Spurious Radiation from a Microstrip Y Junction

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Abstract—The radiation from a microstrip Y junction is calculated as a function of angle. It is shown that the radiated power increases monotonically with the Y angle, decreases inversely as the dielectric constant of the substrate for large permittivities, and increases as the square of the substrate thickness for thin substrates.

I. Introduction

A METHOD of calculating the radiation from discontinuities in microstrip has been applied to a number of special cases, including the open and short circuit, right-angle bend, matched section, and resonator [1]. Watkins [2] applied the same method to resonant-line resonators and showed that the mutual interaction between the ends could not be ignored, and that they led to a decreased Q factor for half-wave resonators and to a considerably increased Q factor for full-wave resonators. The reason for the latter, rather unexpected result is that parts of the microstripline are net absorbers of energy [3], reducing the total spurious radiation as a consequence.

The purpose of the present investigation is to explore the properties of the microstrip Y junction and to discover whether there is an optimum angle at which the spurious radiation could be substantially reduced. Although no such optimum apparently exists, the findings are of general interest and are recorded here.

II. Method

Fig. 1 shows a Y junction, in which it is assumed that the arms of the Y have double the line impedance of the main line, so that the junction is matched to first order. The half-angle of the Y is denoted by β. The Hertzian vector $\mathbf{\Pi}$ is calculated by the process of [1], and the far fields are deduced from the relations

$$E_\theta = k^2\left[\cos\theta\cos\phi\Pi_x + \cos\theta\sin\phi\Pi_y - \sin\theta\Pi_z\right] \quad (1)$$

$$E_\phi = -k^2\left[\sin\phi\Pi_x - \cos\phi\Pi_y\right] \quad (2)$$

where k is the free-space wavenumber. From the far fields the Poynting vector is found and is integrated over a hemisphere to obtain the radiated power. This integration is in principle straightforward, following the lines of the Appendix in [1], but in the present instance it is extremely lengthy and tedious to perform. (This appears to be the case quite generally for other than parallel antennas.)

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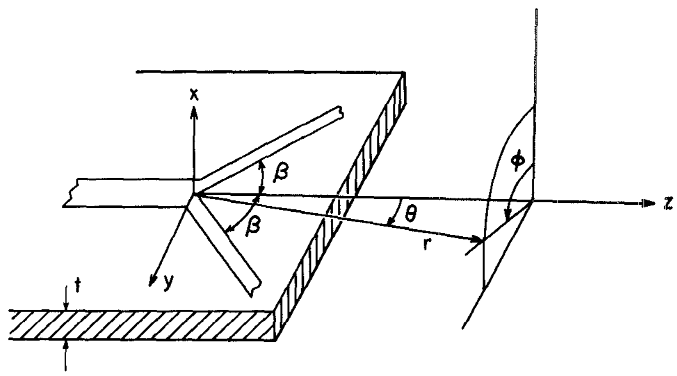

Fig. 1. Geometry of Y junction in microstrip.

III. Results

The far-field electric-field components are found, for a unit current wave incident from the main line to the junction, to be given by

$$E_\theta = \left(-j60kte^{-jkr}/r\right)\frac{\cos\phi}{\epsilon^{1/2}} \cdot\left[1 - \frac{(\epsilon^{1/2}-\cos\theta\cos\beta)(\epsilon^{1/2}\cos\beta-\cos\theta)}{D}\right] \quad (3)$$

$$E_\phi = \left(-j60kte^{-jkr}/r\right)\frac{\sin\phi}{\epsilon^{1/2}} \cdot\left[\frac{(\epsilon^{1/2}\cos\theta-\cos\beta)(\epsilon^{1/2}\cos\beta-\cos\theta)}{D} - \frac{\epsilon^{1/2}\cos\theta-1}{\epsilon^{1/2}-\cos\theta}\right] \quad (4)$$

where

- D $(\epsilon^{1/2}-\cos\theta\cos\beta)^2-\sin^2\theta\sin^2\phi\sin^2\beta$,
- t substrate thickness,
- ϵ substrate (relative) permittivity,
- r distance from Y junction to far-field point.

