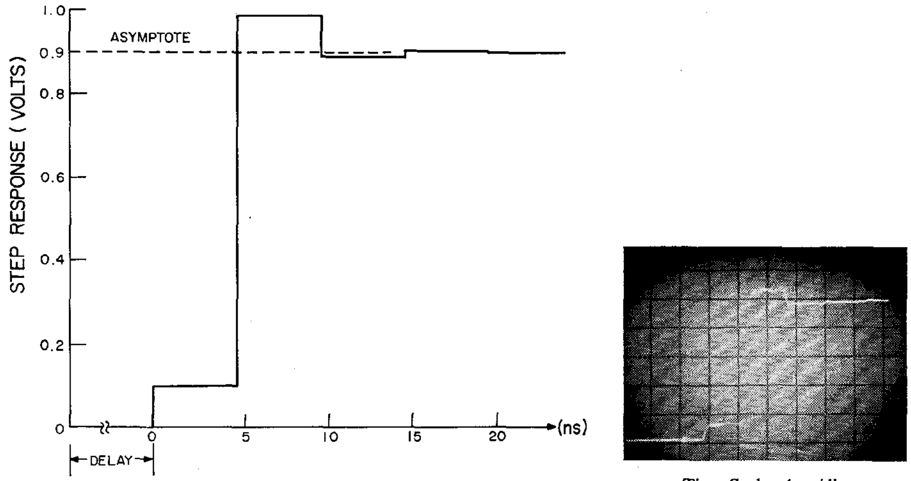

Time Scale: 4 ns/div

(a) (b)

Fig. 6. The step response of the equalizer. (a) Theoretical step response. (b) Measured response.

$$
\begin{aligned}
\hat{H}_{0A}(z) &\doteq \hat{H}_1(z)\hat{H}_2(z), \\
&= \left[\frac{(1+\Gamma_1)(1+\Gamma_2)z}{1+\Gamma_1\Gamma_2 z^2}\right] \\
&\quad \cdot \left[\frac{2}{3}\cdot\frac{\alpha}{(3.15)}\left(1+\frac{1}{\alpha}z^2\right)z^l\right], \\
&= Bz^{l+1}\left[\frac{1+\dfrac{1}{\alpha}z^2}{1+\alpha z^2}\right], \\
B &= \frac{2(1+\Gamma_1)(1+\Gamma_2)\Gamma_1\Gamma_2}{(3)(3.15)}, \\
\alpha &= \Gamma_1\Gamma_2.
\end{aligned} \tag{20}
$$

The *differential* attenuation required between lines in network #2 is equal to the ratio of $1/\Gamma^2 = 1/9$ [see (10) and (13)]: a voltage reduction factor of about 1/9 is achieved by the use of a 19 dB pad. The first pole-zero pair occurs when the magnitude of the real part of p equals

$$
\begin{aligned}
|\operatorname{Re} p| &= \left|\frac{1}{2\tau}\ln\frac{1}{\Gamma^2}\right| = \frac{0.952}{2\tau_1} \\
&= \frac{0.952}{\tau_2}
\end{aligned} \tag{21}
$$

and

$$
|\operatorname{Im} p| = \frac{\pi}{2\tau_1} = \frac{\pi}{\tau_2} = 2\pi f_0
$$

$$
f_0 = \frac{1}{4\tau_1} = \frac{1}{2\tau_2}.
$$

If

$$
\tau_1 = 2.5 \times 10^{-9} \text{ seconds}
$$

then

$$
f_0 = 100 \text{ MHz}
$$

$$
\sigma = \frac{0.952}{5 \times 10^{-9}} = 0.191 \times 10^9. \tag{22}
$$

$\tau_1 = 2.5 \times 10^{-2}$ seconds corresponds to a line length of

$$
\begin{aligned}
L_1 &= \frac{c\tau_1}{\sqrt{\epsilon}} = \frac{3\times 10^8 \times 2.5 \times 10^{-9}}{2.26} \\
&= 0.5 \text{ meter} \\
\Delta L &= 2L_1 = 1 \text{ meter}
\end{aligned} \tag{23}
$$

where the dielectric constant ϵ equals 2.26 for polystyrene.

The overall loss coefficient of the network is B. From (20)

$$
\begin{aligned}
B &= \frac{2(1+\Gamma_1)(1+\Gamma_2)\Gamma_1\Gamma_2}{(3)(3.15)} \\
&= \frac{2\left(\frac{2}{3}\right)\left(\frac{2}{3}\right)\left(\frac{1}{9}\right)}{(3)(3.15)} = \frac{1}{94} \\
&= 39.5 \text{ dB loss} \\
&\qquad \text{(independent of frequency)}
\end{aligned} \tag{24}
$$

The network shown in Fig. 5 was fabricated in the laboratory. It was excited by a unit step generator and the output displayed on a sampling oscilloscope. The theoretical step response was obtained by multiplying (20) by $1/p$, dividing numerator by denominator and inverting the resulting expression term by term and is plotted in Fig. 6(a); a measurement of the actual step response is shown in Fig. 6(b). It can be seen that the results are in close agreement with the theory.

