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

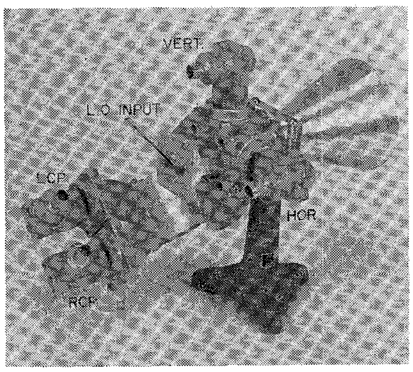


Fig. 1—Four-component polarization resolver for *X*-band simultaneously provides orthogonal linear and orthogonal CP components of an input signal.

*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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* Received by the PGMTT, February 16, 1961; revised manuscript received, January 7, 1962.

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[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 $\lambda_c$ 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 $\lambda_{cr}$.

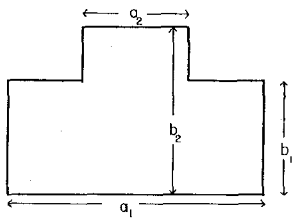


Fig. 1—Basic cross section of the channel waveguide.

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

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

where $K$, the wave number, is equal to $\lambda_{cr}/\lambda_c$ and is the only unknown. $a_1$ is set equal to $\pi$ and $b_1$, $b_2$, and $a_2$ are normalized with respect to $\pi$. A restriction on the equa-

* Received by the PGMTT, February 5, 1962.

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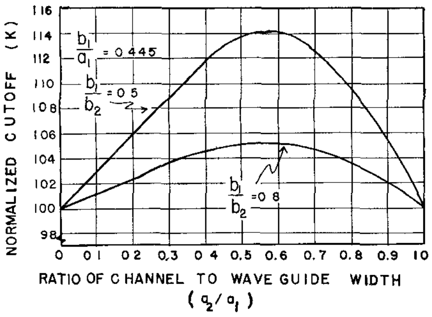


Fig. 2—Ratio of $\lambda_{cr}/\lambda_c$ for two ratios of $b_1/b_2$ with $a_2$ varying for (1).

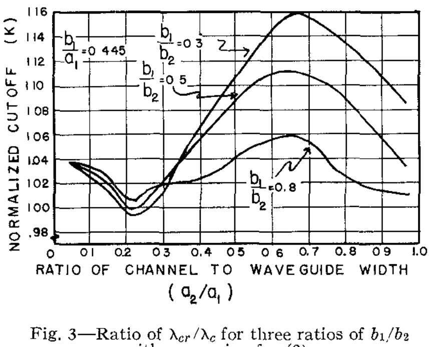


Fig. 3—Ratio of $\lambda_{cr}/\lambda_c$ for three ratios of $b_1/b_2$ with $a_2$ varying for (2).

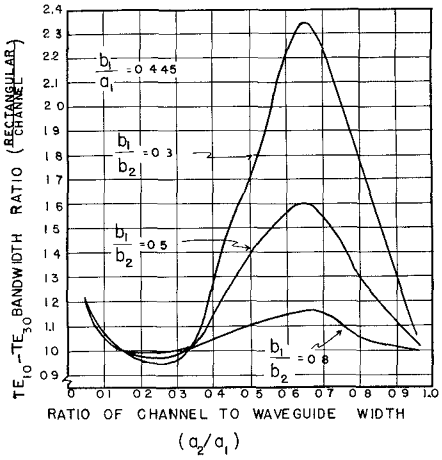


Fig. 4—$TE_{10}-TE_{30}$ bandwidth ratio (rectangular/channel) obtained from (2).

tion is that $K<\pi/b_2$ and $\pi/b_2>1$. The first two terms in the series were used. The results are given in Fig. 2.

A derivation of $\lambda_c$ for the cross section of Fig. 1 using Cohn's method gives (2) which is good for $TE_{2n+1}$ modes.

$$\lambda_c = \frac{7.1\times 10^{11} b_1 c}{\cot\dfrac{\pi(a_1-a_2)}{\lambda_c} - \dfrac{b_1}{b_2}\tan\dfrac{\pi a_2}{\lambda_c}} \cdot \qquad (2)$$

This equation is exactly the same as the one used by Cohn for the ridge waveguide. The discontinuity capacitance $C$ was obtained from a paper by Whinnery and Jamieson.[4] The proximity effect mentioned in this paper was taken into account. The results are given in Fig. 3. The $TE_{10}-TE_{30}$ bandwidth was also found, and the results are given in Fig. 4. The $TE_{10}-TE_{20}$ bandwidth curves are not given; because, as pointed out by Cohn, a symmetrical transmission system will not be affected by the $TE_{20}$ mode.

