm, $\varepsilon_{\rm feed} = 2.2$, $L = 70.0$ mm, $W_1 = 18.0$ mm, $W_2 = 24.0$ mm, $S = 2.0$ mm, $L_a = 54.0$ mm, $W_a = 1.5$ mm, $W_{\rm feed} = 5.0$ mm, $L_{\rm stub} = 14.0$ mm). Frequency range: $f_{\rm start} = 1.35$ GHz, $f_{\rm stop} = 1.75$ GHz, and $f_{\rm step} = 40$ MHz.

even mode with the voltages at the strips of the type (+V, +V, +V), and that at 1.744 GHz is the odd mode with (−V, +V, −V) type of voltage distribution.

The $\rm VSWR = 2$ bandwidth for the two-strip and three-strip cases is almost the same. However, it is easier to control the impedance behavior for the three-strips antenna. Because of symmetrical distribution for the radiating even mode, there is no tilt in the H-plane radiation pattern. Even though the improvement in bandwidth over the rectangular patch is small (∼2 MHz), it is found to increase with the increase in substrate thickness. However, it is difficult to achieve purely real input impedance at resonance if probe fed. This is because of increased probe reactance with increased substrate thickness. Due to this constraint, we decided to study aperture-coupled microstrip antenna, which has the potential of achieving 50 Ω match with increased bandwidth.

III. Design of Aperture-Coupled Coupled-Strips Antenna

Using aperture coupling of the feed microstrip line to the antenna, wide ranges of resistance and reactance values can be achieved by adjusting coupling, i.e., slot size and stub length [3]–[5]. Coupling can also be varied by adjusting the position of the strip resonators with respect to the center of the slot. This is one of the important parameters for the design of coupled-strips geometry. Also, one has to adjust the substrate parameters and the feed parameters, like aperture length and width, stub length, and feed width. For three symmetric coupled-strips geometry with composite dielectric of air and $\varepsilon_r = 2.2$, we optimized the strip widths W and gap widths S, to get the maximum impedance bandwidth. The optimized parameters are given in Fig. 2. It may be noted that the centers of the outer strips (of width W_2) are located at $y = \pm 23$ mm and the length of the coupling slot $L_a = 54$ mm. Therefore, the outer strips are directly coupled to the slot in addition to the parasitic coupling to the central strip. The measured and calculated input impedance for this structure is shown in Fig. 2. The $\rm VSWR = 2$ impedance bandwidth of this antenna is 23% while that of a patch of the same size is only 13%. The measured bandwidth is exactly the same as the predicted one using spectral domain analysis. It was found that increasing the number of coupled strips did not give rise to a further increase in the bandwidth of antenna. The polarization purity of the coupled-strips antenna may be better as reported in [3], since the excited transverse current component is smaller on the narrow strips.

IV. Conclusion

In this letter, we have investigated the coupled-strips antenna. It is found that the enhanced bandwidth of this antenna is due to the even- and odd-mode resonances of the coupled-strips geometry. The even mode is responsible for radiated power and the odd mode increases the bandwidth by shaping the impedance curve. The frequency difference between the two resonant modes will ultimately limit the bandwidth of the antenna. The optimized design for three-coupled strips is presented. More than three coupled strips may not improve the bandwidth but may help in increased polarization purity. A stacked configuration can also be analyzed in terms of even and odd modes.

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A Broad-Band Rectangular Patch Antenna With a Pair of Wide Slits

Kin-Lu Wong and Wen-Hsiu Hsu

***Abstract*—A new broad-band design of a probe-fed rectangular patch antenna with a pair of wide slits is proposed and experimentally studied. The proposed design is with an air substrate, and experimental results show that, simply by inserting a pair of wide slits at one of the radiating edges of the rectangular patch, good impedance matching over a wide bandwidth can easily be achieved for the proposed antenna. With an air substrate of thickness about 8% of the wavelength of the center operating frequency, the proposed antenna can have an impedance bandwidth of about 24%. For frequencies within the impedance bandwidth, good radiation characteristics are also observed, with a peak antenna gain of about 7.2 dBi.**

***Index Terms*—Broad-band patch antenna, microstrip antennas.**

I. Introduction

It has been known that, when a thick air or foam substrate is used, bandwidth enhancement for a microstrip patch antenna can be

