TABLE I
COMPARISON OF CALCULATED AND MEASURED CUTOFF FREQUENCIES

	f_c (GHz)		
	CALCULATED		
	WITHOUT GROOVES	WITH GROOVES	MEASURED
WITH SUBSTRATE	36 3	29 7	29 4
WITHOUT SUBSTRATE	39 3	33 4	33 5

approximation and Gardiol's LSM_{11} mode are about the same for this particular case.)

The measured values were taken as that frequency at which the channel had attenuated the signal 10 dB below the signal level without the test fixture. An additional margin of at least 10 percent should be allowed to avoid the lower signal levels of the waveguide mode below cutoff.

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Optimizing Wide-Band MIC Switch Performance

FULVIO G. ANANASSO

Abstract—**A detailed analysis is presented of a well-known shunt p-i-n diode configuration—tee low-pass filter—particularly useful in high frequency applications and easily realizable in a practical MIC structure. Behavior of both single and double θ-spaced low-pass cell is examined.**

In a microwave switch design, a significant problem is constituted by the mounting parasitics, mainly in the shunt p-i-n diode configuration. Bonding wires and ribbons connecting diode chips to microstrip-line board often exhibit a considerable (inductive) reactance, increasing insertion loss and decreasing isolation. To obviate these inconveniences, it is often worthwhile to incorporate the shunt p-i-n diode in a tee low-pass cell, as indicated in Fig. 1, where C holds for the off p-i-n capacitance, L for the connecting lead, and R_0 being the load resistors. This solution, easily realizable in MIC structure, allows reducing the insertion loss and improving the isolation performance.

The transmission coefficient of such a network is given by

$$\frac{1}{|S_{21}|^2}=(1-\omega^2LC)^2+\frac{\omega^2}{4R_0^2}\left[L(2-\omega^2LC)+CR_0^2\right]^2. \quad (1)$$

This filter response is a function of the quantity $X=L/CR_0^2$; particularly, $X=0.5$ calls for a maximally flat response, while $X<0.5$ curves have a monotonous shape. When X assumes a value higher than 0.5, there results a Tchebysheff response, whose

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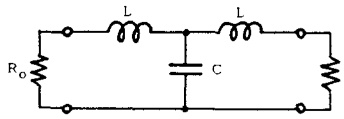

Fig. 1 TEE low-pass filter constituted by an OFF shunt p-i-n diode and two bonding wires

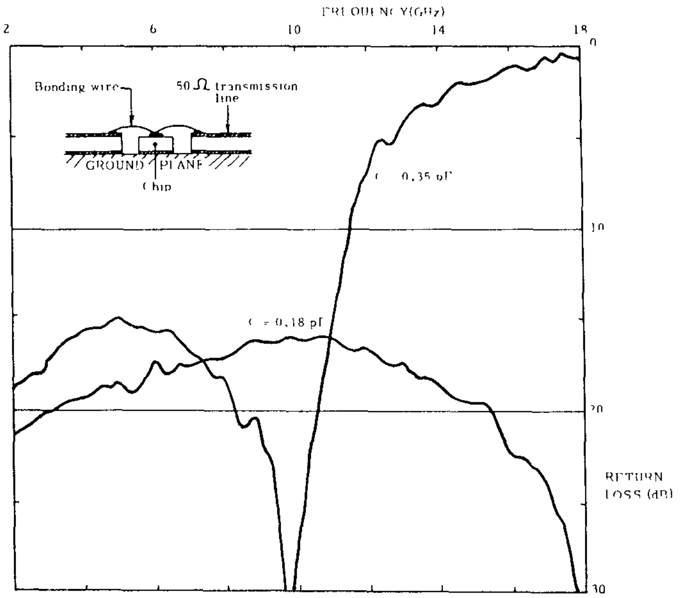

Fig 2. Return loss of a tee filter with $C=0.18$ pF and $C=0.35$ pF.

