Short Papers

A More Accurate Model of the TE_{10} Type Waveguide Mode in Suspended Substrate

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***Abstract*—An analytical model of the suspended substrate transmission line which includes the grooves used to support the substrate is presented. The model predicts a lower cutoff frequency of the TE_{10} type waveguide mode than that without the grooves. The analysis is confirmed by experiments.**

A simplified model of the suspended substrate transmission line is employed by both Schneider and Gardiol [1], [2] to calculate the cutoff frequency of the TE_{10} type waveguide mode in the channel. The approximation omits the grooves which suspend the substrate in the channel. For geometries in which the groove depth is a significant fraction of the channel width (often the case for millimeter structures), this omission predicts a much higher cutoff frequency for the TE_{10} type mode than actually exists. A better approximation for this case employs a transverse resonance analysis of the actual structure as done by Cohn for ridged waveguide [3]. As shown in Fig. 1, a parallel plate waveguide sees an open-circuit looking at distance l_1 to the left of junction J of Fig. 1(b) and sees a short-circuit at distance l_2 to the right. At the cutoff frequency for the TE_{10} type mode (resonance), neglecting the susceptance of the discontinuity,

$$Z_2 \tan\phi_1 - Z_2 \cot\phi_1 = 0 \qquad (1)$$

where

$$Z_1 = K\frac{b_1}{\sqrt{\epsilon_1}}$$

$$Z_2 = K\frac{b_2}{\sqrt{\epsilon_2}}$$

$$\phi_1 = \frac{2\pi l_1\sqrt{\epsilon_1}}{\lambda_c}$$

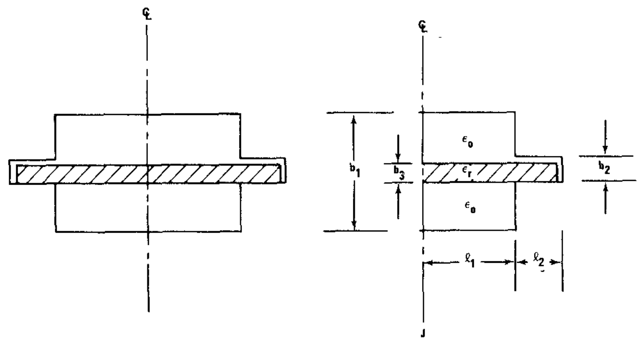
$$\phi_2 = \frac{2\pi l_2\sqrt{\epsilon_2}}{\lambda_c}$$

$$\epsilon_1 = \left[1 - \frac{b_3}{b_1}\left(\frac{\epsilon_r - 1}{\epsilon_r}\right)\right]^{-2}$$

$$\epsilon_2 = \left[1 - \frac{b_3}{b_2}\left(\frac{\epsilon_r - 1}{\epsilon_r}\right)\right]^{-1}$$

$$K = \text{constant}.$$

Thus the cutoff frequency is determined by finding the frequency



(a) (b)

Fig. 1. (a) Suspended substrate transmission line with grooves to support substrate. (b) Parallel plate model.

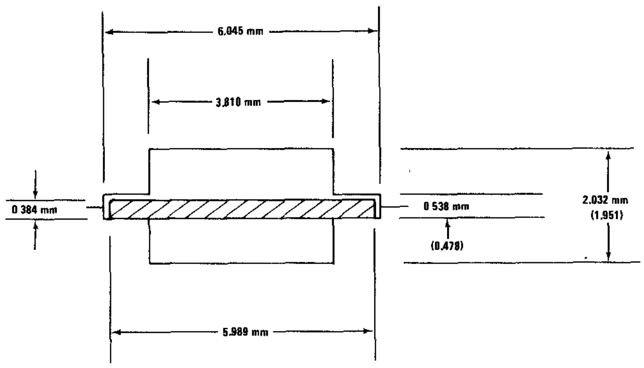


Fig. 2. Dimensions of suspended substrate transmission-line test fixture (numbers in parentheses are for the measurement made without the substrate).

which satisfies the relationship

$$\tan\phi_1 \tan\phi_2 = \frac{b_1}{b_2}\left(\frac{\epsilon_2}{\epsilon_1}\right)^{1/2} \qquad (2)$$

Note that the relationships for ϵ_1 and ϵ_2 are quasi-static approximations which are sufficiently accurate for this calculation.

To confirm the validity of this model, a suspended substrate test fixture with the dimensions given in Fig. 2 was constructed using a blank fused silica substrate for which ϵ_r is 3.78. A 0.06-mm indium gasket was placed between the two halves of the test fixture to prevent cracking the substrate and to insure a good electrical boundary. The cutoff frequency was measured by connecting the channel on each end directly to K_a-band waveguide and sweeping across the band of interest. To further verify the model, an additional measurement was made without the substrate. A gasket was not considered necessary in this case which accounts for two different height dimensions in Fig. 2. The calculated and measured results are compared against the model without grooves in Table I. (The results given by both Schneider's

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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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[2] F. E. Gardiol, "Careful MIC design prevents waveguide modes," *Microwaves*, pp 188–191, May 1977.
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Optimizing Wide-Band MIC Switch Performance

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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

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

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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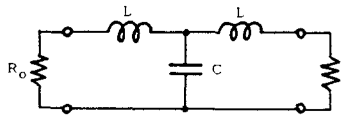

Fig. 1 TEE low-pass filter constituted by an OFF shunt p-i-n diode and two bonding wires

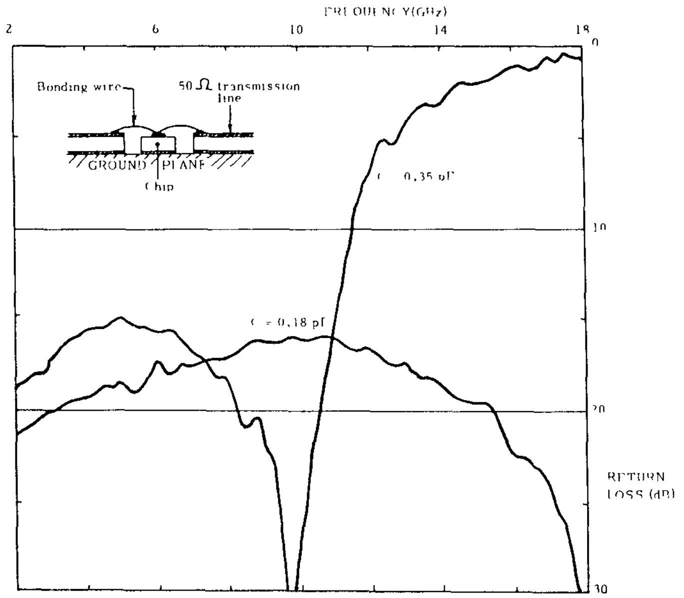

Fig 2. Return loss of a tee filter with $C = 0.18$ pF and $C = 0.35$ pF.

TABLE I

RIPPLE R(dB)	$g_1 = g_3$	g_2	$X = L/CR_0^2$	$b = g_1 R_0 / 2\pi$
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

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

$$f_c = g_2/2\pi R_0 C = g_1 R_0/2\pi L = b/L \quad (3)$$

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

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

Values of g_1 and g_2 are indicated in Table I for various ripple values, together with X and $g_1 R_0/2\pi$ (for a 50-Ω system). Fig. 2