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IMPEDANCE CHARACTERISTICS OF A SLOT ANTENNA FED BY A PARALLEL-PLATE WAVEGUIDE

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ABSTRACT: *A theoretical model to analyze the impedance and radiation characteristics of a slot fed by a parallel-plate waveguide is presented. The slot is excited by a TEM mode propagating between the parallel plates and its effect is simulated by a series impedance in a transmission line. Its value is computed with the use of the Lorentz reciprocity theorem, whereas the radiation pattern is determined from the radiation of an equivalent magnetic current embedded in an infinite conductive surface. Experimental results on a prototype are given to validate the theoretical model. The data fit the predicted results well. © 1997 John Wiley & Sons, Inc.* Microwave Opt Technol Lett *14, 126–128, 1997.*

Key words: *Slot antenna; parallel-plate waveguide*

I. INTRODUCTION

Knowing the impedance and radiation characteristics of a slot embedded in a parallel plate is essential to the design of a parallel-plate slot array. Such an array is attractive because of its relatively high efficiency compared to similarly sized microstrip arrays, which suffer from high losses in the feeding structure [1], especially at high frequencies (millimeter waves). In addition, it has a low profile, is lightweight, and has a low-cost implementation. The advantage of this array over the comparable rectangular waveguide slot array [2], is its superior efficiency because of lower losses in the absence of separating conductive walls in the feeding structure and its low cost. The computation of the equivalent impedance of the slot in a parallel-plate waveguide is the cornerstone in the design of a standing- or traveling-wave array. We present a simple model based on the *Lorentz reciprocity theorem* [3] to calculate the equivalent series impedance of the radiating slot (close to its resonance), excited by a TEM mode propagating between the parallel plates. The radiation pattern is calculated from the electric field radiated by an equivalent magnetic current embedded in a conductive plate [4].

II. THEORY

The basic geometry of the slot embedded in the parallel-plate waveguide and its equivalent circuit are shown in Figure 1. A TEM mode propagating in the $+z$ direction excites electric field in the slot, $\underline{E}_s$. A_{TEM}, B_{TEM}, and C_{TEM} are the incident, reflected, and transmitted waves by the slot, respectively. The medium between the plates has a dielectric constant ε_r; the distance between the top and bottom plates is a, and their width is d, which is characteristic of the spacing between columns in the slot array. The slot length is $2l$ and its width is $2w$ $(2l \gg 2w)$. Because of the asymmetric nature of the scattered field from the slot, $B_{\text{TEM}} = -C_{\text{TEM}}$. Such behavior is typical for an equivalent series impedance on a transmission line. The equivalent normalized series impedance $\overline{Z}_s$ can be expressed in terms of the scattered and incident waves by [5]

$$\overline{Z}_s = \frac{2B_{\text{TEM}}/A_{\text{TEM}}}{1 - B_{\text{TEM}}/A_{\text{TEM}}}. \tag{1}$$

The relation between the backscattered wave B_{TEM} and the electric field distribution in the slot can be established with the use of the Lorentz reciprocity theorem [3] in the form

$$\int_s (\underline{E}_1 \times \underline{H}_2 - \underline{E}_2 \times \underline{H}_1) \cdot \underline{dS} = 0, \tag{2}$$

where $(\underline{E}_1, \underline{H}_1)$ is the scattered field due to the interaction of the slot and the incident TEM mode and $(\underline{E}_2, \underline{H}_2)$ is a normalized TEM mode propagating along the parallel-plate waveguide. Computation of the backscattered mode B_{TEM} through (2), gives the relationship [4]

$$B_{\text{TEM}} = -C_{\text{TEM}} = \frac{\int_{\text{slot}} (\underline{E}_1 \times \underline{H}_2) \cdot \underline{1}_y \, dS}{2S_{\text{TEM}}}, \tag{3}$$

