Prediction of Surface Wave Radiation Coupling on Microwave Planar Circuits

Danielle Vanhoenacker, *Member, IEEE,* and Isabelle Huynen, *Member, IEEE*

***Abstract*—A theoretical and experimental study is presented for spurious radiation coupling induced by radiated surface waves on planar circuits, at frequencies up to 20 GHz. The radiation coupling phenomenon generated by a planar line containing a discontinuity is modelled by an equivalent E-field planar radiation pattern associated with the current waveform along this line in the presence of the discontinuity. It is observed that the direction of the main and side lobes of this pattern varies with frequency. The model is validated by measurements of radiation coupling obtained on microstrip configurations, up to 20 GHz. Theory and measurements agree very well. The model yields for instance the directions of maximum spurious radiation coupling on a planar line in the vicinity of a radiating line.**

I. Introduction

THE INVESTIGATION of radiation phenomena supported by microwave planar circuits is twofold. On one hand, particular efforts have been devoted to the study of "wanted effects," like for instance modeling and designing planar antennas in microstrip [1] and slotline [2] technologies, requiring the calculation of the input impedance and of the radiation patterns. On the other hand, designers of planar microwave circuits are complaining about "unwanted effects," like mismatch and losses induced by radiating planar lines and discontinuities.

It is generally assumed that a uniform planar line exhibits low space radiation losses. These losses are generated when discontinuities are present. As a consequence, a particular attention has been focused on the calculation of an equivalent radiation reactance for open- and shorted-ended microstrip lines [3] and resonators [4], taking into account the power lost by radiation into the surrounding space. On the other hand, planar lines exhibit losses due to surface waves [5] and leakage [6] phenomena. These two phenomena are related to the presence of poles in the spectral Green's function associated to the problem. Finally, it has also to be mentioned that the work dealing with spurious coupling induced by radiation has up to now been concentrated on modeling radiation phenomena induced on lines by incident spatial electromagnetic waves [7], [8].

In this letter we present a new characterization method for spurious coupling on planar lines due to surface waves radiation. The coupling phenomenon induced by surface waves radiated by a planar line containing a discontinuity is described by the way of an equivalent E-field planar radiation pattern associated with the frequency-dependent standing wave current existing along this line due to the discontinuity. When calculated over a wide frequency range, this equivalent radiation pattern is observed to rapidly change with frequency: the direction of the main and side lobes continuously varies with the frequency. This effect is confirmed by measuring radiation coupling on a microstrip line produced by a discontinuous microstrip line: the frequencies where an extremum of measured radiation coupling occurs are the frequencies at which the calculated radiation pattern exhibits an extremum in the direction of the irradiated line. The proposed characterization is wide-band, and yields a fast estimation of the preferential directions where spurious radiation coupling occur between two microstrip elements at a given frequency, without having to integrate the contribution of the resulting fields on the irradiated line.

Manuscript received January 27, 1995. This work was supported in part by the Belgian Scientific Policy of the PAI-program.

The authors are with the Microwaves U.C.L., Bâtiment Maxwell, B-1348 Louvain-la-Neuve, Belgium.

IEEE Log Number 9412819.

II. Theoretical Modelling

Surface waves occur when the spatial form of the Green's function relating the current to the electric field in planar structures exhibits poles, solutions of the characteristic equation of surface waves into dielectric slabs. The equation may exhibit a pole when the product thickness of the layer-frequency is high enough. They are responsible for spurious radiation coupling between various subsystems included into a microwave planar circuit.

We have developed a new theoretical and experimental method for characterizing spurious radiation coupling mechanisms. First we compute the longitudinal dependence of the current along a planar line ended by a discontinuity, taking into account the effect of the discontinuity. This current is then discretized into successive elementary longitudinal currents dipoles, and we calculate the electric field radiated by surface waves for each dipole. This field is extracted from the surface wave term established by Mosig and Gardiol for a current dipole at the interface between air and a dielectric layer on a ground plane [5]. Defining $\overline{az}$ as the longitudinal direction (Fig. 1), $I(z')$ the longitudinal dependence of the current on the line at abscissa z', and (ρ, ϕ, x) the cylindrical coordinate system centered at abscissa z', the surface wave radiated electric field at abscissa z generated by the current dipole $I(z')\,dz'$ is

$$\overline{E}(x,y,z) = \frac{k_r^{3/2} e^{j3\pi/2}}{\sqrt{2\pi}\omega\varepsilon_o} \frac{e^{-jk_r\rho}}{\sqrt{\rho}} R_o u_o \frac{\cosh[u_1(x+H)]}{\cosh(u_1 H)} \cdot \cos\varphi I(z')\,dz'\,\overline{ax} \tag{1}$$