In conclusion, it should be noted that the network can readily be scaled to make, for example, the first pole-zero cluster occur at 1 GHz. Here the line length L must be reduced by a factor of ten i.e., $L = 0.05$ meter. The network loss can be reduced by using wide band isolators and/or a TWT amplifier centered appropriately in the band of interest.

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Mounted Diode Equivalent Circuits

The use of semiconductor diodes at microwave frequencies continues to arouse interest in practical circuit characterization for mounted diodes.

Accurate wideband circuit description and evaluation techniques are needed to design components for predictable performance or optimize component response, and are essential if the design work is to be done by a computer on components employing widely spaced frequencies and wide bandwidths such as occur with parametric amplifiers and multipliers.

Previously Getsinger[1] gave theoretical background on a conceptual scheme for describing microwave frequency behavior of packaged diodes and mounts by the use of lumped-element equivalent circuits. Separation of the circuit of the packaged diode from the circuit of the mount was shown to be possible if the equivalent circuit of the packaged diode described the radial-line TEM-mode impedance found at a terminal surface defined to be at the outer diameter of the cylindrical diode package.

The purpose of this correspondence is to extend the work mentioned above by pointing out that the same point of view can be used even when the diode is unpackaged, that eliminating the diode package does not eliminate the coupling-network problem, that a more rigorous consideration of the diode mounting situation indicates that the diode equivalent circuit and mount equivalent circuit are not completely separable although the common elements are usually very small and can be measured, and finally that the approach of the referenced paper[1] holds even when the diode is not mounted between parallel metal planes, as was specified therein, provided mount parameters can be measured with the diode in place.

The basic diode coupling problem is to relate the electromagnetic fields across macroscopic terminal surfaces of a uniform transmission line to the fields within a microscopic region (the diode junction) of space.

In the preceding work[1] it was pointed out that the solution to this problem could usually be stated in terms of lumped-element circuit mathematics, and thus, in terms of an equivlent circuit or coupling network between the uniform-line terminal surface and the diode junction.

If one considers the total equivalent circuit of a packaged diode mounted in a useful way, and then mentally removes the packaging, it can be seen that the circuit configuration does not change; only the values of some circuit elements are changed. Thus, the circuit problem may not be perceptibly simpler without a package than with one. In fact, it is often helpful to define a cylindrical enclosure for the unpackaged mounted diode and treat it as though it were packaged.

For example, the circuit of Fig. 1 holds for a diode mounted in a waveguide, whether or not the diode is packaged. If packaged, the

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[1] W. J. Getsinger, "The packaged and mounted diode as a microwave circuit," *IEEE Trans. Microwave Theory and Techniques*, MTT-14, pp. 58–69, February 1966.

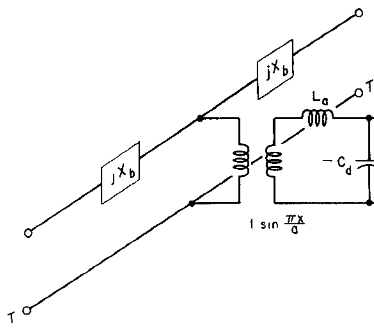

Fig. 1. Equivalent circuit for waveguide-mounted diode.

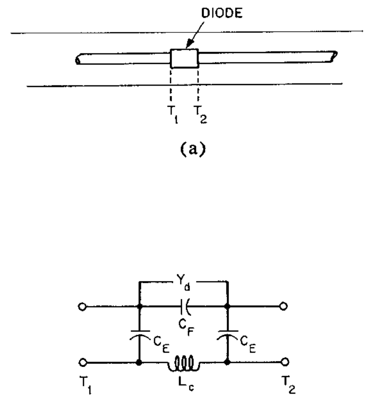

Fig. 2. Diode mounted in a coaxial line.

outer surface of the package is chosen as a convenient radial-line terminal surface. If unpackaged (with the semiconductor die mounted on one broad wall, and connecting wires running to the opposite broad wall), a radial-line terminal surface is chosen to lie at the surface of a hypothetical cylinder of convenient diameter that surrounds the die and connecting wires. Then, the fields external to the hypothetical surface are assigned to the waveguide and coupling network. This allows L_a and C_d to be calculated. The fields inside the surface are assigned to the diode equivalent circuit, which must be measured using the hypothetical surface as a radial-line terminal surface.