in which S_{TEM} is the vector *Poynting* energy of the TEM mode and $\underline{E}_1$ is the electric field in the slot. We assume that the field in the slot is ξ directed $(\underline{1}_\xi = \underline{1}_z)$ and around resonance its distribution is cosinusoidal:

$$\underline{E}_s = \frac{V_s}{2w} \frac{\sin k_s(1 - |\zeta|)}{\sin k_s l} \underline{1}_\xi, \tag{4}$$

with k_s approximated by $k_0 \sqrt{(1 + \varepsilon_r)/2}$ [6], k_0 being the propagation constant in free space and V_s being the voltage in the slot. Substitution of (4) into (3) and evaluation of the integral yields

$$B_{\text{TEM}} = V_s \frac{\tan(k_s l/2) \sin c(kw)}{k_s ad}, \tag{5}$$

in which $k = k_0\sqrt{\varepsilon_r}$ and $\text{sinc}(x) = \sin(x)/x$. Attention will now be restricted to the special case of a resonant slot. A resonant slot is defined by the condition that the forward scattered wave C_{TEM} and the incident wave A_{TEM} are out of phase. This assumption permits a deduction of V_s via power balance $P_{\text{inc}} = P_{\text{refl}} + P_{\text{tr}} + P_{\text{rad}}$ with P_{inc}, P_{refl}, P_{tr}, and P_{rad} being the incident, reflected, transmitted, and radiated power by the slot. The incident power is given by $\frac{1}{2}A_{\text{TEM}}^2 S_{\text{TEM}}$ (no loss in generality results if we assume that A_{TEM} is real) and in like manner we can deduct the reflected and transmitted power by $\frac{1}{2}B_{\text{TEM}}B_{\text{TEM}}^* S_{\text{TEM}}$ and $\frac{1}{2}(A_{\text{TEM}} + C_{\text{TEM}})(A_{\text{TEM}} + C_{\text{TEM}})^* S_{\text{TEM}}$, respectively. By *Schelkunoff's equivalence principle* [3] the radiating slot is equivalent to a magnetic strip

