ability to achieve high quality performance is a function of the instrumentation employed.[7] Use of a swept oscillator and a high directivity coupler is essential.

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[7] P. Foldes and T. B. Thompson, "A waveguide quadruplexer system," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-9, pp. 297–306, July, 1961.

A Microwave Power Limiter*

I. INTRODUCTION

Through the use of the familiar hybrid directional coupler and two *PIN* junction diodes as load impedances, the power delivered to the port not directly coupled to the input may be limited to any desired level, without rigid tolerances on the diode impedances.

Starting with the scattering matrix of the hybrid and specifying arbitrary load reflection coefficients, the general properties of the device will be derived, and then its power-limiting capabilities will be examined in conjunction with the properties of the *PIN* diode.

II. PROPERTIES OF THE HYBRID COUPLER

With the ports of the directional coupler numbered as shown in Fig. 1, its scattering matrix is

$$S = \begin{bmatrix} 0 & \beta & \alpha & 0 \\ \beta & 0 & 0 & \alpha \\ \alpha & 0 & 0 & \beta \\ 0 & \alpha & \beta & 0 \end{bmatrix}. \tag{1}$$

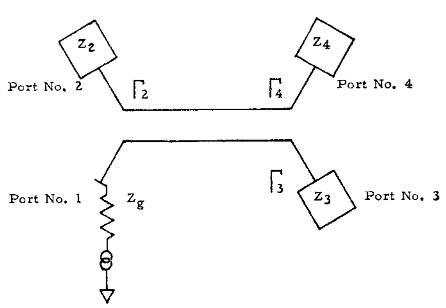

Fig. 1—Schematic representation of 3-db hybrid.

For the hybrid directional coupler, the power delivered to ports 2 and 3 under matched conditions is equal; hence,

$$|\alpha| = |\beta|$$

and from the unitary conditions on the scattering matrix of a lossless network,

$$|\alpha|^2 + |\beta|^2 = 1$$

$$\therefore \quad |\alpha| = |\beta| = \frac{1}{\sqrt{2}}$$

and

$$\beta(\alpha)\dagger + \alpha(\beta)\dagger = 0$$

$$\beta(\alpha)\dagger = -\alpha(\beta)\dagger$$

$$\arg\beta - \arg\alpha = \pm\pi + \arg\alpha - \arg\beta$$

$$\therefore \quad \arg\beta - \arg\alpha = \pm\frac{\pi}{2}.$$

This equation leads to the conclusion that no power is reflected back at port 1 if ports 2 and 3 are terminated in the same impedance.

The power delivered to port 4 is of interest now. The conventional notation will be used for the incident and reflected normalized voltages at the ports of the network, *i.e.*, a_n is incident voltage on a network at port n, and b_n is the voltage reflected from the network at port n. The power delivered to the load at port 4 is

$$P_4 = \frac{|b_4|^2 - |a_4|^2}{2}.$$

Since $a_4 = \Gamma_4 b_4$, where Γ_4 is the reflection coefficient of the load at port 4

$$P_4 = \frac{|b_4|^2}{2}(1 - |\Gamma_4|^2).$$

Consider the scattering equations of the network [(1)]

$$b_1 = \beta a_2 + \alpha a_3$$
$$b_2 = \beta a_1 + \alpha a_4$$
$$b_3 = \alpha a_1 + \beta a_4$$
$$b_4 = \alpha a_2 + \beta a_3.$$

Solving for $|b_4|$ in terms of the network parameters and $|a_1|$

$$P_4 = 2|\alpha|^2|\beta|^2|\Gamma|^2(1 - |\Gamma_4|^2)|a_1|^2$$

where $|\Gamma|$ is the magnitude of reflection coefficient of identical loads at ports 2 and 3.

For a matched load at port 4, $\Gamma_4 = 0$ and since $|\alpha| = |\beta| = 1/\sqrt{2}$,

$$P_4 = \tfrac{1}{2}|\Gamma|^2|a_1|^2.$$

This equation underlines the fact that the power delivered to a matched load depends only on the reflection coefficient at ports 2 and 3, if $\alpha = \pm j\beta$, and on the input power.

Therefore, if the magnitude of the reflection coefficient at ports 2 and 3 should decrease as the power incident on port 1 increases, the power delivered to the load at port 4 will remain constant. Of course, if an exact functional relationship can be formulated between $|\Gamma|$ and $|a_1|$, the expression of P_4 can be written exactly. This may or may not be necessary depending on the application. The *PIN* junction diode is a device whose impedance displays this necessary dependence on power.