TABLE I

RIPPLE R(dB)	$g_1=g_3$	g_2	$X=L/CR_0^2$	$b=g_1R_0/2\pi$
0.01	0.6791	0.9702	0 65	5.00
0 05	0 8794	1.1132	0 79	7.00
0.10	1.0315	1.1474	0 90	8.20
0.15	1.1397	1.1546	0 99	9.10
0.20	1.2275	1.1525	1.07	9 80
0.25	1.3030	1.1461	1.14	10.40
0.30	1.3713	1.1378	1.21	10 95
0 35	1.4332	1.1283	1.27	11.40
0.40	1.4909	1.1180	1.33	11.85
0 45	1.5451	1.1074	1.40	12 35
0.50	1.5963	1.0967	1 46	12.75

(3) → f_c (GHz) = b/L(nH)

characteristics—ripple R, cutoff frequency $f_c, \cdots$—are dependent on $g_1/g_2=X$, g_1 and g_2 being the low-pass prototype filter ($N=3$) elements. We have

$$R(\text{dB})=10\log_{10}\frac{(X+1)^2(8X-1)}{27X^2} \quad (2)$$

$$f_c=g_2/2\pi R_0C=g_1R_0/2\pi L=b/L \quad (3)$$

$$f_z=[f]_{S_{21}=0}=\frac{R_0}{2\pi L}\sqrt{2X-1} \quad (4)$$

$$f_R=[f]_{S_{21}=10^{R/10}}=f_z/\sqrt{3}. \quad (5)$$

Values of g_1 and g_2 are indicated in Table I for various ripple values, together with X and $g_1R_0/2\pi$ (for a 50-Ω system). Fig. 2

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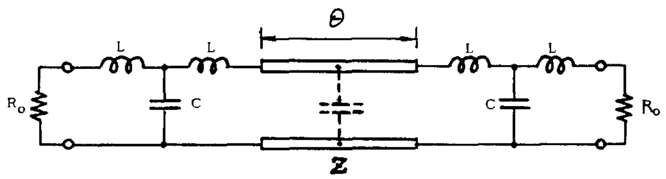

Fig. 3. Double θ-spaced tee cell configuration.

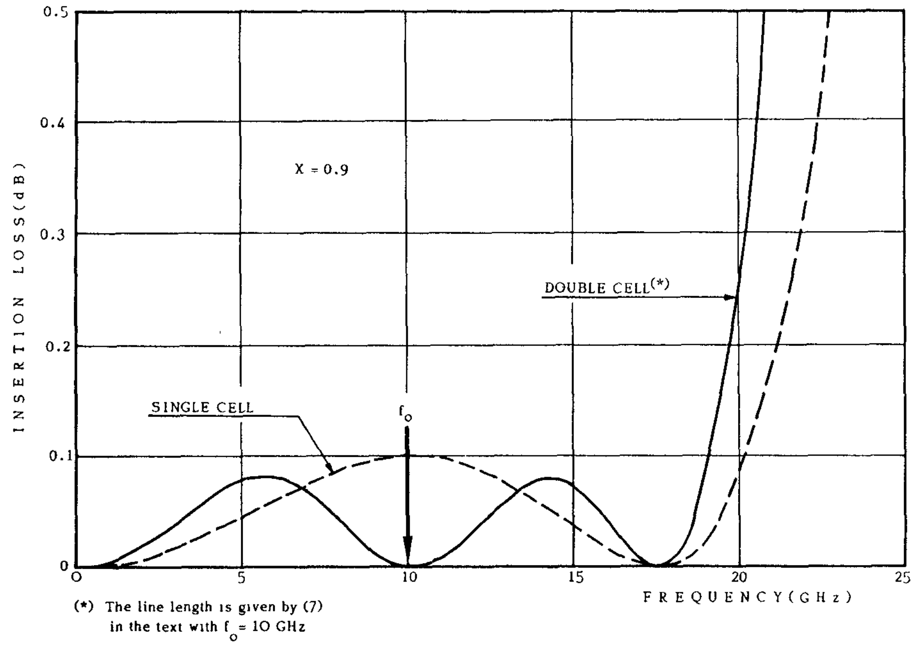

Fig. 4. Performance of both single and double cell ($X = 0.9$).

