Stripline to Microstrip Line Aperture Coupler

G. V. JOGIRAJU AND V M. PANDHARIPANDE

Abstract —**The paper presents the analysis of aperture coupling between stripline and microstrip line as a function of the electric and magnetic polarizabilities of the aperture using Bethe's small-aperture coupling theory. A coupling expression for the general case of an aperture in the common ground plane is derived. Coupling variations with aperture dimensions for the case of a thin slot, a hole, a diamond, a rounded end slot, and an ellipse are determined using McDonald's polynomial approximations for the electric and magnetic polarizability for small apertures. Experimental results for the case of a thin slot and a circular aperture are also presented.**

Nomenclature

V	Amplitude of incident TEM wave.
Z_1	Characteristic impedance of TEM wave in stripline.
Z_2	Characteristic impedance of TEM wave in microstrip line.
b	Height of stripline.
D	Width of parallel-plate equivalent microstrip line ($=b/4Z_0 \cdot 120\ \pi/\sqrt{\epsilon_{eff}}$).
P	Electric dipole moment.
M	Magnetic dipole moment.
A_1, A_2 A_3, A_4	Amplitude coefficients.
$2l$	Length of the rectangular slot aperture.
$2t$	Width of the rectangular slot aperture.
d_0	Circular aperture diameter.
λ	Operating wavelength.
ϵ_0	Dielectric constant of free space.
ϵ_r	Relative dielectric constant of substrate used ($=2.32$).
ϵ_{eff}	Effective dielectric constant of microstrip medium.
μ_0	Free-space permeability.
β	Phase constant of propagating TEM mode in stripline ($=2\pi/\lambda_e = \omega\sqrt{\mu_0\epsilon_0\epsilon_r}$).
λ_{g2}	Wavelength in microstrip line ($=\lambda_0/\sqrt{\epsilon_{eff}}$).
m	$\operatorname{sech}(\pi w/2b)$.
α_e	Electric polarizability coefficient.
α_m	Magnetic polarizability coefficient.

I. Introduction

Microwave coupling through apertures has been investigated by various authors [1]–[3] using guiding systems such as waveguides, striplines, and microstrip lines. The compactness of the microwave integrated circuit using a stripline to microstrip line coupler can be realized with the components mounted at the coupled arm of the microstrip line and the coupling achieved through an aperture in the common ground plane.

This paper presents an analysis of coupling between stripline and microstrip lines through an aperture. The coupling is shown as a function of the electric and magnetic polarizabilities of the aperture. For the microstrip line, an equivalent parallel-plate transmission line is assumed taking into account the effective dielectric constant of the medium. Bethe's [4] polarizability theory is used to derive the general expressions for coupling through an arbitrary aperture. Theoretical coupling is determined for the common apertures, viz. slot, hole, rounded end slot, ellipse, and diamond. The experimental results are presented for the cases of thin slot and circular aperture.

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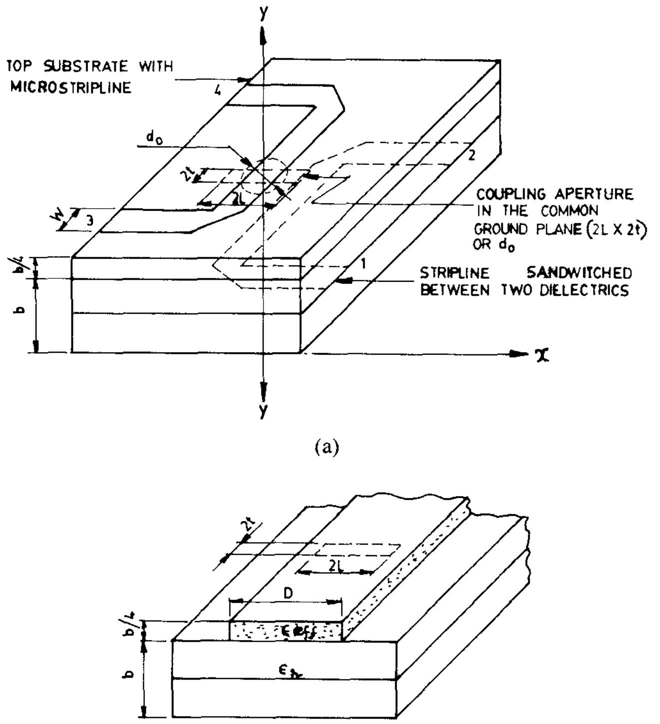

Fig. 1 Stripline to microstrip line aperture coupler. (a) Coordinate geometry. (b) Equivalent parallel-plate configuration.