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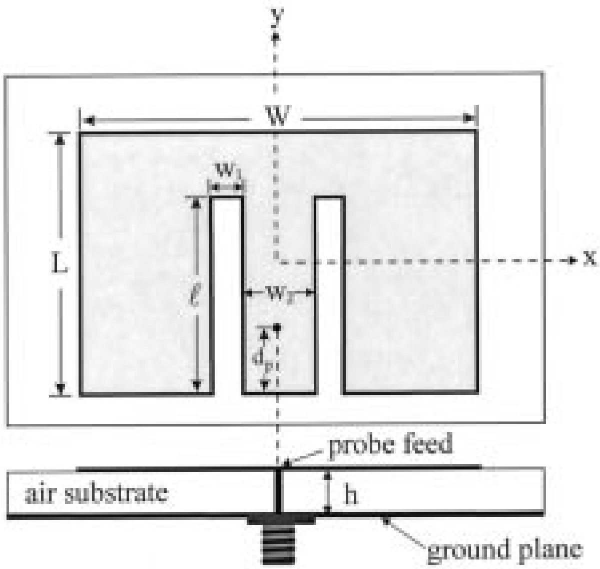


Fig. 1. Geometry of the proposed broad-band rectangular patch antenna with wide slits.

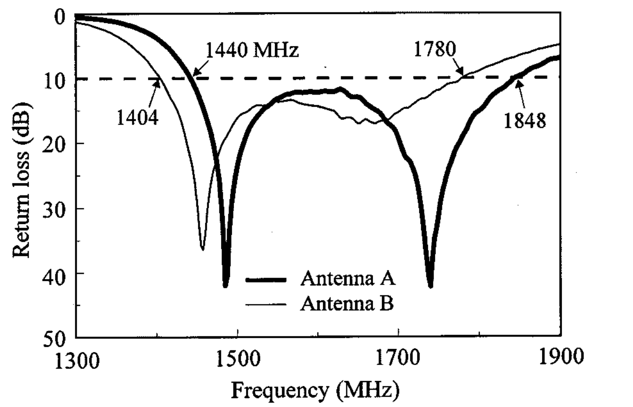
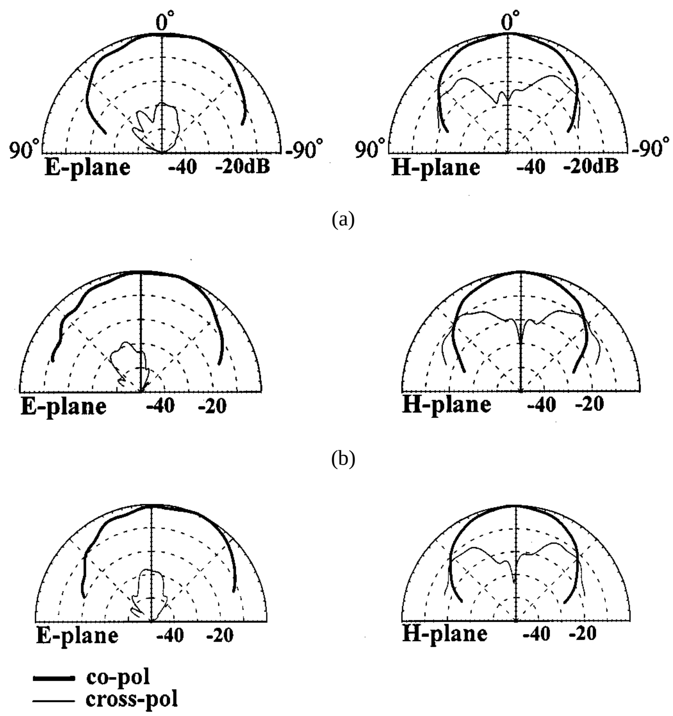


Fig. 2. Measured return loss for the proposed broad-band patch antenna; $L = 65$ mm, $W = 105$ mm, ground-plane size $= 150$ mm $\times$ 150 mm. Antenna A: $h = 14.3$ mm, $\ell = 47$ mm, $w_1 = 6.3$ mm, $w_2 = 15.3$ mm, and $d_p = 10$ mm. Antenna B: $h = 15.7$ mm, $\ell = 53$ mm, $w_1 = 10$ mm, $w_2 = 8$ mm, and $d_p = 13$ mm.



