Short Papers

A Mathematical Model of a Varactor Package in Microstrip Line

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Abstract—**Modeling of special configurations in microstrip becomes very important where computer-aided design is to be used. In order to develop the full capacity of the Gunn diode, the circuit must be understood. The following presentation concerns development of a model for a diode package mounted in a shunt configuration so that impedances can be referred to the semiconductor chip.**

Introduction

Modeling of special configurations in microstrip becomes very important where computer-aided design is to be used. The following example concerns development of a model diode package mounted in shunt configuration in a microstrip line so that impedances can be referred to a semiconductor chip. In order to develop the full capacity of the Gunn diode, the circuit must be understood [1]–[4].

In this presentation a lumped constant equivalent circuit, which can simulate microwave circuit behavior, is given. Since the immittances of the circuit are distributed and are frequency dependent, the equivalent circuit can only approximately describe the resonant frequencies measured. The problem posed is that of developing an equivalent circuit (a mathematical model) for the diode package so that the model will predict the measured parallel resonant frequencies.

Experimental Procedure and Results

An open-circuited straight-line half wavelength resonator using a conventional 99.5-percent pure 0.025-in-thick alumina substrate was driven externally by swept-frequency technique through a loosely coupled probe.

The probe was placed in the position shown in Fig. 1, initially with a coupling gap of 0.001 in. The gap was increased by etching until the parallel resonant frequencies monitored with the Hewlett-Packard Network Analyzer were insensitive to the probe's location. This technique ensures that the probe is not seriously perturbing the field of the resonator.

The resonant frequency assuming a TEM mode can be theoretically calculated:

$$f_n = \frac{nv}{2L\sqrt{\epsilon_{\mathrm{eff}}}} \tag{1}$$

where n is the number of half wavelengths in the direction of the length of the resonator, v is the velocity of light in the vacuum, L is the cavity length which is 2.105 cm, and $\sqrt{\epsilon_{\mathrm{eff}}}$ is the effective dielectric constant which is equal to 2.62 at zero frequency.

Due to dispersive effects the effective dielectric constant increases with increasing frequency. This was accounted for in calculating (1) and is given by Itoh and Mittra [5]. Both the resonant frequencies measured and calculated are tabulated in Table I(a).

A standard empty (less ribbon leads) varactor package, as shown in Fig. 2, was mounted in the center of the resonator. It has been assumed that the empty package can be represented by a lumped-element low-pass pi-network of the form indicated in Fig. 1(b). Placing the diode in shunt with the resonator resulted in the resonances undergoing a frequency shift. The frequency shift in resonance is caused by the addition of L_B, the inductance due to the current path through the surface of the top hat, and the capacitance C_A is due to the abrupt step from the microstrip surface down into the varactor package. Separation of the circuit of the package from the external microstrip circuit is defined to be at the outer diameter of the top hat as shown in Fig. 1(b).

Manuscript received April 24, 1972; revised January 12, 1973. This work was conducted in the former Microwave Products Group of Monsanto, St. Louis, Mo., now a part of Microwave Associates, Burlington, Mass.

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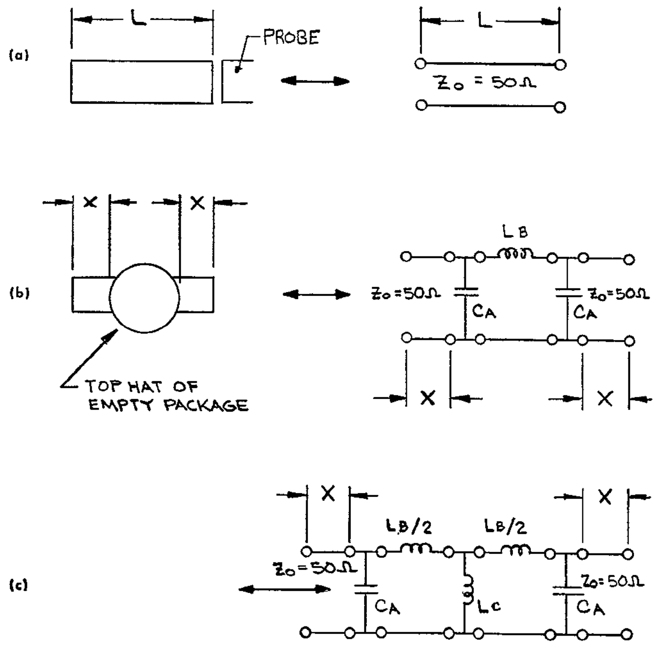

