Correspondence

Direct Coupled Coaxial and Waveguide Band-Pass Filters*

In this note, experimental results will be presented for practical coaxial and waveguide band-pass filters. Filters of both types were originally developed in the 1700 Mc to 2300 Mc frequency range for application in a wide-band microwave radio relay communications system. The specifications for these filters called for very low input VSWR's over an appreciable part of the filter pass bands. This is necessary to minimize the degradation in system performance due to intermodulation noise resulting from feeder distortion of microwave transmission lines[1] and/or group delay distortion within the microwave filters.[2] Both the coaxial and waveguide band-pass filters employed five direct coupled resonators. Filters were designed for Butterworth (maximally flat amplitude) response shapes and nominal filter 3 db bandwidths of 60 Mc.

The coaxial band-pass filter employed open-ended $\lambda/4$ resonators with cylindrical center conductor coaxial to rectangular outer conductors. (See Fig. 1.) A single filter model was capable of covering the entire 1700 Mc to 2300 Mc frequency range using fixed circular apertures as interstage coupling mechanisms. These apertures were designed for minimum frequency sensitivity.[3] Input/output coupling was achieved using adjustable capacitive probes. At any desired center frequency, the gaps between the input/output probes and the coaxial center conductors were varied and the lengths of the five resonator tuning slugs were adjusted for minimum pass band VSWR of maximal flatness. Pass band VSWR vs frequency for a typical alignment of the coaxial band-pass filter is shown in Fig. 2. Filter dissipation loss within the pass band was less than $\frac{1}{2}$ db so that lossless filter theory is applicable. Neglecting dissipation, the measured filter 3 db bandwidth can be approximated by the VSWR$=6.0$ bandwidth which is 65 Mc. For a five resonator maximally flat band-pass filter with a 65 Mc 3 db bandwidth, the theoretical 1.10 VSWR bandwidth is 35.7 Mc. The measured 1.10 VSWR bandwidth is 34 Mc.

The waveguide band-pass filters employed the same coupling mechanism for both input/output and interstage couplings (see Fig. 3). This was a pair of full height inductive posts and an adjustable partial height capacitive screw. Because this type of coupling is quite frequency sensitive, all

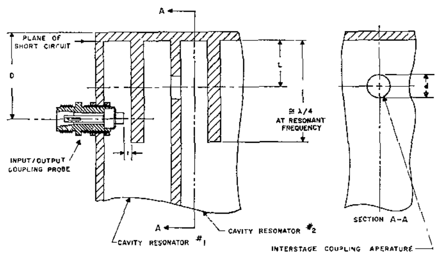

Fig. 1—Coaxial band-pass filter couplings.

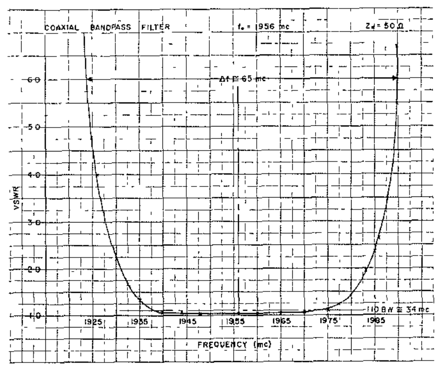

Fig. 2.

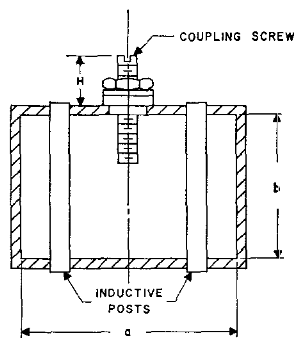

Fig. 3—Waveguide band-pass filter coupling.

couplings were adjustable. Waveguide resonators were TE 101 mode cavities tuned with a centered capacitive screw. Because excessive penetration of the capacitive tuning screws can seriously degrade the resonator unloaded Q's, the 1700 Mc to 2300 Mc frequency range was divided into three 200 Mc bands and three different filter models were employed. This limited maximum penetration of the tuning screws to less than 40 per cent of the waveguide height. ($b=2.150''$ in RG-104/U waveguide). Typical plots of pass band VSWR vs frequency are shown in Figs. 4–6 for the three different waveguide filter models. Filter 3 db bandwidths were between 55 Mc and 65 Mc with 1.10 VSWR bandwidth in excess of 20 Mc. Filter dissipation losses in the pass band were about 0.3 db.