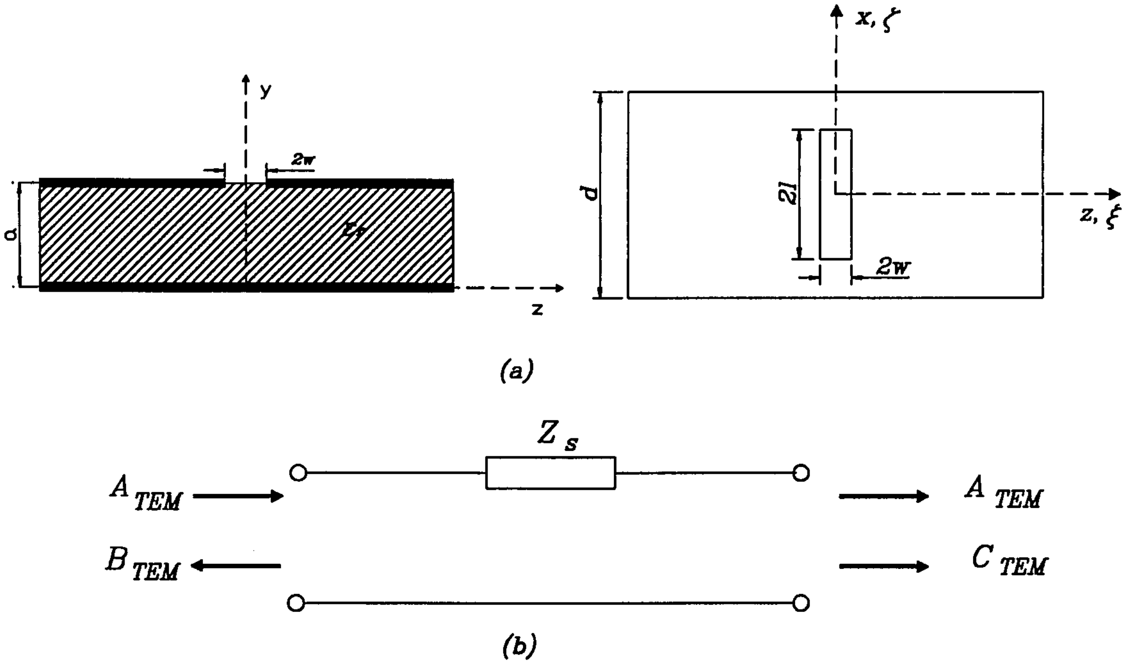

Figure 1 (a) Transverse slot embedded in a parallel-plate waveguide. (b) Equivalent circuit of the slot

dipole radiating into a half space and possessing a current distribution $J_m = 2\mu_0^{-1}E_s 1_\zeta$. If we compute the radiated far-field expression through the electric vector potential [3], the total radiated power by the slot can be computed. Such a computation results into the expression,

$$P_{\text{rad}} = \frac{V_s^2}{2\pi\eta \sin^2 k_s l}\left(\frac{k_0}{k_s}\right)^2 \times \int_0^\pi \left(\frac{\cos k_s l - \cos(k_0 l \cos\theta)}{1 - \left(\frac{k_0}{k_s}\right)^2 \cos^2\theta}\right)^2 \sin^3\theta \, d\theta, \quad (6)$$

in which $\eta = 120\pi$. Substitution of (5) and (6) into the power balance expression determines an explicit expression for the reflection coefficient from the slot, $B_{\text{TEM}}/A_{\text{TEM}}$, which can be substituted into (1) to yield the equivalent normalized series resistance of the slot, $\overline{R}_s$

$$\overline{R}_s = \frac{4\pi \sin^2 k_s l}{ad} \times \frac{\tan^2(k_s l/2)\cdot \sin c^2(kw)}{k_0^2 \cdot \int_0^\pi \left(\frac{\cos k_s l - \cos(k_0 l\cos\theta)}{1 - \left(\frac{k_0}{k_s}\right)^2 \cos^2\theta}\right)^2 \sin^3\theta\, d\theta}. \quad (7)$$

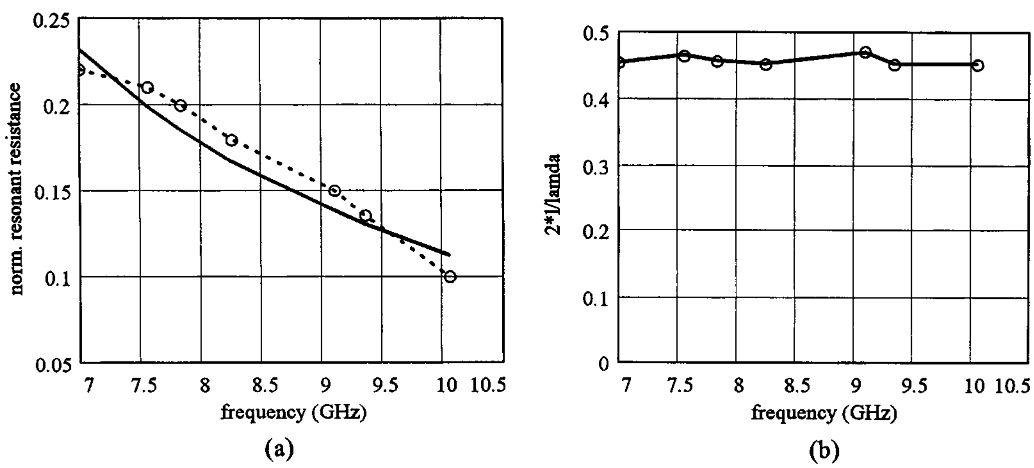

Figure 2 (a) Calculated and measured normalized resonant resistance of the slot versus frequency. (b) Normalized resonant length of the slot versus frequency. Solid lines, calculated. Dashed lines, measured