III. *PIN* JUNCTION DIODES

A *PIN* junction diode is one in which a high resistivity *I* layer is introduced between normally used *P*- and *N*-type semiconducting materials. The presence of the *I* layer results in low capacitance per unit of area, and thus at any given impedance level, permits the use of larger diodes with high burnout power levels. Similar to the PN junction, the *PIN* diode acts as a high *Q* capacitor at low levels. At high levels, the *I*-layer resistivity is greatly reduced by conductivity modulation, resulting in a lower overall resistance. The application of a dc forward voltage also has certain effects worth noting. In this case, electrons and holes are injected into the practically intrinsic region and microwave power is absorbed by the mobile charge carriers thus introduced into the intrinsic part. The mechanism of transport of charge carriers into this region is governed by diffusion and recombination. Theory shows that the excess hole and electron concentrations are almost homogenous if the distance between the *P* and *N* regions is not larger than

$$L = \sqrt{D\tau}$$

where

L = diffusion-recombination length
D = ambipolar diffusion constant
τ = average life of electron-hole pairs.

The time necessary to establish a certain concentration pattern is also of the order of τ. Hence, τ determines the maximum modulation of switching frequency obtainable. It has been found that for germanium τ is in the order of five μsecs.

The microwave equivalent circuit for typical *PIN* junction diodes is illustrated in Fig. 2. The values indicated are for a zero-bias condition where resistivity of the *I* layer is in the neighborhood of 1800 ohm-cm.

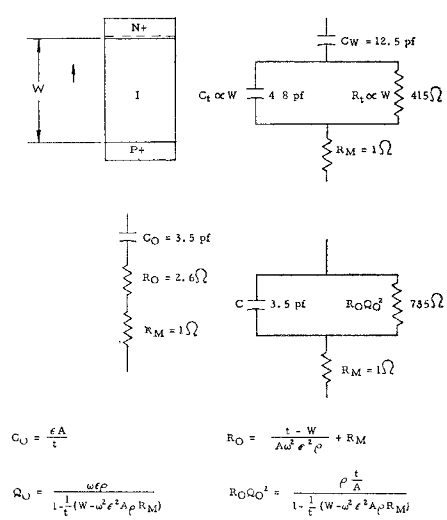

Fig. 2—*PIN* diode equivalent circuit.

IV. DISCUSSION

Since the coupling variation between arms of the hybrid can be controlled to within ±0.1 db, the hybrid itself will not cause any unbalances; however, in order to avoid reflections at the input, the terminations at ports 2 and 3 must be equal in magnitude and phase angle (*i.e.*, $\Gamma_2 = \Gamma_3$). This

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† Denotes complex conjugate.

is easily achieved since this "balance" is really required only during low-level operation. Production of *PIN* diodes can be controlled to match pairs of these devices to good tolerances, relative to their low-level operation. What does become difficult to control is the change of *I*-layer resistance as a function of input power or dc bias. In this case, however, it is not important since, at high power levels, the only requirement is that no excess power be delivered to the detector.

It can be shown, using the scattering technique, that the reflected power at the input is proportional to the square of the difference between the reflection coefficient of the impedances terminating ports 2 and 3, with $\Gamma_4=0$. The return loss, in the case where the *PIN* diodes capacitance differs by 20 per cent, is 15.9 db corresponding to an input VSWR of 1.38:1, which is a negligible amount.

This device has as its principal advantage the fact that it is capable of retaining its power-limiting properties independent of frequency over extremely wide frequency ranges. As an example, with the proper hybrid design, this device can be made operable over a 10:1 frequency band. Its application would be in many cases where power-limiting is required. Typical of these might be the protection of video crystals from high input power levels. Another application which finds wide use, particularly since the advent of extremely low-noise receivers, is the protection of parametric or maser "front ends." These solid-state devices have a very high susceptibility to high power inputs, so much so that they become completely inoperable, and in some cases permanent damage results. The device described here would protect receivers which use masers or parametric techniques from overloading, while at the same time would not introduce insertion losses which would detract from the over-all receiver performance.

