DETERMINATION OF SEMICONDUCTOR JUNCTION DEVICE PACKAGE NETWORKS*

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Abstract

A general method for the determination of a linear, lumped-element equivalent circuit of any semiconductor device package in any microwave circuit mount is presented. Results are given for IMPATT and TRAPATT devices in coaxial and microstrip circuits, together with experimental validation of the results.

I. Introduction

Electrical properties of microwave semiconductor devices are effected by immittance transforming properties of the packages utilized to permit their handling, mounting, and heat sinking. These effects increase in significance with higher frequencies, where the electrical dimensions of the packages may become significant fractions of the wavelength. Determination of lumped element equivalent circuits to represent this transformation permits correlation of theory and experiment in device modeling, and in analysis and design of microwave circuits utilizing such devices. Several authors [1-6] have described efforts related to specific package styles in specific mounting configurations and circuit media. Exact field solutions are usually prohibited because of complicated geometries, and most results have relied on both geometrical considerations and impedance measurements, using specially constructed probes and reference packages with several known internal terminations replacing the device.

The purpose of this paper is to present a general method which uses the HP Automatic Network Analyzer and a digital computer for determining passive, lumped, linear networks to characterize any combination of package and microwave circuit mount, and to present results for an IMPATT device in both coaxial and microstrip circuits at X-band, and for a TRAPATT device in microstrip at S-band. It is noteworthy that the required impedance measurements are taken with the device in situ, without requiring special references such as open or shorted packages, etc.

II. Analytical Basis for the Method

Getsinger[1] points out that the problem of establishing a quantitatively meaningful microwave equivalent circuit for a packaged and mounted diode requires the establishment of transmission line modes and terminal planes. Semiconductor junctions under reverse bias can be modeled by a series combination of a depletion layer capacitance C_j and a loss R_s which accounts for the resistance of the package lead, ohmic contacts and the non-depleted region of the semiconductor substrate. Both of these components are a function of the reverse bias voltage. The value of the capacitance which can be measured at low frequency is assumed to be invariant up through the microwave frequencies of interest. The choice of an inaccessible reference plane at the semiconductor junction results in a series RC termination connected to those terminals. The accessible part of the network is any convenient reference plane for the dominant mode in the transmission medium.

Figure 1 shows the resulting model. N represents a lossless immittance transformation which characterizes both the parasitics of the package and the transformation from the dominant mode in the transmission medium to a longitudinal electric field across the semiconductor junction depletion layer with its associated displacement and/or conduction currents. Because the loss R_s is normally quite small relative to the characteristic impedance of practical microwave networks, the reactive part of the impedance measured at terminals AA' will be similar to that of a lossless circuit terminated in C_j only. A minimum element realization of a lossless driving point reactance may be determined, except for a multiplicative constant, from the location of its poles and zeroes, which alternate and lie on the $j\omega$ axis [7]. The shape of such a curve will then suggest $j\omega$ axis pole-zero locations, from which an L-C ladder may be synthesized. Measured reactance is shown in Figure 2 for a typical case along with the postulated pole-zero location and the resulting lossless approximation. An impedance function is constructed from the poles and zeroes of the form

$$Z(p) = \frac{H(p^2+\omega_{01}^2)\ (p^2+\omega_{03}^2)}{p(p^2+\omega_{p1}^2)} \tag{1}$$

where ω_{0i} and ω_{pi} are the zeroes and poles of the lossless approximation and H is a multiplicative constant. Expansion of this function into an LC ladder is straightforward and is used for an initial guess for element values. In order to satisfy the measured real part of the impedance, a small series resistance R_s is assumed. The end result, which forms the configuration whose element values are to be optimized by the computer algorithm, is shown in Figure 3.

Initial values as described above are perturbed by means of an optimization algorithm using a Fletcher-Powell technique to minimize the mean square error between measured impedances and those calculated from the equivalent circuit. Rapid convergence is experienced as a result of the nearly correct element values predicted by equation (1). The method has been employed for measuring small and large area devices in coaxial and microstrip circuits on dielectrics. Results are described in the last section.

III. Experimental Procedure

The measurement procedure incorporates an automatic network analyzer and yields precise, accurate data suitable for determining equivalent circuit parameters of the semiconductor chip and diode package. For this, reflection coefficients near unity must be measured accurately and repeatably. With an automatic network analyzer, a reference with reflection coefficient of unity magnitude and phase angle equal to that of the unknown is desired for system calibration at each frequency. Standard calibration programs incorporate two reference shorts approximately 180° apart at each frequency. Due to phase-sensitive components in the analyzer, accurate determination of the unity-magnitude reflection coefficient Smith Chart circle is not assured, and small errors in the magnitude of Γ caused by measurements on loads with phase angles other than the calibration angle result in significant errors and scatter in the measured real part of the load impedance.

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A technique was developed to correct these errors in the reflection coefficient magnitude. The reflection coefficient of each calibration short is measured over the entire frequency range. Assuming that each short has a unity magnitude reflection coefficient, a correction to the load reflection coefficient is determined from the two reference shorts which have phase angles approximately the same as the load. This significantly reduces scatter in the real part of the measured load impedance and improves the resulting equivalent circuit characterization.