1051-8207/95$04.00 © 1995 IEEE

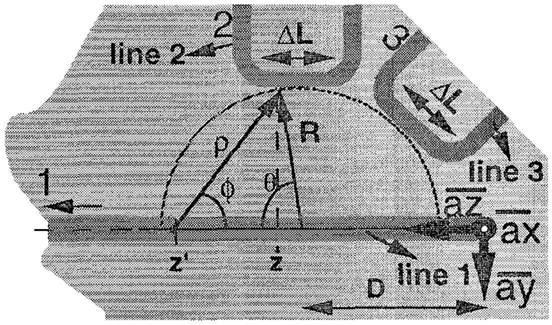

Fig. 1. Circuit configuration for the measurement of surface wave radiation coupling. Measurements are carried between ports 1 and 2 for $\theta = 90°$ and between ports 1 and 3 for $\theta = 135°$.

where the asymptotic expression for the Hankel function of second kind and order 1 has been used. In this expression k_r is the solution of the following characteristic equation for the propagation constant of TM_0 surface waves in a dielectric slab having a thickness H and a permittivity ε_r

$$\mathrm{DTM}\,(k_r) = \varepsilon_r u_o + u_1 \tanh\left(u_1 H\right) = 0. \tag{2}$$

Furthermore one has

$$\begin{aligned} u_o &= \sqrt{k_r^2 - k_o^2} \\ u_1 &= \sqrt{k_r^2 - \varepsilon_r k_o^2} \\ \rho &= \sqrt{y^2 + (z - z')^2} \end{aligned}.$$

Finally, R_o is the residue associated to the TM_0 surface wave, obtained as

$$R_o = \lim_{k \to k_r} \frac{k - k_r}{\mathrm{DTM}} = \left(\left. \frac{\partial \mathrm{DTM}\,(k)}{\partial k} \right|_{k = k_r} \right)^{-1}. \tag{3}$$

Only the E-field is considered, since it is $\overline{ax}$-oriented as the quasi-TEM E-field component of the dominant mode on a microstrip line, and will therefore induce the major contribution to the radiated field absorbed by such a line.

We obtain at each point (R, θ, x) the total E-field component (1) from the integration, along the z'-axis in (4), of the contributions of the elementary dipoles over the length of the radiating line. The result is an equivalent E-field pattern $A(R, \theta, x)$ of the radiating line, centered at a distance D from the open-ending discontinuity of the radiating line, which is assumed to be fed by a matched source at its other end (abscissa $z' = \infty$)

$$A(R, \theta, x) = \frac{\cosh\left[u_1(x + H)\right]}{\cosh\left(u_1 H\right)} \int_0^{\infty} \frac{k_r^{3/2} e^{j3\pi/2}}{\sqrt{2\pi\omega\varepsilon_o}} \frac{e^{-jk_r\rho}}{\sqrt{\rho}} \cdot R_o u_o \cos\varphi I(z')\, dz' \tag{4}$$

with

$$\begin{aligned} z &= D + R\cos\theta \\ \cos\varphi &= \sqrt{1 - (R\sin\theta/\rho)^2} \end{aligned}.$$

The radiation pattern $A(R, \theta, x)$. is integrated along the x-axis in order to obtain the radiation coupling voltage induced on a length ΔL of the irradiated line. Hence, the resulting radiation coupling pattern $T(R, \theta, \Delta L)$ is equal to (5) shown at the bottom of the page, where β_{strip} and Z_{cstrip} are the propagation constant and characteristic impedance of the microstrip lines respectively, while $Z_{r\mathrm{surf}}(\theta)$ is the equivalent TEM characteristic impedance of the surface wave in the direction of propagation of the quasi-TEM mode along the ΔL line, and is defined as the ratio of the total electric field component $E_x(\theta, R)$ by the total magnetic field component $H_R(\theta, R)$.