shows return loss of a tee low-pass mounting—it also contains a sketch of a microstrip-line board circuit—in which $C = 0.18$ pF and $C = 0.35$ pF, both yielding—with modified L values—$X = 0.9$ in a 50-Ω system.

Typical isolation obtainable from a tee low-pass network such as in Fig. 1 (diode ON, that is a low resistance in place of C capacitance) rarely exceeds 25–30 dB, depending on p-i-n characteristics and frequency. Thus, only one cell is often not sufficient to assure good electrical performances, and a multicell configuration ($N = 5, 7, \cdots$) has to be employed. A very practical approach is shown in Fig. 3, where two tee low-pass cells are separated by a convenient Z-Ω transmission line length. If Z is kept very low, carefully designing the spacing θ in order to realize the suitable capacitance value yields a very wide band, high isolation structure, constituted by a 7-element low-pass Tchebysheff filter. Alternatively, we can suitably choose the spacing θ in a way to place a zero of attenuation at a particular frequency. Putting $Z = R_0$, the transmission coefficient of Fig. 3 network results

$$\frac{16R_0^4}{|S_{21}|^2} = \omega^4\left(B - CR_0^2\right)^4 + \left[A^2 + \omega^2\left(B + CR_0^2\right)^2\right]^2 \cdot \left\{1 - \frac{2\omega^2\left(B - CR_0^2\right)^2}{A^2 + \omega^2\left(B + CR_0^2\right)^2}\cos 2\left[\theta - \tan^{-1}\frac{A}{\omega\left(B + CR_0^2\right)}\right]\right\} \quad (6)$$

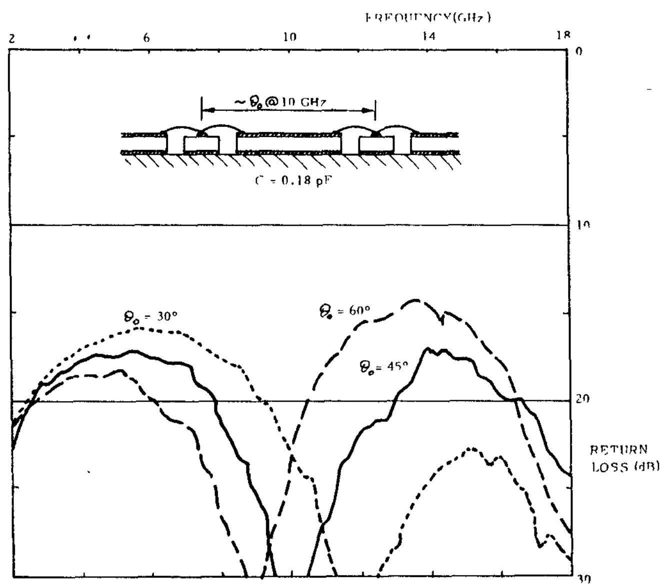

Fig. 5. Return loss of a double θ-spaced tee cell configuration with $C = 0.18$ pF and various line electrical lengths.

where

$$A = 2R_0(1 - \omega^2 LC)$$

and

$$B = L(2 - \omega^2 LC).$$

From (6) we can find the optimum θ_0 spacing value minimizing the insertion loss

$$\theta_0 = \tan^{-1} \frac{2R_0(1-\omega_0^2 LC)}{\omega_0[L(2-\omega_0^2 LC)+CR_0^2]}. \tag{7}$$