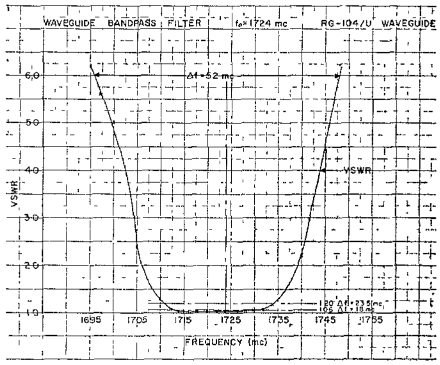

Fig. 4.

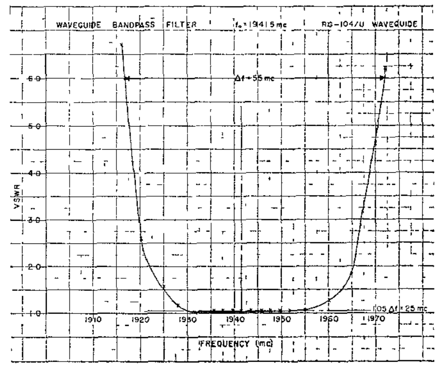

Fig. 5.

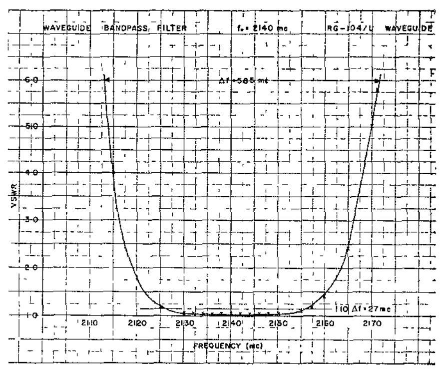

Fig. 6.

The microwave filters discussed in this paper were developed using the techniques of Dishal[4,5] and Cohn.[6] Performance capabilities of all filter coupling elements were determined by experimental methods. The

* Received by the PGMTT, December 22, 1961.

[1] R. G. Medhurst, "Echo-distortion in frequency modulation," *Electronic and Radio Engr.*, pp. 253–259; July, 1959.

[2] L. E. Thompson, "Distortion in multichannel frequency-modulation relay systems," *RCA Rev*, vol. 11, pp. 453–464; December, 1950.

[3] R. M. Kurzrok, "Design of interstage coupling apertures for narrow band tunable coaxial bandpass filters," (Correspondence) IRE Trans. on Microwave Theory and Techniques, vol. MTT-10, p. 143; March, 1961.

[4] M. Dishal, "Dissipative band-pass filters," Proc. IRE, vol. 37, pp. 1050–1069; September, 1949.

[5] M. Dishal, "Alignment and adjustment of synchronously tuned-multiple-resonator-circuit filters," Proc. IRE, vol. 39, pp. 1448–1455; November, 1951.

[6] S. B. Cohn, "Direct-coupled resonator band-pass filters," Proc. IRE, vol. 45, pp. 187–195; February, 1957.

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

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

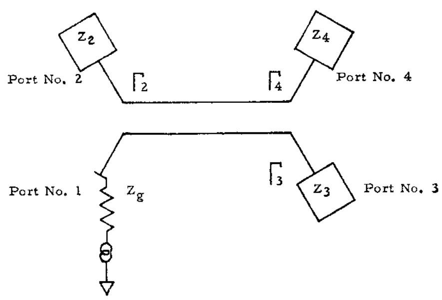


Fig. 1—Schematic representation of 3-db hybrid.

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

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

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

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

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

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

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

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

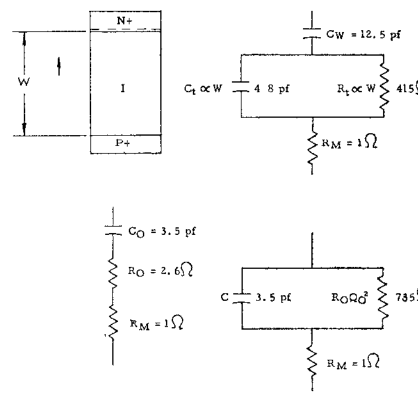
$$L = \sqrt{D\tau}$$

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.



$$C_O = \frac{\epsilon A}{t} \qquad R_O = \frac{t - W}{A\omega^2\epsilon^2\rho} + R_M$$

$$Q_O = \frac{\omega\epsilon\rho}{1 - \frac{1}{t}(W - \omega^2\epsilon^2 A\rho R_M)} \qquad R_O Q_O^2 = \frac{\rho\frac{t}{A}}{1 - \frac{1}{t}(W - \omega^2\epsilon^2 A\rho R_M)}$$

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.