(c)

Fig. 3. Measured E-plane (y–z plane) and H-plane (x–z plane) radiation patterns for antenna A. (a) $f = 1485$ MHz. (b) $f = 1644$ MHz. (c) $f = 1740$ MHz.

obtained. However, when such a method is applied to a probe-fed microstrip patch antenna, it usually has a limited achievable bandwidth less than 10%, which is largely due to the increased inductance associated with the long probe pin in the thick substrate layer. Recently, it has been shown that, by embedding a U-shaped slot in the rectangular patch, an impedance bandwidth greater than 20% can easily be achieved for a microstrip antenna with a probe feed [1], [2]. In this letter, we demonstrate a new and simpler broad-band design of a rectangular patch antenna with a pair of wide slits (see Fig. 1). The wide slits are inserted at one of the radiating edges of the rectangular patch. The proposed design has a wide impedance bandwidth and good radiation characteristics similar to those reported for a U-slotted patch antenna. Also, the proposed design has a simpler structure than the design with a U-slot [1]–[3]. Details of the proposed antenna and experimental results are presented. Some simulation results obtained by IE3D™ are also given.

II. Antenna Design

The proposed design is shown in Fig. 1. The rectangular patch has dimensions of $L \times W$ and is supported by nonconducting posts of height h (not shown in the figure) from the ground plane. The two wide slits are of same length ℓ and same width w_1 and are inserted at the bottom edge of the patch. The separation of the two wide slits is w_2, and the two slits are placed symmetrically with respect to the patch's center line (y axis). Thus, there are only three parameters (ℓ, w_1, w_2) for the wide slits used here. Along the patch's center line, a probe feed at a distance d_p from the patch's bottom edge can be located for good excitation of the proposed antenna over a wide bandwidth.

III. Results and Conclusions

Fig. 2 shows the measured return loss for prototypes of antennas A and B in which different air–substrate thicknesses are selected. First note that, for achieving good impedance matching over a wide bandwidth, the slit length (ℓ) is found to be about 0.7 to $0.85L$, and the spacing between outer edges of the two slits $(2w_1 + w_2)$ is about 0.27 W. From the results, it is seen that two adjacent resonant modes are excited, which leads to a wide bandwidth. This characteristic is also similar to that obtained for the U-slotted patch antenna [1], [3]. For antenna A, the impedance bandwidth, determined from a 10-dB return loss, is 408 MHz or about 24.8% with respect to the center frequency at 1644 MHz, average of the measured lower and higher frequencies with a 10-dB return loss. For antenna B, the obtained bandwidth is 376 MHz or about 23.6% referenced to the center frequency at 1592 MHz.

Radiation characteristics of the antenna are also studied. Fig. 3 plots three different operating frequencies for antenna A, which are all of same polarization planes and similar radiation patterns. The cross-polarization radiation in the E-plane patterns is also seen to be less than -20 dB. The H-plane patterns, however, show relatively larger cross-polarization radiation. This behavior is also similar to that reported for

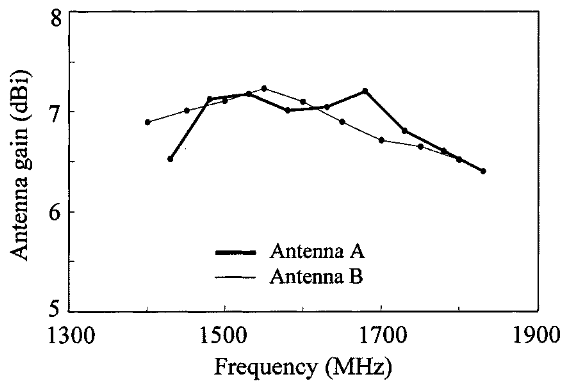


Fig. 4. Measured antenna gain in the broadside direction for antennas A and B.

