Letters

A Small MIC Coupler with Good Directivity

R. R. WEIRATHER

Abstract—**The results of tests made on a novel coupler design suitable for MIC's are described. The design utilizes resistive termination of fields above the dielectric. Directivity of 14 dB was achieved with a λ/16 coupler at 3.6 GHz. Uses for this design technique are signal sampling where very flat coupling is not required.**

Microwave integrated circuit (MIC) designers are frequently required to realize a coupler on high dielectric materials. Edge-coupled microstrip couplers on dielectric materials have been examined by several investigators [1]–[3] and each was aware of the lack of directivity MIC couplers can exhibit. The fundamental problem is that of differing phase velocities for the even- and odd-mode impedances. Another solution to this problem has been examined and it is easily implemented. The solution is for lightly edge-coupled lines on high dielectric materials such as sapphire or alumina.

It is frequently necessary to sample a microwave signal, and the need for a directional coupler becomes evident. The MIC designer must examine the even- and odd-mode fields as required by a familiar technique [4]. As shown in Fig. 1, the odd-mode field lines exist proportionally more in air. To achieve cancellation of the mode components necessary for high directivity, the two mode velocities must be equal. This is not realizable with edge-coupled microstrip couplers without compensation. Instead of compensation for the mode inequalities, it is possible to introduce some resistive material in between the couplers to terminate the odd-mode field in air that causes the poor directivity.

This concept has been tried and the results are encouraging. A need existed to sample a signal in high *S* band with a minimum of circuit area. The coupling desired was approximately 20 dB. To compute the line spacing, even-mode impedance, odd-mode impedance, and coupling region length, formulas as presented by Jones and Bolljohn [4] and Bryant and Weiss [5] were used. To minimize size, a λ/16 coupling region was chosen. The resistive overlay was extended to the middle of the coupling region as shown in Fig. 2. Test results on the coupler are shown in Fig. 3. The performance of the coupler is nearly that as predicted for a single coupling section <λ/4, nonuniform coupling with frequency. This detraction may or may not penalize some wide-band usages. Directivity was typically 14 dB compared to 3 dB or less for this coupler without the resistive overlay. The return loss of ports 1 and 2 were essentially that of the loads used, and is not shown in Fig. 3. Increased directivity was noted by adding open shunt stubs approximately λ/4 away from the coupling region on the coupled line. This further improved the directivity but narrowed the bandwidth. With this configuration, a minimum of 20-dB directivity was obtained over a 20-percent bandwidth centered at 3.6 GHz. The return loss of port 3 remained greater than 20 dB.

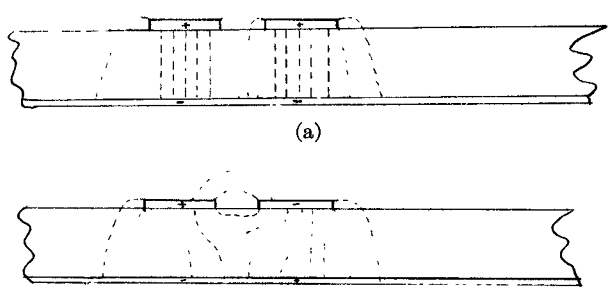

Fig. 1. Illustration of MIC coupler mode fields. (a) Even-mode electric field. (b) Odd-mode electric field.

