the empty waveguide into the ferrite loaded region of the waveguide, its transverse electric field distribution becomes different. In particular, Fig. 2 illustrates the effect of a finite sample. The transverse electric field distribution measured a short distance after the incident wave has entered the ferrite section (marked Position 1) indicates one mode-type or combination of modes whereas further on along the sample, the transverse electric field distribution (marked Position 2) is of another mode-type. Figs. 1–4 are for various thicknesses of ferrite slab. The frequency is 9.275 kmc.

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Feeding RF Power From a Self-Excited, Pulsed Source into a High-Q Resonant Load*

A problem frequently arising in microwave electronics is the feeding of pulsed power from a self-excited source into a high-Q resonant load. A typical example is the one of feeding power from a pulsed magnetron into a high-Q microwave cavity[1] such as that used in a linear accelerator. In the past, it has been customary to use a stabilizing load in a series tee system[2] which results in approximately half the magnetron power being fed into the high-Q cavity.

In this letter, a high-power ferrite isolator system is described which is capable of feeding 78 per cent of the available power from a 5586, S-band megawatt magnetron into a microwave cavity having an unloaded Q of 14,400. The performance of this ferrite system is compared to the series tee and shown to result in 38 per cent increase in power with no reduction in stability characteristics.

The experimental setup used is shown in Fig. 1. The microwave power from the

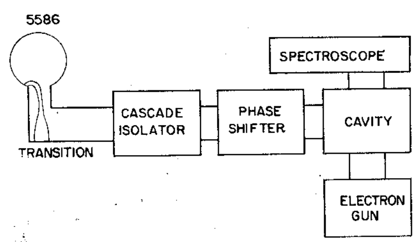

Fig. 1.

5586, S-band magnetron is fed through a coax-to-waveguide transition, then through a Cascade S-159 isolator and phase shifter into a TM_{010} mode cavity. The cavity was resonant at 2840 mc. It had an unloaded Q of 14,400, a shunt resistance of 3.32 megohms, and a VSWR on resonance of 1.2 (coupling greater than critical).

The power delivered by the magnetron to the cavity was determined by accelerating a 20 kv electron beam injected into the system by means of an electron gun and then measuring the exit electron momentum with a spectroscope. The electron momentum is determined by the peak axial field which in turn is determined by the power delivered to the cavity.

Input power to the magnetron was adjusted so that, for a magnetic field of 2650 gauss, 50 amperes peak anode current was delivered for each of the load conditions. The voltage pulse was obtained from a 3-μsec line-type modulator operating at 60 cps repetition rate.

For tee stabilization, the required voltage input was 26 kv peak or 1.30 megw peak input power. For this case, the loaded VSWR was 1.8 resulting in an output power of 453 kw peak of which 233 kw peak, or 51.4 per cent, was delivered to the cavity. The peak axial electric field was 311 kv/cm as determined from a measured exit momentum of 3940 gauss-cm.

For isolator stabilization, the required voltage input was 25.8 kv peak or 1.29 megw peak input power. For this case, the loaded VSWR was 1.11 resulting in an output power of 410 kw peak of which 321 kw peak, or 78.3 per cent, was delivered to the cavity. The peak axial electric field was 365 kv/cm as determined from a measured exit momentum of 4550 gauss-cm.

Thus, relative to the power delivered to the cavity by the tee-stabilized system, isolator stabilization permitted an increase in cavity power of 37.8 per cent with no deterioration of stability in performance. This increase was possible even though the shift in operating point caused by the mismatch of the tee required a higher power output from the magnetron. Hence the increase in cavity power was obtained along with an improvement in operating conditions resulting from the lower VSWR.

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* Received by the PGMTT, February 27, 1959.

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An Image Line Coupler*

For many microwave transmission systems, it is possible to make use of high-conductivity "image planes" as ground planes or as shields. Strip transmission lines, image plane lines, and dielectric image lines are examples of such systems. In order to make use of both sides of the ground plane, energy can be coupled from one side of the plane to the other by means of holes in the metal plate. Obviously, many existing coupling devices can be redesigned for such use. One such application is the directional coupler. In this paper, it is applied to the dielectric image line.

The dielectric image line[1–3] has been used for certain applications.[4,5] This line consists of a half round dielectric rod mounted on an image plane, (see Fig. 1).

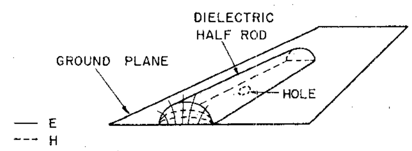

Fig. 1—Dielectric image line.