II. Analysis of Coupling

Fig. 1 shows the coordinate system and the associated structure of a stripline to microstrip line coupler. Let the normalized TEM fields in the stripline be represented by [2]

$$E_y = -\frac{Va_y}{b\sqrt{F(m)}}\left[\frac{1}{1+m\sinh^2\frac{\pi(x+jy)}{b}}\right]^* e^{-j\beta z}$$

$$H_x = Y_1 E_y. \tag{1a}$$

On the ground plane, at $y=b$ the fields are

$$E_y|_{y=b} = \frac{-V_{ay}}{b\sqrt{F(m)}}\left[\frac{1}{1+m\sinh^2 \pi x/b}\right]e^{-j\beta z}$$

$$H_x|_{y=b} = -Y_1 E_y|_{y=b}. \tag{1b}$$

In the above expressions V is the amplitude of the incident TEM wave and

$$F(m)=\frac{4}{b^2}\int_{-b/2}^{b/2}\int_{-\infty}^{\infty}\frac{dx\,dy}{\left|1-m^2\cosh^2\frac{\pi(x+jy)}{b}\right|}.$$

The parallel-plate equivalence of microstrip line is shown in Fig. 1(b) having a width

$$D=\frac{b}{4Z_0}\cdot\frac{120\pi}{\sqrt{\epsilon_{\text{eff}}}}. \tag{2}$$

Van Bladel [5] showed that the electric dipole moment $\boldsymbol{P}$ can be expressed in terms of the electric polarizability dyadic $\bar{\tau}$ and electric field normal to the aperture. Similarly, the magnetic dipole moment $\boldsymbol{M}$ can be expressed in terms of the magnetic polarizability dyadic $\boldsymbol{\rho}$ and the tangential magnetic field. $\boldsymbol{P}$ and $\boldsymbol{M}$ can be obtained from the following relations [1]:

$$\boldsymbol{P}=\epsilon\tau\cdot\boldsymbol{E}_y \tag{3a}$$

where

$$\epsilon=\epsilon_0\left(\frac{\epsilon_r\epsilon_{\text{eff}}}{\epsilon_r+\epsilon_{\text{eff}}}\right)$$

and

$$\boldsymbol{M}=\boldsymbol{\rho}\cdot\boldsymbol{H}x. \tag{3b}$$

Let the fields coupled into the coupled arm by the electric dipole be

$$\boldsymbol{E}_1'=A_1\exp\left(-j\frac{2\pi}{\lambda g_2}\cdot z\right)\qquad z>0 \tag{4a}$$

$$\boldsymbol{H}_1'=-\frac{A_1}{z_2}\exp\left(-j\frac{2\pi}{\lambda g_2}\cdot z\right)\qquad z>0 \tag{4b}$$

and

$$\boldsymbol{E}_1'=A_3\exp\left(j\frac{2\pi}{\lambda g_2}\cdot z\right)\qquad z<0 \tag{4c}$$

$$\boldsymbol{H}_1'=\frac{A_3}{z_2}\exp\left(j\frac{2\pi}{\lambda g_2}\cdot z\right)\qquad z<0 \tag{4d}$$

where

$$\lambda_{g_2}=\lambda_0/\sqrt{\epsilon_{\text{eff}}}.$$

Similar expressions for coupling due to the magnetic dipole can be written replacing A_1 with A_2 and A_3 with A_4.

The forward and backward couplings due to electric and magnetic dipoles are given by

$$A_1=A_3=-\frac{1}{P_{ms}}(j\omega P_y) \tag{5a}$$

$$A_2=-A_4=\frac{j\omega\mu_0}{P_{ms}}\cdot\alpha_m\cdot\frac{Y_0VY_2}{b\sqrt{F(m)}} \tag{5b}$$

where the normalizing factor P_{ms} is given by

$$P_{ms}=-2\int_0^w\int_0^{b/4}Y_0\,dx\,dy=\frac{wb\sqrt{\epsilon_{\text{eff}}}}{240\pi}.$$

Using (5a) and (5b), the coupling between ports 1 and 3 is given as

$$A_{13}\text{ in dB}=20\log_{10}\left(\frac{A_1+A_2}{V}\right)$$
$$=20\log_{10}\left\{\frac{j\omega}{P_{ms}\sqrt{F(m)}}\left(\epsilon\alpha_e+\frac{\mu_0\alpha_m}{z_1z_2}\right)\right\}. \tag{6}$$