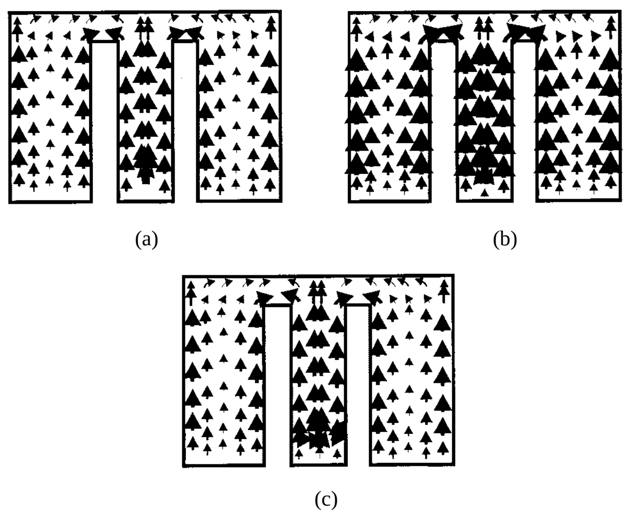


Fig. 5. Simulated current distributions on the rectangular patch of antenna A. (a) $f = 1485$ MHz. (b) $f = 1644$ MHz. (c) $f = 1740$ MHz.

the U-slotted patch [1], [3], and is largely due to the thick substrate thickness (about 0.08 times the wavelength of the center frequency in this study) and long probe pin in the substrate layer. Measured radiation patterns for antenna B (not shown here for brevity) also show similar results. Measured antenna gain for frequencies within the impedance bandwidth is also presented in Fig. 4. For both antennas A and B, the antenna gain variation is seen to be less than 0.8 dBi, with the peak antenna gain at about 7.2 dBi.

Finally, the excited patch's surface current distribution of the proposed antenna is also studied by using IE3D™. Fig. 5 shows the simulated patch's surface current distribution for antenna A. Results for three typical frequencies at 1485, 1644, and 1740 MHz are shown. It is seen that all the three frequencies have similar surface current distributions on the rectangular patch. This characteristic agrees with the results that similar radiation patterns for the three frequencies are measured (see Fig. 3).

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Stability Characteristics of Absorbing Boundary Conditions in Microwave Circuit Simulations

Wenhua Yu, Raj Mittra, and Sourav Chakravarty

***Abstract*—In this paper, we examine the stability properties of several absorbing boundary conditions in the finite-difference time-domain (FDTD) simulations of microwave circuits. The numerical experiments show that the stability characteristics of absorbing boundary conditions, e.g., MURss and perfectly matched layers (PML), can depend upon the discretization of the computational domain.**

***Index Terms*—Absorbing boundary conditions, finite-difference time-domain (FDTD), stability property.**

I. Introduction

Finite-difference time-domain (FDTD) methods are frequently used to simulate various electromagnetic problems involving microwave circuits [1], [2]. Absorbing boundary conditions, e.g., Mur's [3] and perfectly matched layer (PML) [4], play a very important role in FDTD simulations dealing with open region problems. In most microstrip circuit applications, the thickness of the substrate layer is relatively small compared to its length and often relative to its width as well. Even so, for accurate simulation of microstrip circuits, it is common to employ a fine discretization for the substrate layer. However, to reduce both the memory requirement and run time of the computer, we should use as few cells in the vertical direction as possible. In this communication, we examine the issues pertaining to the spatial discretizations in the horizontal and vertical directions, to see how they affect the stability properties of two absorbing boundary conditions widely used in FDTD simulations, *viz.*, Mur's and PML.

We show that the FDTD result can become unstable when the dimension of the computational domain in the vertical direction (z) is not sufficiently large compared to spatial discretization in the horizontal directions (x and y) and when the free-space region in the vertical direction, i.e., the space between the surface of the substrate and the absorbing boundary, is not sufficiently big. In the next section, we investigate the stability characteristics of absorbing boundary conditions in FDTD simulations of microwave circuits with a view to developing some guidelines that help assure the stability of the algorithm.

II. Numerical Analysis

We begin with a microwave stripline structure, shown in Fig. 1. In all of the numerical experiments in this communication, the time step is assumed to be consistent with the Courant Stability criterion and is given by

$$\Delta t = \frac{0.995}{c\sqrt{\left(\frac{1}{\Delta x}\right)^2 + \left(\frac{1}{\Delta y}\right)^2 + \left(\frac{1}{\Delta z}\right)^2}}$$

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