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A Four-Component Polarization Resolver*

Numerous ways are known for resolving a polarized microwave signal into *pairs* of orthogonally polarized components. Although pairs of orthogonal components contain redundant information, in many experimental applications it is desirable to have both linearly polarized and circularly polarized components available for comparison purposes. These four polarization components are derived simultaneously in the simple resolver-mixer assembly of Fig. 1. The

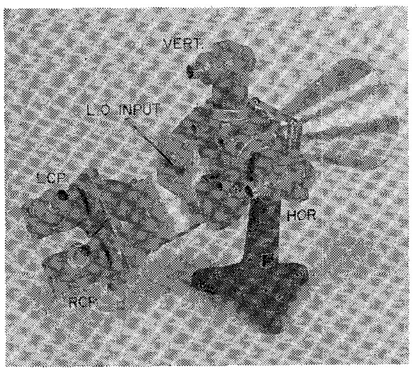

Fig. 1—Four-component polarization resolver for X-band simultaneously provides orthogonal linear and orthogonal CP components of an input signal.

X-band device shown represents an extension of principles applied earlier to an instantaneous polarimeter.[1]

In the four-component resolver a trimode turnstile junction is supplemented with a quadrature hybrid junction of the short-slot variety, which for compactness is contained in a 90-degree bend. The device resolves one-half the signal power into orthogonal linear components and the remaining half into orthogonal circular components. Local oscillator drive, injected through the "coaxial" port of the trimode turnstile junction, divides equally among the four mixer crystals. Symmetry of the plumbing insures phase preservation, although at moderate IF's it is not necessary that detector arm lengths be electrically equal.

The device is well adapted to measuring the polarization backscatter characteristics of radar targets, and is currently being employed very successfully in model range polarization studies. When used in an instantaneous polarimeter, the four-component resolver offers advantage over the original polarimeter design[1] in that the "sense" of a polarization pattern being observed is readily indicated by a simple determination of which circularly polarized component is the greater.

A variation of the four-component resolver, recently used in a revealing radar backscatter study,[2] employs a *conventional* turnstile junction and a quadrature hybrid junction. Auxiliary circuitry was employed for local oscillator injection, although for less critical applications video detectors might be employed on the four ports of the resolver to obtain relative amplitude information for the four different polarizations.

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[1] P. J. Allen and R. D. Tompkins, "An instantaneous microwave polarimeter," Proc. IRE, vol. 47, pp. 1231–1237; July, 1959.

[2] I. D. Olin and F. D. Queen, "Measurements Using A Polarization Instrumentation Radar on Navigational Buoys," U. S. Naval Res. Lab., Washington, D. C., NRL Rept. No. 5701; November, 1961.

The Channel Waveguide*

A waveguide propagating the TE_{10} mode can carry more power than the normal rectangular waveguide if it has a symmetrically placed channel in the E-plane as shown in Fig. 1. The greater height of the channel in the center of the waveguide will allow a higher voltage to be applied before dielectric breakdown occurs. The TE_{10} cutoff wavelength λ_c was investigated using the methods of Iashkin[1,2] and Cohn[3] to find out if the cutoff wavelength of the channel waveguide was equivalent to that of the rectangular waveguide λ_{cr}.

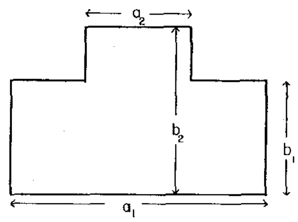

Fig. 1—Basic cross section of the channel waveguide.

An equation derived by Iashkin that fits the cross section in Fig. 1 is

$$\cot K\left(\frac{a_1-a_2}{2}\right)=\frac{b_1}{b_2}\left\{\tan K\left(\frac{a_1-a_2}{2}-\frac{\pi}{2}\right)+2K\sum_{n=1}^{\infty}\frac{\tanh\sqrt{\left(\frac{n\pi}{b_2}\right)^2-K^2}\left(\frac{a_1-a_2}{2}-\frac{\pi}{2}\right)\sin^2\frac{n\pi b_1}{b_2}}{\sqrt{\left(\frac{n\pi}{b_2}\right)^2-K^2}\left(\frac{n\pi b_1}{b_2}\right)^2}\right\} \tag{1}$$

where K, the wave number, is equal to λ_{cr}/λ_c and is the only unknown. a_1 is set equal to π and b_1, b_2, and a_2 are normalized with respect to π. A restriction on the equa-

* Received by the PGMTT, February 5, 1962.

[1] A. I. Iashkin, "A method of approximate calculation for waveguides of triangular and trapezoidal cross-sections," *Radio Engrg.*, vol. 13, pp. 1–9; October, 1958.

[2] A. I. Iashkin, "The calculation of the fundamental critical wavelength for a rectangular waveguide with longitudinal rectangular channels and ridges," *Radio Engrg.*, vol. 13, pp. 8–14; March, 1958.

[3] S. B. Cohn, "Properties of ridge waveguides," Proc. IRE, vol. 35, pp. 783–788; August, 1947.