Now if a hole is made between the two sides of the image plane, coupling exists between the two sides. The purpose of this paper is to indicate what effect this coupling produces. According to H. A. Bethe,[6] the coupling may be determined approximately by considering it as arising from coupling by two dipoles. The electric field on the secondary side of the plane (near the hole) is similar to that generated by an oscillating electric dipole, with its dipole moment parallel to the electric field of the incident wave (near the hole) on the primary side of the plane. The magnetic field behaves as if the hole contained a magnetic dipole moment parallel to the incident magnetic field but in the opposite direction. Making the usual approximations as to aperture size, thickness, and extent of the image plane the following equations can be obtained:

Coupling:

$$C = 20 \log_{10} \frac{\pi h^3}{12\lambda_0 S}\left[\frac{1}{\eta} E_n^2 \frac{F_e(t)}{F_H(t)} + 2\eta H_t^2 \cos\theta\right] F_H(t) \text{ db.}$$

Directivity:

$$D = 20 \log_{10} \frac{\left[2\eta H_t^2 \cos\theta + \frac{1}{\eta} E_n^2 \frac{F_e(t)}{F_H(t)}\right]}{\left[2\eta H_t^2 \cos\theta - \frac{1}{\eta} E_n^2 \frac{F_e(t)}{F_H(t)}\right]} \text{ db.}$$

For maximum directivity:

$$\cos\theta = \frac{E_n^2}{2\eta^2 H_t^2} \cdot \frac{F_e(t)}{F_H(t)}.$$

Here, $F_e(t)$ and $F_H(t)$ are attenuation factors for the electric and magnetic fields, respectively in propagating through the circular

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[5] B. Packer and D. L. Angelakos, "An Image Line Coupler," Univ. of Calif., Berkeley, Inst. of Engrg. Res., Rep. No. 188, series no. 60; July, 1957.

[6] H. A. Bethe, "Lumped Constants for Small Irises," Mass. Inst. Tech., Cambridge, Rad. Lab. Rep. No. 43-22; March, 1943.

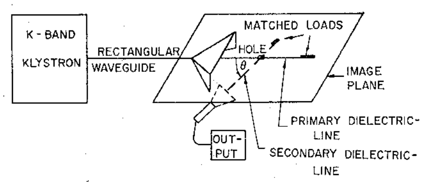

Fig. 2—Experimental set-up.

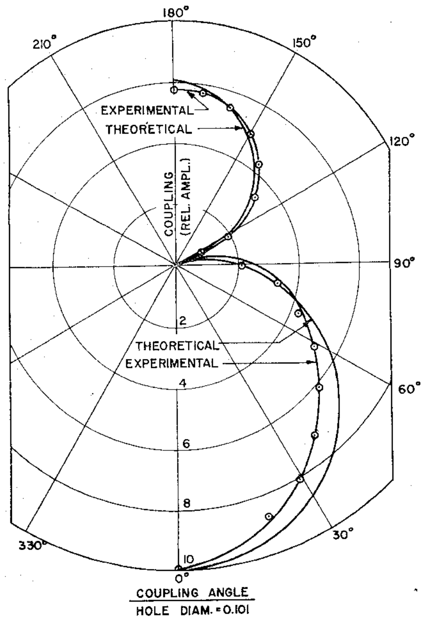

Fig. 3—Coupling as a function of angle θ.

hole acting as a waveguide beyond cut-off. E_n is the normal component of the electric field and H_t is the transverse component of the magnetic field both in the dielectric guide at the coupling hole. The longitudinal component of the magnetic field H_z is neglected. S is a power normalizing factor, h is the diameter of the hole, λ_0 is the free-space wavelength, and η is the specific impedance of free space. Fig. 2 defines the angle θ. If t, the thickness of the plane, is very small, the ratio $F_e(t)/F_H(t) \sim 1$ and the equations further simplify

$$C = T[1 + g \cos \theta]$$

and

$$D = \frac{1 + g \cos \theta}{1 - g \cos \theta}.$$

C and D are given in relative values, T is a constant of proportionality, and g is approximately the ratio of magnetic to electric coupling.

An experimental dielectric image line coupler was investigated at 24.4 kmcps. Fig. 2 illustrates the arrangement of the experiment. The ground plane thickness was 0.026 inch and the hole size was 0.101 inch in diameter. For these dimensions, $F_H(t)$ was 1.112 and $F_e(t)$ was 0.827. The ratio $F_e(t)/F_H(t)$ then is 0.742 and should be used (instead of 1.00) if greater accuracy is called for. Fig. 3 shows the variation of the coupling as a function of the angle θ. Plotted in the same figure is a curve called "Theoretical" which is of the form: $C = T(1 + g \cos \theta)$. The factor g is approximately equal to 4, the ratio of magnetic to electric coupling. The magnitude of the coupling is normalized so that at $\theta = 0°$, $C = 10$. Similar data were taken for hole diameters of 0.078, 0.082, 0.093, 0.111, 0.128 inch with no substantial differences apparent.