For the limiting case of $b_1/b_2 \to 0$, (1) and (2) cannot be used, but the trend toward a shorter $\lambda_c$ can be noted except when $a_2 \to a_1$, or 0. When $b_1/b_2 \to 1$, (2) again cannot be used, but (1) definitely gives the correct result of $K=1$ for all values of $a_2$.

In general, the results seem to show that $\lambda_c$ will be shorter for the channel waveguide than the rectangular waveguide but that it can approach $\lambda_{cr}$. A shorter $\lambda_c$ is not necessarily a disadvantage. The channel waveguide can act as a high pass filter and carry the same or greater power than a rectangular waveguide which would allow lower frequencies to propagate. The $TE_{10}-TE_{30}$ bandwidth for the channel waveguide can also be made to approach the bandwidth for the rectangular waveguide. The power handling capacity naturally will be greater for the channel waveguide. It can be noted that there are discrepancies between the results for the two methods. Cohn's equation can be used when $b_2>a_1$, where Iashkin's equation cannot be used. Iashkin's equation can give results when $a_2 \to 0$ and when $b_1/b_2 \to 1$, while Cohn's equation cannot be used in these areas. These discrepancies will be resolved by experimental data.

The authors would like to acknowledge the use of the IBM 1620 Computer of the Marquette University Computing Center in obtaining these results.

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## Frequency Diplexing with Waveguide Bifurcations*

A relatively simple form of microwave frequency diplexer consists of a waveguide $E$- or $H$-plane $Y$-junction terminated with bandpass filters in each of the two output ports. The $E$- or $H$-plane bifurcation, or zero-degree $Y$-junction, allows a compact diplexer design as shown in Fig. 1.

The conditions required of such junctions for ideal diplexing and the limitations imposed by the impedance functions of the filters are outlined. The design technique for achieving "optimum" diplexing with two contiguous filters and experimental verification of the design are then given in the following sections.

* Received by the PGMTT, September 2, 1960; revised manuscript received, February 12, 1962.

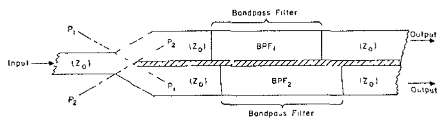


Fig. 1.

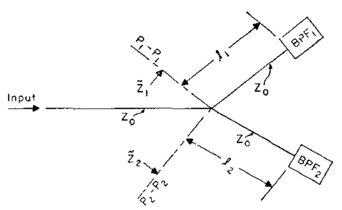


Fig. 2.

### Diplexing Requirements

The basic design requirement for diplexing is that maximum power transfer occur throughout the passband of each filter channel when both output ports are terminated in their characteristic impedance ($Z_0$). One method for satisfying the basic diplexing requirement is to achieve zero filter junction reactances (with an $E$-plane bifurcation) at $P_1-P_1$ and $P_2-P_2$ of Fig. 2 throughout the passbands of both filters. In general, conjugate filter junction reactances are necessary for maximum power transfer; zero filter junction reactances are a special case when using the $E$-plane bifurcation. During the review of this correspondence by the PGMTT Editorial Board, the writer was made aware of the similarity between the filter characteristics required and those possessed by lumped constant complementary filters which have been treated extensively by Bell Laboratories and other authors.

A reduction in maximum power transfer through the receiving channel, or diplexing loss, occurs because the filter junction impedance is not purely reactive throughout the receiving passband. The impedance of a filter[1–4] is complex and a rapidly changing function of frequency so that the conditions for ideal diplexing are not physically realizable. Ideal diplexing may be approached only when the filter passbands are noncontiguous or if a definite stopband exists between the passbands.

The diplexer bandwidth is limited by the bandwidth of the diplexing junction. A well designed junction will exhibit negligible junction and discontinuity effects so that its bandwidth will approach that of the waveguide. The VSWR-frequency response may be used as a measure of quality for the diplexing junction.

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