TABLE I

Approx. Height	$\omega_0/2\pi$	C_D+C (pF)		L (pH)		G_L (mS)		G_D+G_C (mS)		$G_L+(G_D+G_C)$ (mS)		ρ_0		Q_L	
of Resonant Cap (mm)	(GHz)	SC	ML	SC	ML	SC	ML	SC	ML	SC	ML	SC	ML	SC	ML
1	9.54	3.8	4.1	73	68	2.0	4.6	1.7	2.1	3.7	6.7	0.007	0.13	61	36
2	9.50	3.2	3.2	89	89	1.7	3.7	1.3	1.7	3.0	5.4	0.021	0.14	64	36
3	9.45	2.7	2.7	105	105	1.5	2.9	1.0	1.3	2.5	4.2	0.036	0.15	64	38
4	9.28	1.6	1.4	179	206	1.3	2.7	0.7	1.0	2.0	3.7	0.096	0.21	46	22

was bias dependent. In practice, measurements satisfied this criterion if these spurious resonances were separated from the required resonance over a frequency range equal to several times $\Delta\omega$ or $\Delta\omega_B$. This range had to include the oscillation frequencies of the IMPATT's when biased above breakdown otherwise the derived circuit values would be irrelevant to the operation of the IMPATT diode when it was oscillating. The difference of oscillation frequency and the absorption frequency below breakdown was small enough to easily satisfy this criterion in our measurements, because our circuits had a large dominance of the equivalent circuit susceptance over the variable part of the diode's dynamic susceptance. Several types of cavity were used and results are shown for a waveguide cap circuit [5], [6] in Table I. Care has to be taken with this circuit because there is a resonance between the diode and the circuitry associated with the mounting post and radial cap and a further resonance in the waveguide between the sliding short circuit and the post. The results in Table I were taken for two conditions which were singly resonant. In one (marked SC) the short-circuit was an odd integral number of quarter wavelengths from the post so that it reflected a high impedance at the post plane. This was necessary both to avoid interaction between the two resonances and to give a known circuit condition for comparison with the next case. The positioning of the short circuit did not appreciably alter the results for small movements. In the second condition (marked ML) the sliding short circuit was replaced by a matched load. The objective of this was to increase the effective G_L by a factor of 2 because of the effective source conductance at the post terminals is half the characteristic conductance in this latter case. This was done because we had no other means of independently checking the measurements. It can be seen that this behavior is shown in Table I for various heights of the radial cap. A further point to note is the comparability of the useful load conductance G_L and internal loss conductance $G_D + G_C$. The ratio of these two is the ratio of the useful and lost power in the active oscillator. When the diode is operating as an oscillator the modulus of its negative conductance G_N must be equal to the sum of the positive conductances $G_L + (G_D + G_C)$. From (5) and (6),

$$G_N = G_L + (G_D + G_C) = \frac{(1 - n\rho_0)^{1/2}}{[\rho_0(n-1)]^{1/2}} \cdot \Delta\omega \cdot (C_D + C). \quad (9)$$

It may be argued that G_D is not the loss conductance present in the diode when it is biased into useful operation so invalidating the equality with G_N in (9). However, many diodes we have tested have an absorption bandwidth $\Delta\omega$ which becomes substantially independent of bias voltage just below breakdown implying that the residual losses are not caused by undepleted active regions of the diode and may still be present at the higher bias voltages of useful operation.

Once the cavity equivalent circuit parameters have been obtained it is possible to gain further information about the diode susceptance when it is oscillating or amplifying. An example is shown in Fig. 3 where the bias current induced variation of the diode susceptance ΔB_S at constant load impedance has been obtained from the bias dependence of the oscillation frequency $\Delta\omega_S$ where