Fig. 4 illustrates how this arrangement may be used as a directional coupler. The

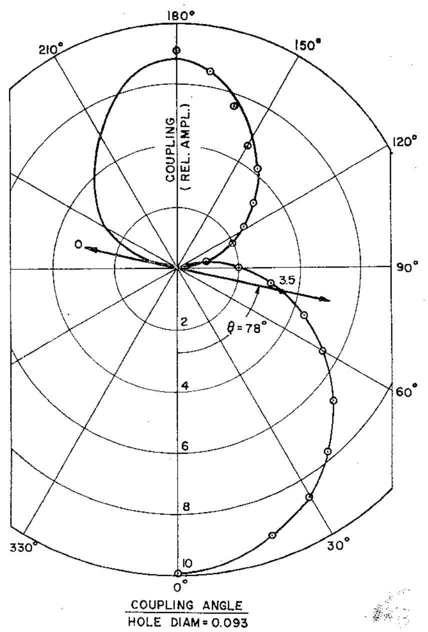

Fig. 4—Direct coupling.

solid line at $\theta = 78°$ represents the orientation of the secondary guide with respect to the primary guide. In one direction, the coupling is approximately zero whereas in the other direction it is 3.5 relative units.

Many other possible arrangements of holes and slots may be used to produce equivalent results or to improve the coupling and directivity behavior.

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An Extension of the Concept of Stop and Pass Bands of a Zobel Type Filter to a General Reciprocal Two Port Network Which has a Nonloxodromic Transformation*

The conventional treatment[1] of the Zobel filter starts with symmetrical T or π sections of pure reactances and then develops the iterative measures of the network; the fixed points and the propagation constant. It is shown that the propagation constant is either pure real (stop band)[2] or pure imaginary (pass band). These iterative measures can be worked out for the general T section. Fig. 1 shows the nomenclature used for the symmetrical T section and the general two port.

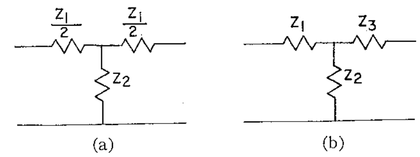

Fig. 1—(a) The symmetrical T section, (b) the asymmetrical T section.

TABLE I

(a) The symmetrical T section
$\cosh \gamma = 1 + \frac{Z_1}{2Z_2}$
$Z_{it} = Z_0 = \pm \sqrt{Z_1 Z_2 + \frac{Z_1^2}{4}}$
(b) The asymmetrical T section
$\cosh \gamma = 1 + \frac{Z_1 + Z_3}{2Z_2}$
$Z_{it} = Z_1 - Z_3 \pm \sqrt{\left(\frac{Z_1 + Z_3}{2}\right)^2 + (Z_1 + Z_3)Z_2}$

Table 1 lists the equations for the fixed points and the propagation constant for both networks. The network properties can also be developed as a bilinear transformation,[3] and the network can be classified by its type of transformation. For the two port, the input impedance is related to the output impedance by

$$Z' = \frac{\frac{(Z_1+Z_2)Z}{Z_2} + \frac{Z_1Z_2+Z_2Z_3+Z_3Z_1}{Z_2}}{\frac{1}{Z_2}Z + \frac{Z_2+Z_3}{Z_2}}.$$

The trace of the normalized transformation is $2 + (Z_1+Z_3)/Z_2$ which is twice $\cosh \gamma$. For the transformation to be nonloxodromic the trace must be real, hence the sum of the two series impedance phasors is either in phase or 180° out of phase with the phasor impedance of the shunt arm. It has been shown[4] that it is always possible in a microwave two port, to find reference planes at which the transformation is nonloxodromic. Thusly for any microwave network, reference planes can be found where $\cosh \gamma$ is real. If the transformation is hyperbolic ($|a+d| > 2$)γ will be real. If the transformation is elliptic ($|a+d| < 2$)γ will be imagi-

* Received by the PGMTT, March 9, 1959.

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[3] E. F. Bolinder, "Impedance and polarization-ratio transformations by a graphical method using isometric circles," IRE Trans. on Microwave Theory and Techniques, vol. MTT-4, pp. 176–180; July, 1956.

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