We have computed this equivalent pattern at several frequencies up to 20 GHz, for a microstrip open-ended line with $D = 10$ mm. The result is shown at Fig. 2 for an integration length equal to $2D$. An azimuthal variation of the radiation pattern is observed when the frequency varies. Hence, the direction where maximum coupling occurs between the radiating line and any other planar element in its vicinity varies with frequency: at a fixed position, the coupling level induced on an irradiated line changes with frequency. As examples, Fig. 2(a) shows that at 8 and 17 GHz there is no radiation in the direction perpendicular to the line, while at 9.8, 15, and 20 GHz [Fig. 2(b)] there is a maximum of radiation in that direction.

III. Experimental Validation

The calculated effect, illustrated on Fig. 2, has been experimentally verified with success, using the microstrip configuration shown on Fig. 1 (width of the lines = 1.5 mm, relative dielectric constant of the substrate = 2.33, thickness H of the substrate = 0.5 mm, characteristic impedances of the lines = 50 Ω). For the measurements, a vector network analyzer was used, as well as broadband test fixtures for the microstrip-to-coaxial transitions. The edges of the circuit of Fig. 1 were covered by an absorbing material in order to avoid spurious reflections of the radiated waves. Fig. 3(a) shows the measured transmission between the input port 1 of the radiating open-ended microstrip line 1 and port 2 of the curved microstrip line 2 with its other output matched while Fig. 3(b) shows the same measurement between port 1 and 3 for the oblique line 3. Fig. 3 shows that, at the frequencies used to calculate Fig. 2, relative extrema occur in the measured transmissions, in accordance with the prediction of the equivalent radiation coupling patterns: in the direction $\theta = 90°$ [Fig. 3(a)], low

$$T(R, \theta, \Delta L) = 20 \log 10 \left(\frac{\left| \sqrt{Z_{r\mathrm{surf}}(\theta)}\, j(\beta_{\mathrm{strip}}/Z_{\mathrm{cstrip}}) \dfrac{e^{-j\beta_{\mathrm{strip}}\Delta L/2}}{\beta_{\mathrm{strip}}} \displaystyle\int_{-H}^{0} A(R, \theta, x)\, dx \right|}{2\sqrt{Z_{\mathrm{cstrip}}}} \right) \tag{5}$$

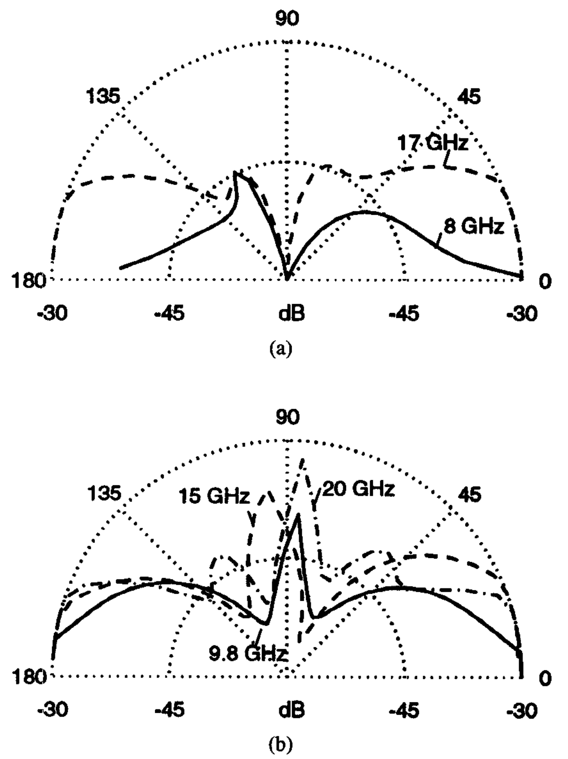

Fig. 2. Radiation coupling patterns calculated at several frequencies for $D =$ 10 mm, a source line of length $2D$, and an irradiatied line of length $\Delta L =$ 10 mm. (a) 8 GHz (solid) and 17 GHz (dashed), (b) 9.8 GHz (solid), 15 GHz (dashed) and 20 GHz (dot-dashed).

values are indeed obtained at 8 and 17 GHz [Fig. 2(a)] and high values at 9.8, 15 and 20 GHz [Fig. 2(b)], while in the direction $\theta = 135°$ [Fig. 3(b)], intermediate values are obtained at 8 and 9.8 GHz [Fig. 2(a)] and high values at 15, 17, and 20 GHz (Fig. 2(b)). Hence, the predicted transmission levels are in good agreement with the measured ones.