$$\Delta B_S = -2(C_D + C) \cdot \Delta\omega_S.$$

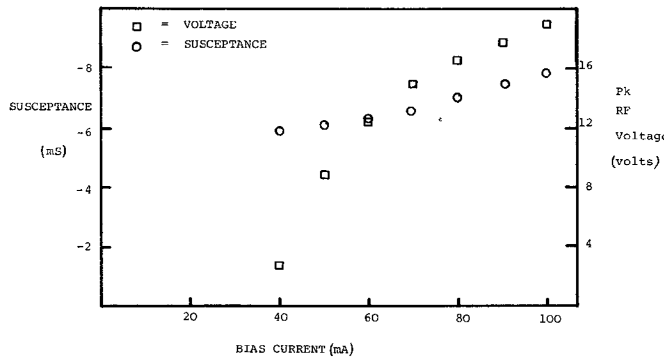

Fig. 3. Bias current variation of diode susceptance and RF voltage at $G_N = 2.7$ mS, $\omega_0 = 2\pi \times 10^{10}$ rad/s^{-1} in a coaxial cavity.

In conclusion we have developed a simple technique to characterize the diode and circuit of an IMPATT oscillator without mechanically interferring with the diode or its circuit. With some reduction of convenience and accuracy, we have also found the technique useful in characterizing the bandwidth-limiting and loss characteristics of circuits used for transferred-electron reflection amplifiers by substituting an IMPATT diode in the appropriate circuit.

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Very Large Impedance Steps in Microstrip

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***Abstract*—Experimental determination of an equivalent circuit for very large impedance steps in microstrip is described. The equivalent circuit is shown to be valid in the frequency range from 1 to 2 GHz on 0.0635-cm-thick alumina substrate, although the experi-**

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mental method used is applicable to any frequency range and any geometry.

I. INTRODUCTION

In recent years, several authors have discussed the effects of relatively small impedance steps in microstrip and balanced stripline [1]–[6]. However, during a development program for high power microwave transistor amplifiers, it became apparent that accurate computer optimization of microstrip matching networks could not be accomplished without accounting for the effects of very large impedance steps. At the suggestion of Dr. L. Young of the Naval Research Laboratory, Washington, D.C., a short length of lossy microstrip transmission line was selected as a possible representation of the junction discontinuity effect. The results obtained indicate that this equivalent circuit models the junction well over at least one octave bandwidth.

II. EXPERIMENTAL PROCEDURE

Several circuits were assembled, similar to that shown in Fig. 1(a), with the low Z_0 line having a different impedance value in each circuit; the length of the low Z_0 line was kept approximately constant at 3.81 cm. The input impedance at a reference point was measured for each circuit, from 1 to 2 GHz, using a Hewlett-Packard 8410A manual network analyzer. The measurements were computer corrected to account for inherent equipment errors [7].

III. EQUIVALENT CIRCUIT MODEL

The junction effect model, shown in Fig. 1(b), was used with a network optimization computer program, to establish Z_0, l, and α for the "junction" lines. The network optimization program was allowed to vary only the parameters of the hypothetical transmission lines used to represent the junction effects. The parameters thus obtained are those for which the input impedance to the hypothetical network was the same as that measured for the actual network. This particular equivalent circuit representation was chosen because it intuitively corresponds well to physical reality, and because it is analytically and computationally simple to handle.

IV. RESULTS

The results are shown in Fig. 2. For example, consider a point in a physical network at which there is a dimensional step of $(D/d) = 12$. The electrical analysis of this portion of the network is then correct if a section of line is assumed to exist between the two lines forming the step, whose parameters are given in Fig. 2; for $(D/d) = 12$, they are $Z_0 = 0.39 \times Z_{\text{high}}$, $l = 0.083$ cm, and $\alpha = 1.55$ dB/cm. The reason for the necessity of a loss parameter α is not entirely clear; however, it is presumed to account for radiation losses, and is required for accurate results.