Fig. 1. Top view of microstrip circuit and equivalent circuit. (a) $L=0.797$ in. (b) $x=0.336$ in. (c) Same as (b), except package contains ribbon leads.

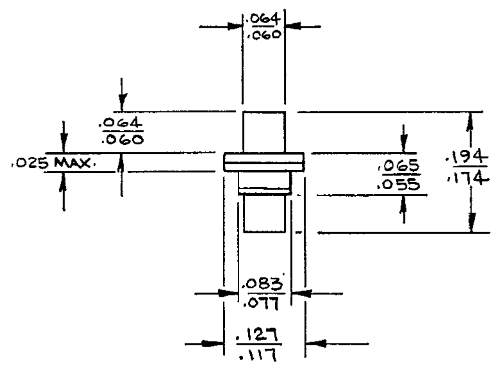

Fig. 2. Varactor package.

TABLE I

(a) Resonator Only					
f_{meas} (GHz)	2.768	5.160	7.870	10.264	
f_{calc} (GHz)	2.697	5.352	7.961	10.530	
(b) Open-Circuit Package					
f_{meas} (GHz)	2.91	4.76	8.35	9.90	
f_{calc} (GHz)	2.96	4.70	8.30	10.05	
Error functions (GHz)	0.05	−0.06	−0.05	0.150	
(c) Short-Circuit Package					
f_{meas} (GHz)	2.412	2.923	6.468	8.415	10.595
f_{calc} (GHz)	2.600	2.950	6.500	8.250	10.450
Error functions (GHz)	0.188	0.027	0.032	−0.165	−0.145

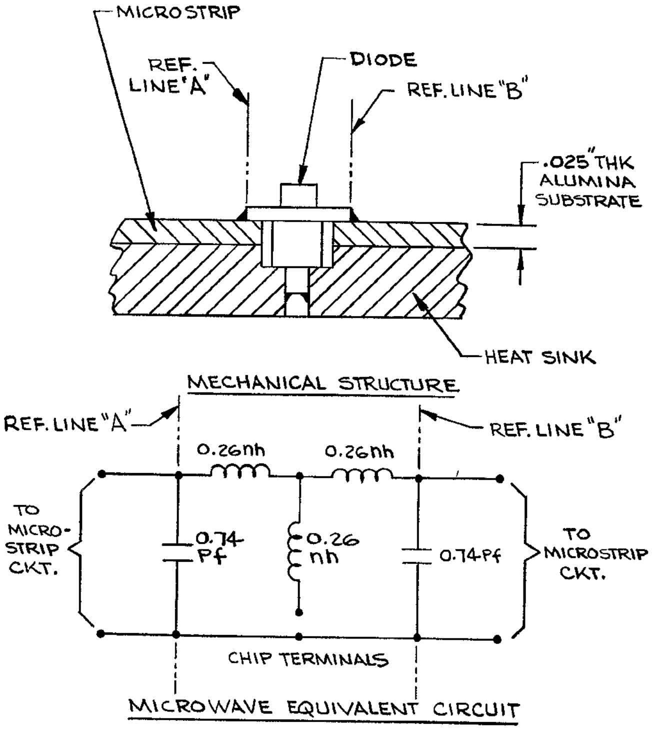

Fig. 3.

The matrices for the low-pass pi-network for the empty package can be represented by the matrix multiplication of the cascaded unit cells

$$T_\pi = \begin{vmatrix} 1 & 0 \\ jB_a & 1 \end{vmatrix} \begin{vmatrix} 1 & \dfrac{1}{jB_b} \\ 0 & 1 \end{vmatrix} \begin{vmatrix} 1 & 0 \\ jB_a & 1 \end{vmatrix} = \begin{vmatrix} 1 + B_a/B_b & -j/B_b \\ jB_a(2 + B_a/B_b) & 1 + B_a/B_b \end{vmatrix} \tag{2}$$

where B_a and B_b are the susceptances of C_A and L_B, respectively.