IV. Conclusion

We have developed a simple method for evaluating the directional effects of surface wave radiation coupling on microstrip planar circuits. It is based on the definition of an equivalent E-field planar radiation coupling pattern for the radiating line. The frequency-dependent behavior of the radiation pattern is illustrated. It is successfully confirmed by measurements carried out from 0–20 GHz on very simple microstrip configurations. This model will help the design of layouts for planar microstrip circuits minimizing spurious surface wave radiation coupling.

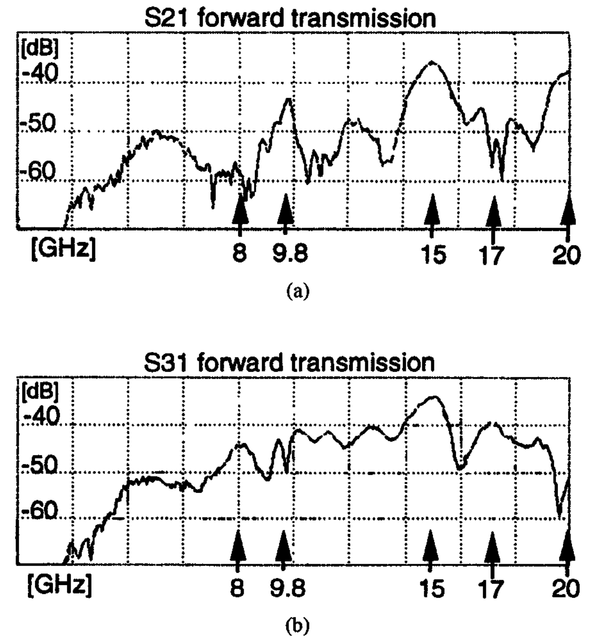

Fig. 3. Measured radiation coupling versus frequency. (a) Transmission between ports 1 and 2 (90°). (b) Transmission between ports 1 and 3 (135°).

Acknowledgment

The authors like to thank Prof. A. Vander Vorst for a number of stimulating discussions and Mr. R. Sciuto who measured the circuits.

References

[1] J. R. James, P. S. Hall, and C. Wood, *Microstrip Antennas Theory and Design.* Stevenage, U.K.: Peregrinus Ltd., 1981.

[2] A. Rabbaa, "An experimental technique for increasing the bandwidth of printed slot antennas," *Microwave and Opt. Technol. Lett.*, vol. 4, no. 4, pp. 161–166, Mar. 1991.

[3] L. J. van der Pauw, "The radiation of electromagnetic power by microstrip configurations," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-25, no. 9, pp. 719–725, Dec. 1977.

[4] E. J. Denlinger, "Radiation from microstrip resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, no. 4, pp. 235–236, Apr. 1969.

[5] J. Mosig and F. Gardiol, "Rayonnement d'une antenne microruban de forme arbitraire," *Ann. Télécommun.*, vol. 40, pp. 181–189, Mar./Apr. 1985.

[6] A. A. Oliner and K. S. Lee, "The nature of the leakage from higher order modes in microstrip line," in *IEEE MTT-S Int. Microwave Symp. Dig.*, 1986, pp. 57–60.

[7] C. Paul, "Frequency response of multiconductor transmission lines illuminated by an electromagnetic field," *IEEE Trans. Electromagn. Compat.*, vol. EMC-18, no. 4, Nov. 1976.

[8] R. Ciccetti and C. Pirone, "On the radiation sensibility of a microstrip transmission line," *Microwave and Opt. Technol. Lett.*, vol. 6, no. 14, Nov. 1993.