The radiated power takes the form

$$P = 60(kt)^2F \quad (5)$$

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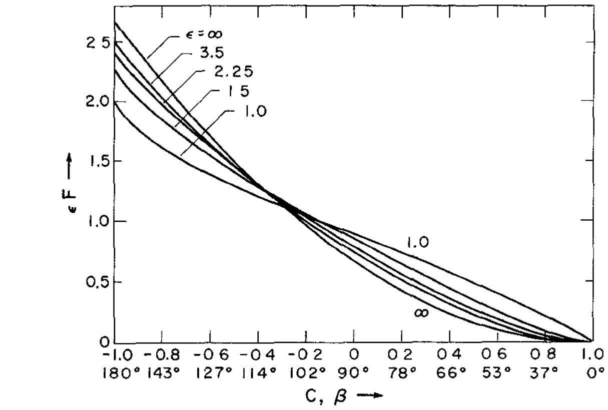

Fig. 2. Plot of ϵF versus c or β.

where

$$F=\frac{\epsilon+1-2c^2}{4c^2}\cdot\frac{s}{(\epsilon-c^2)^{1/2}}\ln\frac{(\epsilon-c^2)^{1/2}+s}{(\epsilon-c^2)^{1/2}-s}-\frac{1-c}{1+c}$$

$$\cdot\left\{\frac{(\epsilon+1)(1-c)(1+3c)}{4c^2}\frac{1}{\epsilon^{1/2}}\ln\frac{\epsilon^{1/2}+1}{\epsilon^{1/2}-1}+\frac{2(\epsilon-c)}{\Gamma}\ln\frac{\epsilon-c+\Gamma}{\epsilon-1}\right\} \tag{6}$$

and where $c=\cos\beta$, $s=\sin\beta$, and $\Gamma=[2\epsilon(1-c)-s^2]^{1/2}$. For large ϵ it is found that

$$F \sim 2(1-c)^2/3\epsilon. \tag{7}$$

Fig. 2 shows the quantity ϵF plotted against c or β for a range of values of ϵ. For β small the radiation is clearly negligible. It rises to a maximum at $\beta=180°$, corresponding to an open-circuit condition; it can be verified that (6) then reduces to (14) in [1] when $c\to-1$.

It can also be shown that as $\epsilon\to1$ (6) takes the limiting form

$$F_{\epsilon=1}=2\frac{1-c}{1+c}\ln\frac{2}{1-c}+\frac{s^2}{c^2}\ln s. \tag{8}$$

There may be some coupling between the junction arms, particularly for small β, and this has not been considered in the above calculations. This stipulation apart, it is clear that the spurious radiation can be kept low by using small junction angles and large substrate dielectric constants.

References

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Air-Gap Effect in Rectangular Waveguide Containing a Lossy *H*-Plane Dielectric Slab

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***Abstract*—The propagation coefficient for a partially filled rectangular waveguide containing a lossy *H*-plane slab against the broad wall can be significantly altered if an air gap exists between the slab and the wall of the waveguide. The solutions of the dispersion equation show that the attenuation and phase coefficients may be increased as well as decreased by the presence of an air gap. For a fully filled waveguide the effect of an air gap is maximized if the gap is equally distributed at the top and bottom of the sample.**

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I. Introduction

FOR CERTAIN dielectric materials the *H*-plane loaded waveguide provides a convenient method [1] for determining material properties. The theoretical and experimental procedures are simplified if the sample is in complete contact with the broad wall of the waveguide. However, it is often difficult to eliminate air gaps completely due to problems such as imperfections in the waveguide, for example, rounded internal corners or warping of the sample itself. It has been shown previously [2] that the attenuation and phase coefficient decrease rapidly as the sample is moved away from the waveguide

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