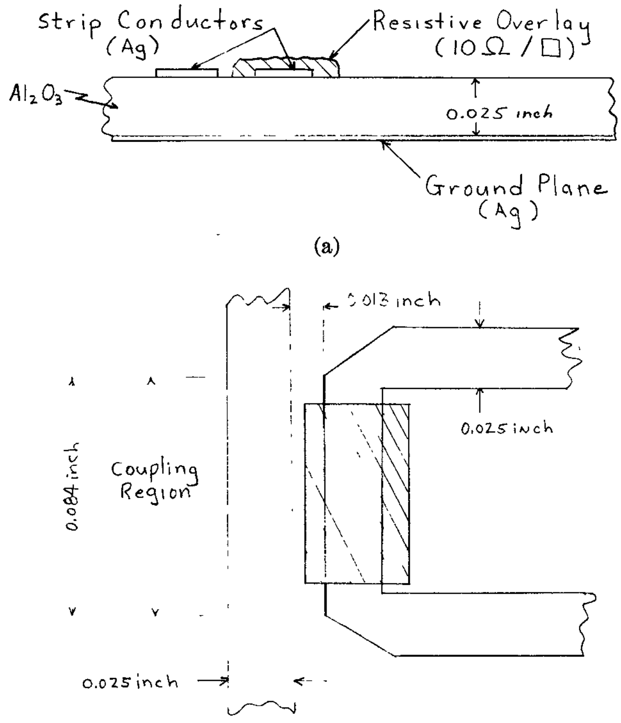

Fig. 2. MIC coupler. (a) Cross section. (b) Top view.

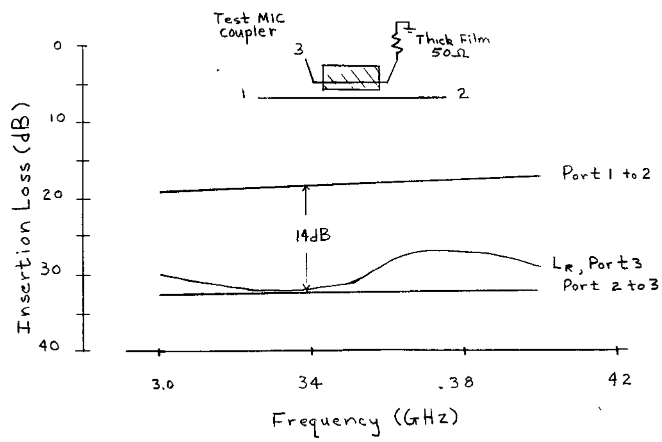

Fig. 3. Response of small MIC coupler as shown.

Conclusion

A small (0.01 in^2) *S*-band MIC coupler with good directivity has been examined. The design utilizes resistive termination of the fields above the edge-coupled region. No exact quantitative analysis has

References

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Comments on "A New Concept for Broadbanding the Ferrite Substrate Circulator Based on Experimental Modal Analysis"

STEVEN MARCH

***Abstract*—In the above paper,[1] the authors concluded, from a modal analysis of microstrip ferrite circulators, that the combination of two adjacent modes should be used for broad-banding circulators, which do not require quarter-wavelength matching transformers. Upon careful examination of their results, it can be shown that Miura and Hashimoto actually produced conventional below-resonance circulators—transformers included.**

Recently, Miura and Hashimoto proposed that broad-band microstrip circulators on ferrite substrates, without conventional matching transformers, can be realized by a combination of the $n = 1$ mode, as proposed by Bosma [2] and Fay and Comstock [1], and a separate, nonresonant mode. This second mode would be different from, but compatible with, the former one. Instead of magnetizing just the area enclosed by the shield diameter d a region of substantially larger diameter D was magnetized (Fig. 1). The optimum results were obtained from equi-isolation curves; the optimum ratio was $D/d = 1.90$. The optimum w/h was found to be 1.0. These results were obtained for a YIG substrate having a saturation magnetization $4\pi M_s$ of 1750 G. The experimental circulator had a center frequency of 6.3 GHz. Since Miura and Hashimoto did not present values of the relative dielectric constant of their YIG substrate nor the bias magnetization, a relative permittivity ϵ_r of 16.0 and a magnetization of 1250 G (approximately equal to the remanence magnetization) will be assumed.

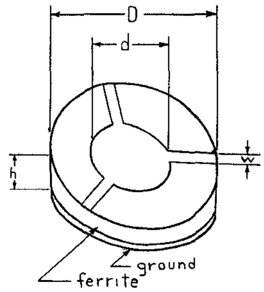

Fig. 1. Ferrite circulator geometry.