The matrix for the transmission line can be represented by

$$T_L = \begin{vmatrix} \cos\theta & jz_0 \sin\theta \\ jY_0 \sin\theta & \cos\theta \end{vmatrix} \tag{3}$$

where

$$\theta = 2\pi \frac{x}{\lambda_m}, \qquad x = 0.336 \text{ in}$$

$$Y_0 = 1/z_0, \qquad z_0 = 50\ \Omega.$$

By repeated application of matrix multiplication, the equivalent circuit for the empty package and the long resonator can be represented by the $ABCD$ matrix

$$|T_L|\ |T_\pi|\ |T_L| = \begin{vmatrix} A & B \\ C & D \end{vmatrix}. \tag{4}$$

The input admittance is

$$Y = \frac{AG_L + B}{CG_L + D} \tag{5}$$

where G_L is the load conductance of the open-circuited transmission line and was taken to be 0.001 mho. When $Y = 0$, parallel resonance is achieved.

A digital computer using the Monsanto Company's Microwave Circuit Analysis program was used to determine the element values for the diode package. The procedure consists of selecting values for C_A and L_B, then calculating the $ABCD$ matrix, equation (4), and then solving for the admittance Y equal to zero. An error function was used to minimize the frequency differences between the measured and the calculated resonant frequencies. A subroutine prints out the calculated resonant frequencies and the error function. The optimized values for L_B and C_A were 0.52 nH and 0.74 pF, respectively. The experimental and computed circuit model resonances are tabulated in Table I (b).

The empty package was replaced with an identical package in which a 6-mil ribbon lead was connected from the top hat to the pedestal at the base of the package. With the addition of the inductance L_C, the resonances again shifted frequency and a new resonant frequency occured. A matrix similar to the open-circuited package case with the same values for L_B and C_A was used to obtain the optimum lead inductance L_C. The computed value was 0.26 nH. Both the experimental and computed resonances are shown in Table I (c).

In conclusion, a very simple lumped circuit gave fairly good agreement. Better results could have been achieved if the parameters of the pi-network were not invariant with change in measured frequency. The transformation of impedances to the active chip terminals (Fig. 3) now becomes a matter of ordinary RLC network theory.

References

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The Impedance and Scattering Properties of a Perfectly Conducting Strip Above a Plane Surface-Wave System

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***Abstract*—The impedance and scattering properties of a perfectly conducting strip above a dielectric-coated conducting plane is investigated both theoretically and experimentally. An integral equation for the induced current is presented and solved numerically using a point-matching technique. The values of the reflection and transmission coefficients are calculated from the computed current distributions. The results of the computations are compared to the measured values and the agreement is quite good. In addition the impedance and fractions of power reflected, radiated, and transmitted are computed and displayed graphically.**

I. Introduction

Whenever an obstacle is placed in the vicinity of an unshielded surface-wave system, part of the scattered field radiates away from the surface, part is backscattered in the form of a surface wave, and the remainder is transmitted in the forward direction also as a surface wave. The obstacle can be thought of as an antenna fed by a surface-wave transmission line [1], [2]. It is, therefore, altogether appropriate to characterize the obstacle by the usual parameters from transmission-line theory; namely, by impedance or scattering matrices. An aspect of this type of scattering problem which has been largely neglected in the literature is that of the prediction of the fraction of power lost by radiation.

Gillespie and Gustincic [3], [4] have computed the reflection coefficients of strips above a dielectric-coated conducting plane and of plane conducting annuli surrounding a Goubau line. Using results presented in [4], Gillespie [5] calculated the shunt impedance of plane annuli on a Goubau line, as well as the fractions of power radiated, reflected, and transmitted. It is the purpose of this short paper

Manuscript received May 30, 1972; revised January 26, 1973.
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