While it is normally a very good approximation to assume that the equivalent circuits of the diode mount and the packaged (real or defined) diode are completely separable it is not strictly so. First, consider the shunt-mounted diode, exemplified by a diode mounted across a waveguide, (Fig. 3(a) of Getsinger[1]). The circuit is given here in Fig. 1. It has equal series elements X_b on either side of the shunt connection. These elements were neglected in Getsinger[1] because they are small for small diameter diodes. The value for X_b depends on the construction of the diode package and internal connections to the junction. A value for X_b can be found for any given diode and waveguide by imposing equal but opposite RF voltages either side of the waveguide reference plane T, and measuring the impedance introduced when the diode is mounted at the reference plane.

A Sylvania D-5047 diode, 0.080 inch in diameter, mounted at the center of a waveguide 0.900 by 0.060 inches in cross section gave a value for X_b of -0.7 ohms at about 10.8 GHz.

The measurement was made using a diode mount at the middle of a long length of waveguide terminated at each end in a small coupling aperture. The resonances occurring at those frequencies for which the structure (with diode absent) is an even number of half guide wavelengths long place equal voltages of opposite sign on either side of the terminal plane. Placing the diode in its mount causes the even numbered resonances to undergo frequency shifts from which X_b can be calculated.

Now consider a series-mounted diode in coaxial line, shown in Fig. 2(a). The diode length is taken to be much smaller than a wavelength, but the diode package diameter will be considered to be larger than the coaxial-line center conductor in which it is mounted.

Since the region between terminal surfaces T_1 and T_2, bounded also by the outer conductor and the surface of the diode package, is small in wavelengths, we may represent the network connecting the coaxial-line terminal surfaces to the packaged-diode radial-line terminal surface by lumped elements, in accordance with the arguments of Section II, of Getsinger.[1]

If equal in-phase RF voltages are applied at T_1 and T_2, and the admittance is measured at T_1, that admittance ($j\omega C_E$) is found to result from shunt capacitance due to fringing from the end of the diode package and to direct capacitance between the internal conductors of the diode package and the outer conductor modified slightly by diode and line inductance. This measurement gives the shunt elements of an equivalent π network in which the packaged-diode radial-line equivalent circuit can be incorporated as part of the series arm.

Consideration of the mounting configuration indicates that the inductance L_c of the region between T_1, T_2, the diode package surface, and the outer conductor should be in series with the packaged-diode equivalent circuit. Also, it would be expected that the electric fields in the region between the diode ends, but external to the diode package would be different than when the diode is mounted between parallel metal plates. This calls for a small corrective capacitance C_F which might be positive or negative, across the packaged-diode terminal surface. The resulting equivalent circuit is shown in Fig. 2(b) where Y_d is the total packaged-diode admittance as measured at its radial-line terminal surface.

This can be seen to be equivalent in form to the circuit of Fig. 6(b) of Getsinger[1] drawn for the situation where the diode package was smaller than the coaxial-line center conductor.

Sometimes it is desired to have a single terminal surface for the coaxial line at the middle of the diode, rather than terminal surfaces at each end. The equivalent circuit is then of the same form as in Fig. 2(b), but with values for the elements C_E and L_c adjusted for the assumption that the coaxial line now continues to the middle of the diode. Again, only L_c can be calculated; it is now the difference between the inductance external to the diode package over its length and the inductance that would exist over the same length if the diode were replaced by a continuation of the coaxial-line center conductor. For the illustration, Fig. 2(a), where the diode package diameter is greater than the center conductor diameter, L_c would have a negative value.

In these circuits, the capacitances C_E are appropriately charged against the mount, but their values depend partially on the structure of the diode internal to its package as well as on the construction of the mount. However, this contribution from the internal parts of the package is probably quite small, as was found in the waveguide situation.

It should be observed that the determination of element values for these circuits ultimately becomes a problem in measurements, because the complex mechanical configurations of diode packages or connecting straps and wires, and variations in these structures from one unit to the next make accurate analytical methods impractical.

The author hopes to subsequently describe practical measurement techniques for diodes and mounts.

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[2] Operated with support from the U. S. Air Force.

Dual-Mode Coupler

The dual-mode coupler discussed in this correspondence is a junction in which two rectangular waveguides excite orthogonal dominant modes (TE_{10} and TE_{01}) in a square waveguide. The coupler performs well over a relatively broad frequency band, and its simple configuration lends itself to fabrication by electroforming.

Several dual-mode couplers of various configurations have been reported previously.[1]–[12] The coupler described here is similar to that reported by LaPage[12] except that the side, or auxiliary, arm is coupled to the square guide through a resonant iris which is opposite the sloped face of the taper.

The basic design procedure was to design a linear double taper[13] for the straight-through arm and a resonant coupling iris for the side arm. The position of the side arm then was adjusted for the most desirable impedance characteristics over the design band, and it was matched with an inductive iris.