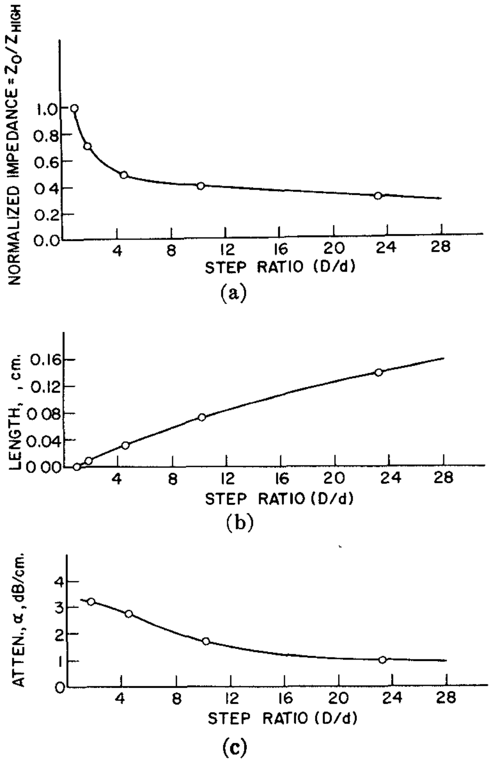

Fig. 2. Transmission line junction model parameters versus dimensional step ratio. (a) Normalized impedance. (b) Length. (c) Attenuation factor.

V. DISCUSSION

The curves of Fig. 3 show the effectiveness of the use of this step representation information. The measured input impedance curves for a microstrip network are shown, along with computed input impedance with and without accounting for the large step discontinuities. It is seen that the computations without accounting for

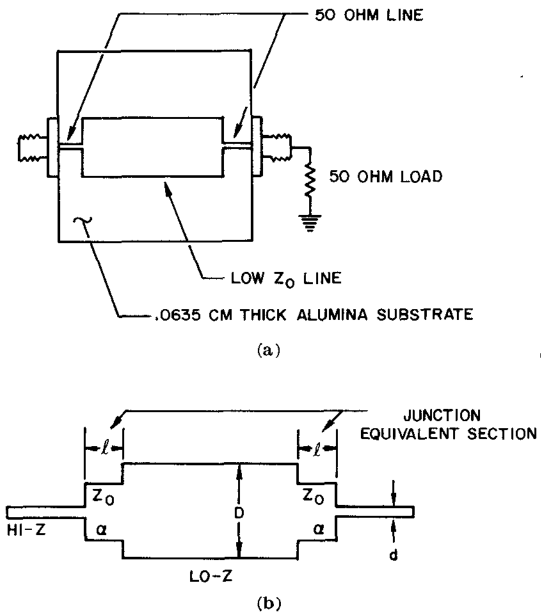

Fig. 1. (a) Experimental hardware. (b) Transmission line junction model.

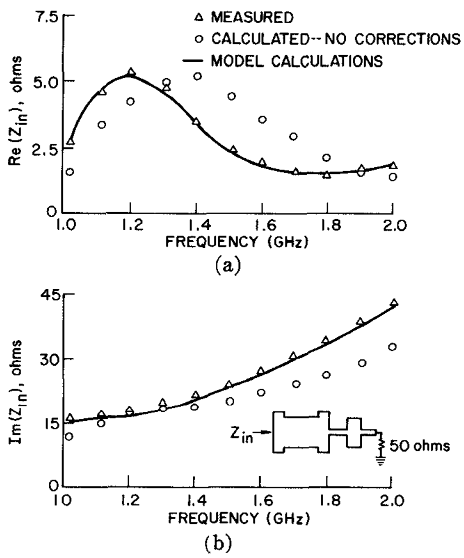

Fig. 3. Input impedance to sample circuit. (a) Input resistance. (b) Input reactance.

the large step discontinuities are appreciably in error. Errors of this magnitude render impossible the accurate computer design of complex microstrip matching networks for microwave power transistor amplifiers. On the other hand, with a set of equations approximating the junction model curves incorporated in a microstrip network optimization computer program, such amplifiers have been developed at the Naval Research Laboratory on a "work the first time" basis. The program directly generates corrected physical dimensions of the networks.

VI. CONCLUSIONS

No attempt is made to generalize the results presented herein. The equivalent circuit has been shown to be valid for large steps from 50-Ω lines, in the frequency range from 1 to 2 GHz, on 0.0635-cm-thick alumina substrate material. Other tests have indicated that the curves are usable for steps between any two impedances, neither of which is 50 Ω.

It should be recognized that great precision is required in the measurement of input impedances and physical dimensions, and in the knowledge of substrate dielectric constant, in producing the parameters for this junction model. This work shows that very large microstrip step effects can be accounted for, gives equivalent circuit parameters for a specific case, and shows the utility of the results.