If (7) is applied, at $f = f_0$ a zero of attenuation occurs. A wide band of operation can therefore be used across f_0, if the elementary low-pass cell characteristics are suitably chosen. In fact, if the tee cell is individually well matched, the same will be true for the overall network. Thus, optimizing the initial tee low-pass filter in order to limit its maximum mismatching loss to 0.20–0.25 dB ($X \leq 1.15$), and choosing f_0 about coinciding with the frequency corresponding to maximum of ripple, i.e., to maximum mismatching (this frequency, see (5), is $f_z/\sqrt{3}$), the best performance can be obtained in a multioctave band (see Fig. 4). Fig. 5 shows the return loss of a double θ_0-spaced tee cell configuration, in the case $X = 0.9$. From (5), maximum mismatching occurs at $f = 10$ GHz. Choosing $f_0 = 10$ GHz, applying (7) yields $\theta_0 = 45°$ at 10 GHz. As we can see from the figure, when $\theta_0 = 45°$ the network is well matched at 10 GHz, and in the range 2–18 GHz, VSWR is less than 1.35 : 1. When θ_0 is different from 45°, VSWR increases. However, for $30° \leq \theta_0 \leq 60°$, we note a maximum VSWR of about 1.5 : 1. Outside this electrical length range we observe a relevant degradation of performance. Isolation typically results 40–50 dB. Higher values can be obtained increasing the number of θ-spaced tee cells.

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Letters

Correction to "Asymptotic High-Frequency Modes of Homogeneous Waveguide Structures with Impedance Boundaries"

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The treatment of Section V in the above paper[1] was incomplete and, as such, a bit misleading. In fact, the existence question for solutions of equation (53) for $f_2 = (\pi_2, m_2)$ did not properly take into account the degeneracy of the basic modes $f_1 = (\pi_1, m_1)$. It is known that for a solution to exist, the right-hand side of a deterministic equation like (53) must be orthogonal to all solutions of the homogeneous adjoint problem, which in this case is the basic problem with solutions f_1. Without degeneracy, equation (56) would be that condition. However, since there are at least two linearly independent solutions f_{1i}, there are at least two such conditions, which leads to a contradiction except if f_1 in (53) is chosen in a special way. Let us denote the admissible f_1 in (53) by f_1' and it can be written as a linear combination of any complete set of degenerate basic modes corresponding to the same parameter β_1: $f_1' = \Sigma \alpha_i f_{1i}$. The orthogonality condition reads

$$({}_1f_i^*, Lf_2) - (f_{1i}^*, Bf_2)_b = 2\beta_2(f_{1i}^*, f_1') - 2j\beta_1(f_{1i}^*, Mf_1')_b = 0 \tag{1}$$

and it should be satisfied for all $i = 1 \cdots n (n \geqq 2)$. This is an algebraic equation for the matrix $\boldsymbol{\alpha} = (\alpha_i)$

$$j\beta_1 \boldsymbol{M} \cdot \boldsymbol{\alpha} = \beta_2 \boldsymbol{F} \cdot \boldsymbol{\alpha} \tag{2}$$

where we denote $\boldsymbol{M} = (f_{1i}^*, Mf_{1j})$, $\boldsymbol{F} = (f_{1i}^*, f_{1j})$. The unknown coefficient β_2 is obtained as a solution of the eigenvalue equation

$$\det(j\beta_1 \boldsymbol{M} - \beta_2 \boldsymbol{F}) = 0 \tag{3}$$

and the corresponding admissible right-hand side of a (53) from the eigenvectors $\boldsymbol{\alpha}$ of (2).

The expression (56) giving β_2 in terms of f_1 is not incorrect but ambiguous, because if we would substitute f_1' for f_1 we would obtain the correct β_2. Hence, the general conclusions following (56) in Section V are valid if f_1' is understood everywhere in place of the ambiguous f_1.

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The author wishes to thank Dr. C. Dragone for a discussion on the above problem and providing with a copy of his recent paper [1] on the same subject, where a slightly different technique is applied.

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