After substituting for values of P_{ms}, z_1, z_2, and ϵ and simplifying, we obtain the general expression for coupling due to an arbitrary aperture:

$$(A_{13})_{\text{dB}}=20\log_{10}\frac{20}{3\lambda b^2\sqrt{F(m)}}\left\{\frac{\alpha_e\epsilon_r\epsilon_{\text{eff}}}{\epsilon_r+\epsilon_{\text{eff}}}+\alpha_m\sqrt{\epsilon_r\epsilon_{\text{eff}}}\right\}. \tag{7}$$

Case 1: Circular Aperture

For the case of a circular aperture, after substituting the electric and magnetic polarizability coefficients the coupling expression takes the form

$$(A_{13}')_{\text{dB}}=20\log_{10}\frac{5d_0^3}{9\lambda b^2\sqrt{F(m)}}\left\{\frac{-\epsilon_r\epsilon_{\text{eff}}}{\epsilon_r+\epsilon_{\text{eff}}}+2\sqrt{\epsilon_r\epsilon_{\text{eff}}}\right\}. \tag{8}$$

Case 2: Slot Aperture

The polynomial expressions for α_e and α_m deduced by McDonald [6], [7] are

$$\alpha_e=\frac{\pi}{2}\alpha^2l^3(1-0.5663\alpha+0.1398\alpha^2) \tag{9a}$$

$$\alpha_m=\frac{1.054l^3}{\log_e\left(1+\frac{0.66}{\alpha}\right)}$$

where

$$\alpha=\frac{t}{l}. \tag{9b}$$

The coupling is computed by substituting (9a) and (9b) in (6). In a similar way, the variation of coupling with aperture dimensions is computed for various geometrical shapes. Fig. 2 shows the theoretical coupling as a function of major dimension for a thin ellipse, rounded end slot, and diamond. For experimental verification, the thin slot and the hole were chosen as coupling apertures.

III. Results

The factor $F(m)$ of the stripline is evaluated by a Fortran computer program [8]. The substrate parameters used in the above program are $b=3.2$ mm, $w=2.4$ mm, and $\epsilon_r=2.32$ at a frequency of 4 GHz, giving a value of $F(m)$ of 11.862. Fig. 3 shows the theoretical backward coupling computed from expressions (6) and (7) for a thin rectangular slot and a circular aperture, respectively.

A 50 Ω stripline and a microstrip line were etched back to back on RT Duroid substrate with $\epsilon_r=2.32$. The two different transmission lines are edge grounded by a thin copper foil and mounted in an aluminum package by means of screws. The apertures with different dimensions were etched in the common ground plane. Fig. 3 shows the experimental data on coupling variation with hole diameter and slot length measured at 4 GHz using a network analyzer.

IV. Conclusion

The above analysis of coupling, which takes into account the fields in the stripline as well as the equivalent parallel-plate configuration for microstrip line, shows good agreement with the