Circulators and Resonators

In the design of conventional Y-junction circulators, investigators have assumed that the z-component of the electric field traverses the periphery of the ferrite disk in only one period, i.e., only the $n = 1$ mode is present. For a below-resonance circulator operating in this mode, Fay and Comstock have found the diameter of the ferrite puck D_f to be related to the free-space wavelength λ_0 by:

$$D_f = \frac{0.586\lambda_0}{(\epsilon_{\text{eff}}\mu_{\text{eff}})^{1/2}} \quad (1)$$

where ϵ_{eff} and μ_{eff} are the effective permittivity and effective permeability of the ferrite, respectively.

Watkins [3], in studying the modal patterns of disk sections in microstrip, concluded that the $n = 1$ mode is dominant because, for a given frequency, it requires a minimum diameter. This resonator mode is also characterized by (1). Thus there exists a duality between circulators and circular disk resonators in microstrip.

In his study of circular resonators for microwave integrated circuits, Schwarzmann [4] reported the quality of the principal resonance to be poor, indicating a low value of Q. He found the following analytic expression for the unloaded Q of the disk resonator,

$$Q_u = 120(f_{\text{GHz}})^{1/2}. \quad (2)$$

Schwarzmann also reported the resonant frequency of the principal mode to be slightly lower than predicted by (1). This is due to fringe effects, not accounted for in the theoretical analysis, which decrease the resonant frequency. The effect of radiation is similar to fringing, but of lesser magnitude. Since the principle resonance is equivalent to a TM_{110} mode, the presence of a conducting ground perturbs the magnetic field and tends to increase the resonant frequency [5]. As reasonable estimates the combined effects of fringing and radiation would warrant a reduction in diameter by 8 percent, offset 3 percent by the magnetic field perturbations. Thus (1) should be used to define an effective ferrite or resonator diameter. A reasonably accurate estimate of the physical diameter would be

$$D_f' = \frac{0.557\lambda_0}{(\epsilon_{\text{eff}}\mu_{\text{eff}})^{1/2}} = 0.557\,\lambda_g. \quad (3)$$

How (3) relates to the work of Miura and Hashimoto will be made clear in the following.

Circulator Design

If we set D_f' equal to d of Fig. 1, and if we assume conventional circulator design which includes quarter-wave matching transformers, then it would be reasonable to conclude

$$D = d + 2\left(\frac{\lambda_g}{4}\right). \quad (4)$$

Then, from (3) and (4), it follows that D is 1.057 λ_g and the ratio D/d equals 1.898, in excellent agreement with the optimum results of Miura and Hashimoto. Thus the ratio D/d due to Miura and Hashimoto is nearly identical to that of conventional circulator design, assuming the presence of the $n = 1$ mode only.

As would be expected, the dimensions of Miura and Hashimoto for a circulator at 6.3 GHz also conform with established theory. In a microstrip media, the effective permittivity must be used in lieu of the relative dielectric constant. To determine the effective dielectric constant, (5) due to Schneider [6] yields highly accurate results at zero frequency

$$\epsilon_{\text{eff}_0} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\,\frac{1}{(1 + 10h/w)^{1/2}}. \quad (5)$$

For the disk structure at 6.3 GHz, Miura and Hashimoto used a shield diameter of 8 mm and a substrate thickness h of 1.5 mm. In calculating the zero-frequency effective dielectric constant of the disk structure, a w/h of 2.67, $(1/2)d/h$, was assumed; for the connecting lines, a w/h of 1.0 was used. Equation (5) gives effective permittivities of 10.76 and 11.94 for the connecting lines and disk, respectively.

Dispersion in microstrip increases the effective dielectric constant with increasing frequency. Schneider [7] has also studied the frequency dependence of the effective permittivity. In a modified form,

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The author is with Andersen Laboratories, Inc., Bloomfield, Conn. 06002.
[1] T. Miura and T. Hashimoto, in *1971 IEEE G-MTT Symp. Dig.*, pp. 80–81.