Manuscript received April 6, 1967. The work reported here was supported by the U. S. Naval Ship Systems Command.

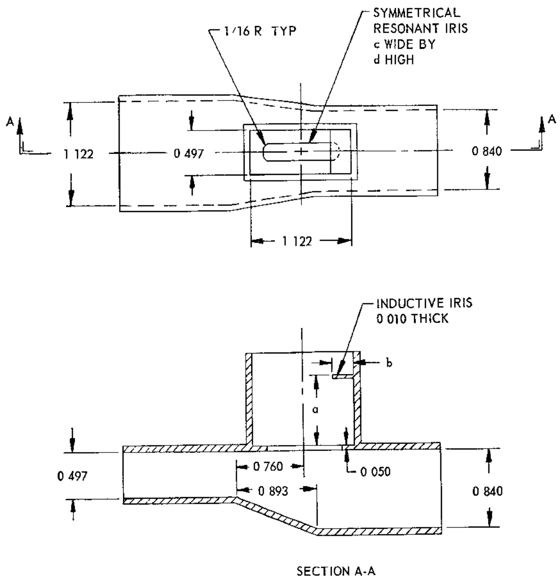

Fig. 1. Configuration of the dual-mode coupler.

TABLE I

VALUES OF LETTER DIMENSIONS FOR THE COUPLER OF FIGURE 1

Dimension	Model 1 (inches)	Model 2 (inches)
a	0.765	0.465
b	0.200	0 196
c	0.780	0.910
d	0.230	0.406

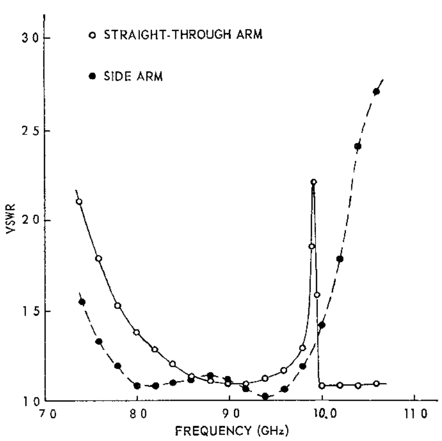

Fig. 2. VSWR versus frequency for Model 1 of the dual-mode coupler.

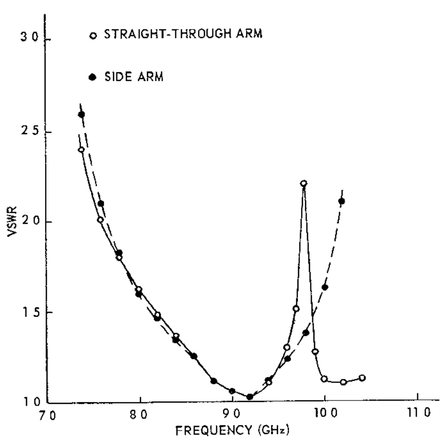

Fig. 3. VSWR versus frequency for Model 2 of the dual-mode coupler.

Two models were designed and tested. Their configuration including dimensional information is shown in Fig. 1; specific dimensions of the irises are given in Table I. The smaller resonant iris in Model 1 yields a wider bandwidth; however, it reduces the power-handling capabilities of the side arm.

Both models were tested for VSWR, power handling capabilities, and isolation. The VSWR versus frequency characteristics are shown in Figs. 2 and 3. Both couplers were tested at atmospheric pressure with a pulsed power source having a pulse length of 1.0 μs, a pulse repitition frequency of 600 Hz, and a frequency coverage from 8.5 to 9.6 GHz. The straight-through arm of both models handled over 250 kW peak; the side arm handled over 125 kW peak in Model 1 and over 250 kW peak in Model 2. In both couplers, the isolation between the rectangular arms exceeded 50 dB over the test frequency bands of 8.2 to 9.8 GHz and 8.4 to 9.6 GHz for Models 1 and 2, respectively.

The performance reported here does not necessarily represent the ultimate limits for this type of coupler. It is felt that with changes in the taper and with a different shape for the resonant iris the coupler would perform well over a broader frequency band.

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Comment on "Isolation of Lossy Transmission Line Hybrid Circuits"

I should like to call attention to several errors appearing in the above correspondence.[1] These errors are presumably typographical, and indeed do not affect the calculated isolation values. However, they could impair the utility of the correspondence to the reader wishing to extend the analysis to calculate the remaining properties of the two hybrid types. The errors are listed in the following.

Equation (13) should read

Manuscript received May 31, 1967.

[1] R. M. Kurzrok, *IEEE Trans. Microwave Theory and Techniques (Correspondence)*, vol. MTT-15, pp. 127–128, February 1967.