IV. Experimental Results

An X-band IMPATT diode was characterized as described above in a 7mm 50Ω coaxial test fixture. The TEM reference plane was taken at the interface between the diode package flange and the coaxial center conductor. Figures 4 and 5 show measured and calculated real and imaginary parts of the impedance of the diode biased just below breakdown. The solid curves indicate behavior calculated from the resultant lumped equivalent circuit shown in the inset of Figure 5 after the initial element values were optimized by the computer. Verification of the equivalent circuit over the measured frequency band is given by the fact that element values obtained for other bias conditions remained invariant except for changes in R_s and C_j to account for the bias change. Element values for these bias conditions are shown in Table I(a).

The method was applied to a TRAPATT diode shunting the open end of a 20Ω microstrip line on a .031" thick teflon-polyglass substrate ($\varepsilon_r \sim 2.2$). Impedance measurements were taken from 2.0 to 12.4 GHz using two different transitions from the 50Ω measurement system to the 20Ω medium in order to obtain a low VSWR transformation over the entire band. A reference plane for the measurement was established by means of an indium-covered shorting block at the position of the diode, and a correction for the dispersion and loss of the microstrip line was obtained at each frequency from the measured impedance of the reference shorting block. The equivalent circuit derived from the measurement is similar to Figure 3 except with L_1 equal to zero. Element values for the resulting equivalent circuit are listed in Table I(b). The validity of the circuit was examined with a probe consisting of an identical threaded pill package with a type ARRM subminiature 50Ω coaxial line leading through the base to the position of the diode chip, and connected to the package flange by a gold bonding wire similar to the diode lead wire. Figure 6 shows measured reactance seen by the probe, together with reactance at the chip location calculated from the equivalent circuit of Table I(b), for a reference load consisting of a short located 1.400" from the package centerline in 20Ωmicrostrip. The agreement is good over this range of frequencies.

The third case consisted of an IMPATT diode shunting the open end of a .750" length of 50Ω microstrip line of .025" thick alumina ($\varepsilon_r \sim 9.6$) substrate. The same measurement technique was utilized as described above with the 20Ω($\varepsilon_r \sim 2.3$) microstrip circuit. With the diode biased near breakdown, the element configuration found for this case was the same as for the TRAPATT diode but with values given in Table I(c).

V. Conclusions

Measurements of impedance versus frequency for a semiconductor junction device in a microwave transmission line may be interpreted to provide the configuration and approximate values of a passive linear lumped element network which characterizes the immittance transformation associated with device package and mount. A digital computer algorithm has been developed which refines the approximations to provide useful design values for these parasitic reactances. Package transformations for GaAs IMPATT diodes in coaxial and microstrip circuits, and a Si TRAPATT diode in a microstrip circuit have been determined. The method is general and useful for any junction device in any package or mounting configuration.

Acknowledgement

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Table I

Equivalent Circuit Element Values

a. GaAs IMPATT Diode in Coaxial Cavity

Bias Voltage	R_s	L_1	C_1	L_2	C_2	C_j
0	.625	.201	.032	.518	.189	5.10
10	.584	.201	.032	.518	.189	1.70
40	.479	.201	.032	.518	.189	1.26

b. Si TRAPATT Diode in 20Ω Microstrip Circuit

Bias Voltage	R_s	L_2	C_1	C_2	C_j	L_1
100 V	1.08	.454	.265	.443	1.7	0.

c. GaAs IMPATT Diode in 50Ω Microstrip Circuit

Bias Voltage	R_s	L_2	C_1	C_2	C_j	L_1
47.5 V	1.3	.330	1.48	.06	.85	0.

(Resistances in ohms, capacitances in picofarads, inductances in nanohenries)

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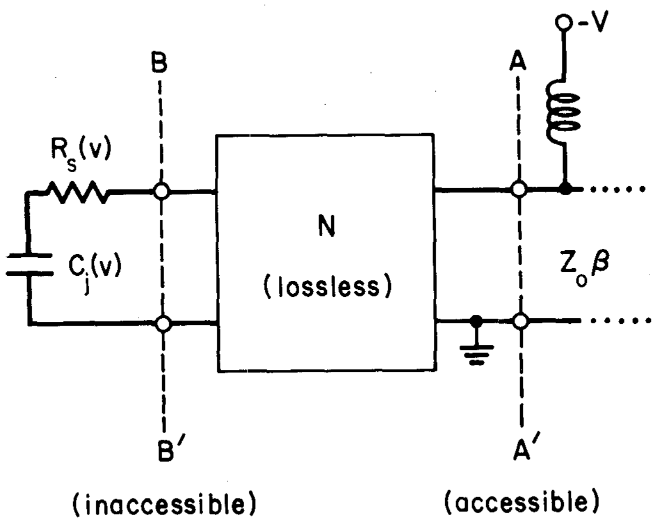

Figure 1. Model of packaged semiconductor junction device mounted in a microwave circuit.