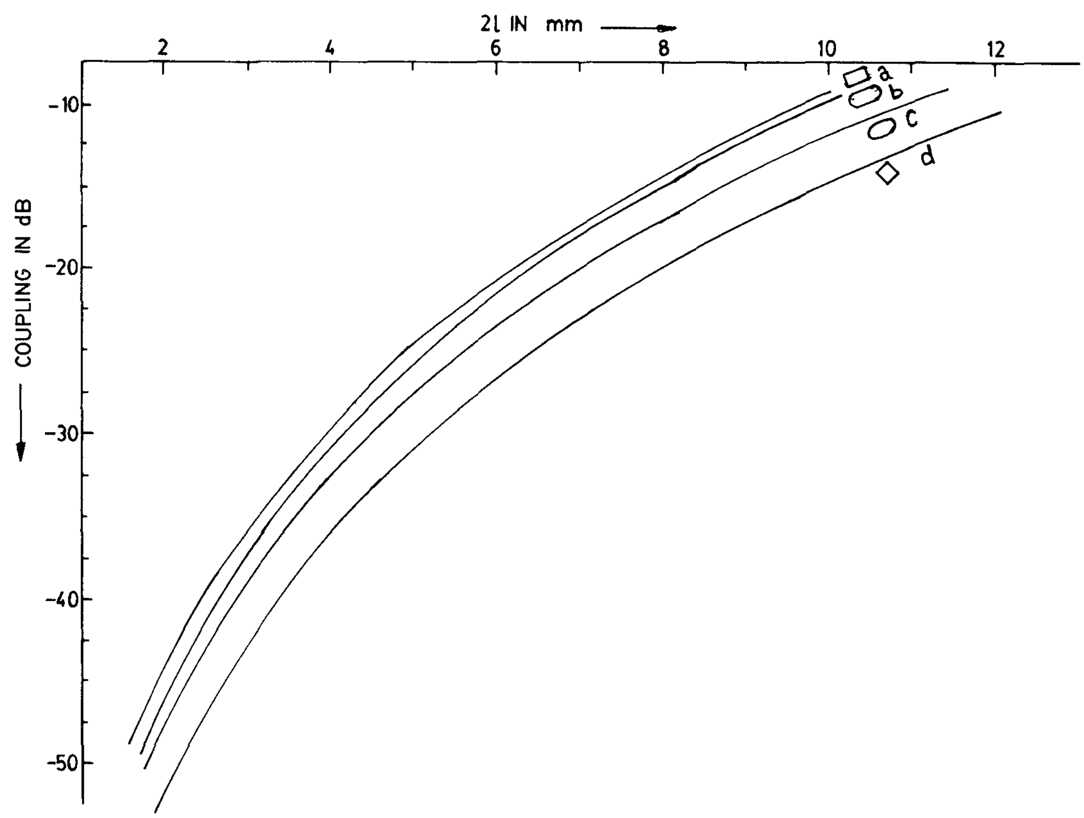

Fig 2. Variation of coupling with aperture dimensions for different aperture shapes: *a* rectangular slot, *b* rounded end slot; *c* elliptical aperture; *d* diamond-shaped slot.

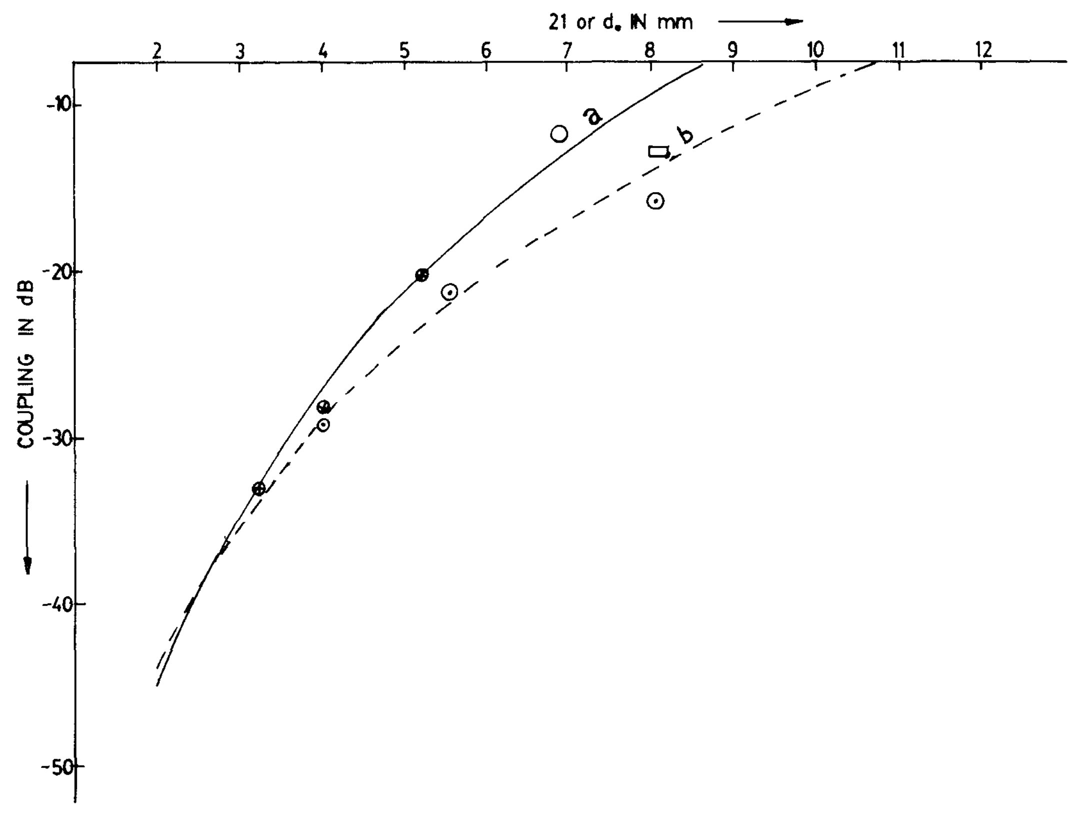

Fig 3 Variation of coupling with aperture dimensions for *a* circular aperture and *b* rectangular slot· ··· experimental results for slot; ⊗ ⊗ ⊗ experimental results for circular aperture

experimental data. Such couplers are useful for sensing the power of an airborne transponder and for other communication applications. The results are useful for the design of a well-matched multiaperture directional coupler between stripline and microstripline.