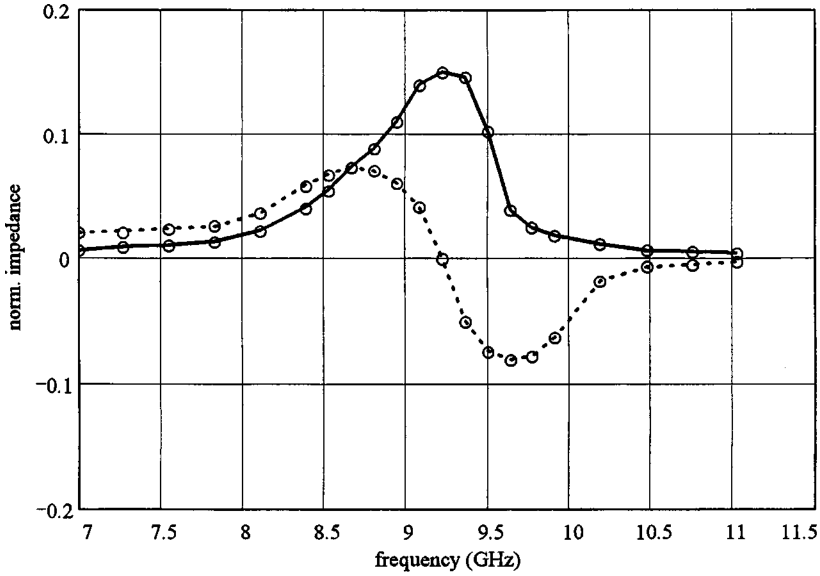

Figure 3 Real and imaginary components of the slot impedance versus frequency. Solid line, normalized Re{Z}. Dashed line, normalized Im{Z}

III. NUMERICAL AND EXPERIMENTAL RESULTS

To validate the theoretical results developed in Section II, an experimental module was built. The experiment was performed in X band. The module consists of two rectangular WR-90 waveguide ports with two sectional TEM launching horns at the transmitting and receiving ends and a central section of parallel plate waveguide with an embedded slot. The aperture of the launching horns was 20 × 1 cm, and the parallel-plates section was 20 cm long, 20 cm wide, and 1 cm high. The slot width was 1.5 mm and had a variable length. The impedance of the slot was measured in two steps. In the first step the slot was covered with conducting tape and the transmission coefficient T_s of the slot module was measured. Then, the transmission coefficient T with the slot uncovered was measured. One can show [5] that the complex transmission coefficients T_s and T are related to the normalized impedance of the slot $\overline{Z}_s$ by the expression $\overline{Z}_s = 2(T_s/T - 1)$. Figure 2 shows the normalized resonant length of the slot and a comparison between experimental and computed results of the equivalent normalized resonant slot resistance $\overline{R}_s$. One can observe that the slot resonant length is approximately $0.46\lambda_0$, and we obtained good agreement between the measured and computed normalized slot resistance at resonance. Figure 3 shows the variation with frequency for both the normalized resistance and reactance of the slot. The slot length was 1.49 cm. One can observe the bell shape (resistance) and S shape (reactance) of the slot impedance, which are typical to a resonant device.

IV. CONCLUSIONS

A simple theoretical model of the equivalent series impedance of the slot close to its resonance was developed. A test module was built and the impedance characteristics of the radiating slot in the parallel waveguide was measured. The experimental data fit the computed data well. The theoretical model of the slot can be used to design traveling and standing-wave slot arrays embedded in a parallel-plate waveguide.

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AN ACCURATE DETERMINATION OF THE Q FACTOR OF A DIELECTRIC RESONATOR IN A SUSPENDED SUBSTRATE ENVIRONMENT

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ABSTRACT: *An analytical technique for the determination of the Q factor of a dielectric resonator placed in a suspended substrate environment is presented in this article. Here, loss due to the conductor and*