It is hoped that the presentation of this information will stimulate further theoretical work on very large impedance steps in microstrip, so that more general analytical expressions can be developed to replace the specific results presented herein.

ACKNOWLEDGMENT

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Design of Partial Height Ferrite Waveguide Circulators

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Abstract—**This short paper presents a design procedure for the widely used three-port waveguide circulator that has a partial height ferrite post in the junction region. Design formulas and curves are derived for two configurations of partial height circulators: one has a short circuit at one end of the ferrite post, while the other has dielectric spacers at both ends. The design method was used to build two circulators for operation at 38 GHz and 60 GHz, respectively. Excellent agreement between theory and experiment was obtained in predicting the center frequency and required matching structure of these devices.**

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I. INTRODUCTION

The most commonly used broad-band waveguide circulator incorporates a partial height ferrite post in the junction. However, due to the complexity of this configuration, no exact theory for designing such a circulator has been obtained. Theories for only the full height ferrite post have been attempted [1], [2]. These analyses involve matching the dominant mode fields in the connecting waveguides to a summation of fields due to modes within the junction and thus require elaborate computer programs. Owen and Barnes [3] proposed that the partial height ferrite junction circulator operates in a turnstile fashion with rotating modes propagating along the ferrite axis. Later on, Owen [4] measured the phase-frequency responses of the eigenvalues from which the principal field modes could be experimentally identified and built an X-band circulator based on these measurements. He could adjust the ferrite geometry to achieve a 120° separation of the eigenvalues over a broad frequency range. However, the instrumentation for these measurements involved multiple phase shifters, attenuators, couplers, a 3-way power divider, and an HP network analyzer. This equipment is not readily available for devices operating above X-band.

In this paper, approximate formulas are presented which simplify the design of partial height waveguide circulators. They apply to both the single-ended and double-ended configurations of the so-called compact turnstile device. The design is used to build 38- and 60-GHz circulators. Good correlation is achieved between measured and predicted performance.

II. THEORY

The two types of partial height ferrite circulators shown in Fig. 1(a) and (b) are the one-sided compact turnstile device, which has a short circuit at one end of the ferrite rod, while the other is a double-sided compact turnstile with dielectric spacers on both ends. With either type, two of the three eigen excitations propagate axially along the rod in rotating modes which are circularly polarized in opposite senses. Magnetizing the ferrite increases the propagation constant of one of these modes and decreases the propagation constant of the other since these modes experience different permeabilities. The third eigen-excitation does not couple into the ferrite rod along the symmetry axis. It simply uses the rod as a dielectric resonator with its center frequency and phase φ_3 a function of the effective dielectric constant of the dielectric–ferrite combination and the effective radius (including the effect of fringing fields) of the resonator. With the ferrite magnetized, a phase displacement of $2n\pi$

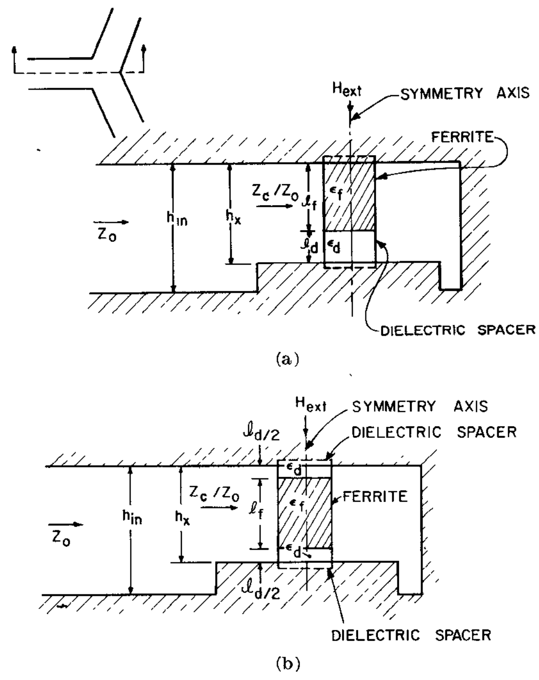

Fig. 1. (a) One-sided compact turnstile circulator. (b) Double-sided compact turnstile circulator.