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Additional Approximate Formulas and Experimental Data on Micro-Coplanar Striplines

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Abstract —**Additional approximate design formulas and experimental data of micro-coplanar striplines, which have recently been proposed for high-packing-density MMIC's, are given in this paper to be applied to four types of substrate materials: GaAs, plastics, and alumina with different dielectric constants.**

I. Introduction

In connection with the design of high-packing-density microwave monolithic integrated circuits (MMIC's), Pucel has discussed the necessity of solutions for two types of proximity effects [1]. One type is observed between the microstrip conductor and closely located conductor having ground potential on the top of the same substrate [2]. The other type occurs when microstrip lines are near a substrate edge [3].

The former effects have been analyzed by one of the authors [2] and the effects of the location of the upper ground conductor on the characteristics of the microstrip lines have been discussed. The micro-coplanar stripline (MCS) structure, shown in Fig. 1, has been proposed as a measure to avoid this type of proximity effect. Namely, the structural dimensions of micro-coplanar striplines are so designed that the characteristic impedance is kept at a constant value even when the ground potential conductor is located close to the strip conductor. With this structure, the packing density of MMIC's can be enhanced, and shunt element connection between the strip conductor and the upper ground conductor can be easily realized.

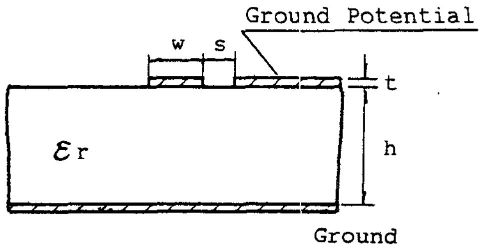

Fig 1 Cross-sectional view of micro-coplanar striplines.

In this paper, we present several approximate design formulas of MCS's in addition to previous ones [2] assuming the use of other substrate materials with different dielectric constants. The analysis is based on the rectangular boundary division method given in the previous paper [2]. These approximate formulas are derived by applying a least-square curve-fitting procedure to the results of our computation. Experimental data on the MCS's with an alumina substrate are also compared with theoretical values.

II. Design Formulas of MCS's for Different Substrate Materials

The design procedures of the MCS's consist of finding the strip width (w) for keeping the characteristic impedance at 50 Ω against specified values of separation (s), the subsequent guide wavelength (λ) of such 50 Ω MCS's, and preferably, their conductor attenuation constant (α_c). Since theoretical derivations of relevant expressions are the same as those in the previous paper [2], they are not repeated here. Our numerical computation covers four types of substrate materials in current use, namely, GaAs ($\epsilon_r = 12.9$), plastics ($\epsilon_r = 2.22$), and two kinds of alumina (ϵ_r values of 9.7 and 10.1). To simplify the expressions, the calculated results of w, λ, and α_c are normalized by w_0 (the strip width of the 50 Ω microstrip line), λ_0 (free-space wavelength), and $\sqrt{f/f_0}$ respectively. Here f_0, the frequency for the numerical estimation of the attenuation constant, is taken as 10 GHz, and f is the frequency to be used.

Using a least-square curve-fitting procedure, all the theoretical data of w/w_0, λ/λ_0, and $\sqrt{f/f_0}\,\alpha_c$ are expressed in simple polynomial formulas as follows:

$$\frac{w}{w_0} = \sum_{i=0}^{M} a_{0i}u^i + \left(\frac{t}{h}\right)\sum_{i=0}^{M} a_{1i}u^i \tag{1}$$

$$\frac{\lambda}{\lambda_0} = \sum_{i=0}^{N} b_{0i}u^i + \left(\frac{t}{h}\right)\sum_{i=0}^{N} b_{1i}u^i \tag{2}$$

and

$$\sqrt{f/f_0}\,\alpha_c = \sum_{i=0}^{P} C_{0i}u^i + \ln\left(\frac{t}{h}\right)\sum_{i=0}^{P} C_{1i}u^i + \left[\ln\left(\frac{t}{h}\right)\right]^2\sum_{i=0}^{P} C_{2i}u^i \tag{3}$$

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