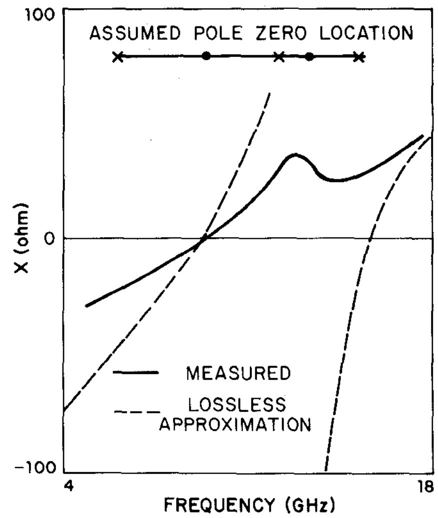

Figure 2. Reactance versus frequency for an X-band silicon IMPATT device.

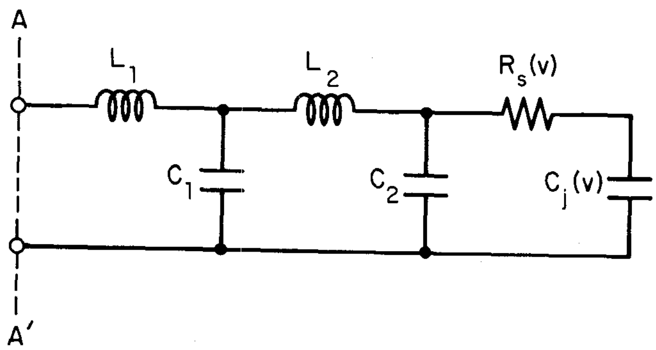

Figure 3. Lumped element equivalent circuit for realizing reactance behavior of Fig. 2.

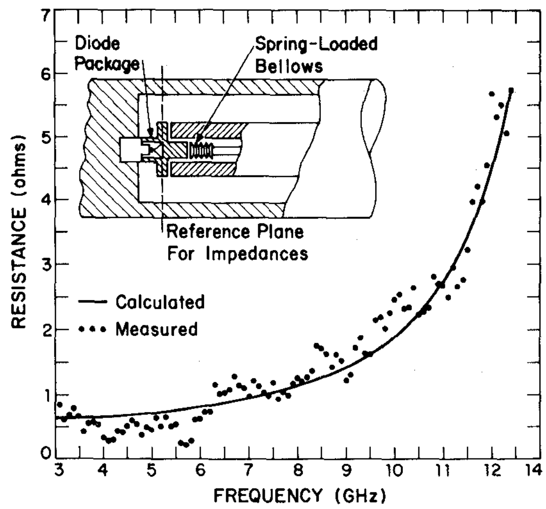

Figure 4. Resistance versus frequency for X-band Si IMPATT diode in coaxial mount (V = 40V).

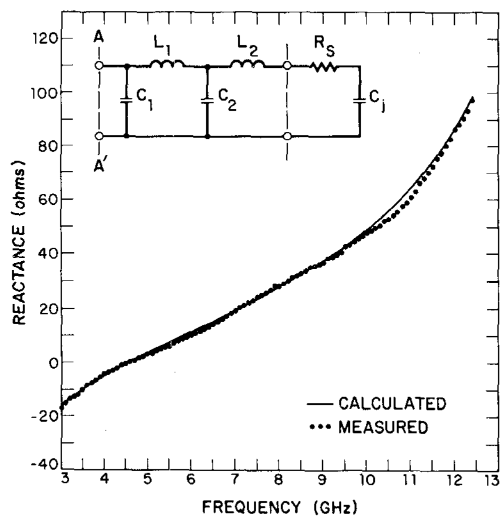

Figure 5. Reactance versus frequency for X-band Si IMPATT diode in coaxial mount (V - 40V).

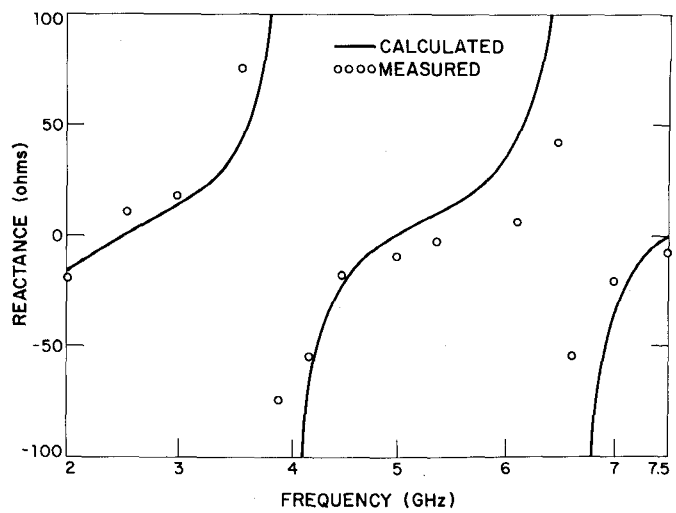

Figure 6. Calculated and measured reference impedances transformed through threaded pill package in microstrip.