junctions are made with silver solder. The temperature necessary for silver soldering limits the thinness of material to be worked. In both type cavities, the sections with the terminations were chosen long enough so that the fields in these sections of "beyond cutoff" waveguide will be attenuated adequately and hence will not radiate. Some radiation from the cylindrical cavity occurred because the number of cross-plates was insufficient to prevent the propagation of modes in the outer portion.

Each of the cavities of both types considered was originally fabricated with the terminations left unsoldered and held in place only by friction (increased by adding external clamps). This was done so that the cavities might be tuned to a reasonable frequency by sliding the terminations in and out. The frequency of each design was near 9435 mcps, the nominal frequency of Crain refractometer measuring cavities.

After the terminations were properly placed and soldered, measurements of the voltage standing wave ratio (VSWR) due to cavity input impedance were made as a function of frequency. Since there is no convenient means of calculating the size of feed hole which gives critical coupling, feed holes were cut small and gradually enlarged until the plot of VSWR vs frequency showed a minimum value near unity.

Following the technique given by Montgomery,[8] the unloaded Q of these cavities was calculated to be 3420 for the rectangular cavity and 2310 for the cylindrical cavity.

The next pertinent test was to determine the relation between cavity Q and the axial length of the terminations. When the length of termination sections on the rectangular cavities was decreased by increments, it was found that the Q was essentially constant for terminating section lengths greater than two inches but dropped rapidly as the length of the stub was decreased below two inches.

Since physical dimensions, as well as index of refraction, determine the resonant frequency of a cavity, the coefficient of thermal expansion of the material of the cavity will cause an erroneous indication of change in index unless the proper temperature correction is known. A possible means of temperature compensation involves the use of a material with a larger temperature coefficient for the divider in the terminating stub than for the material in the body of the cavity, but extensive efforts in this direction were not carried out.

The rectangular cavity must be considered a satisfactory, working piece of equipment with sufficiently good characteristics to make it acceptable for its designed function as a governing cavity for a Pound oscillator. The cross sectional obstruction due to the termination is only 1.66 per cent of the waveguide cross section area. With sufficient effort, it should be possible to temperature compensate such a cavity to any degree required.

D. C. Thorn
A. W. Straiton
Elec. Eng. Res. Lab.
Univ. of Texas
Austin, Tex.

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[8] C. G. Montgomery, ed., "Technique of Microwave Measurements," Mass. Inst. Tech., vol. 11, McGraw-Hill Book Co., Inc., New York, N. Y.

Transverse Electric Field Distributions in Ferrite Loaded Waveguides*

The transverse electric field distribution in dielectric and ferrite loaded waveguides has been measured by several investigators.[1,2] Knowledge of the actual field distribution within the waveguide is needed in the design of field displacement isolators, phase-shifters, and similar microwave devices. In making such measurements, care must be taken to choose lengths of samples of ferrites or dielectrics sufficiently long enough so that the distribution in one transverse plane will be the same (except for an attenuation effect) as in another transverse plane.

To show that variations in distribution are indeed present, measurements were made with a transverse electric field detector[1,2] at various positions along a ferrite slab (see the sketch in the lower half of Fig. 1). As the incident wave penetrates from

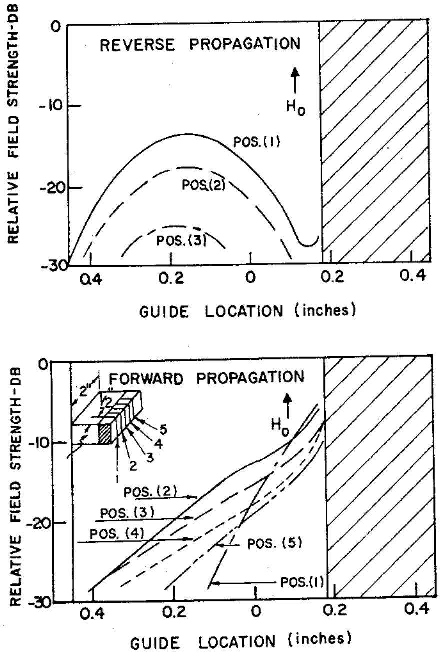

Fig. 1—Electric field distribution at 0.250 inch intervals along ferrite slab 0.259 inch thick with an external dc field of 2000 gauss.

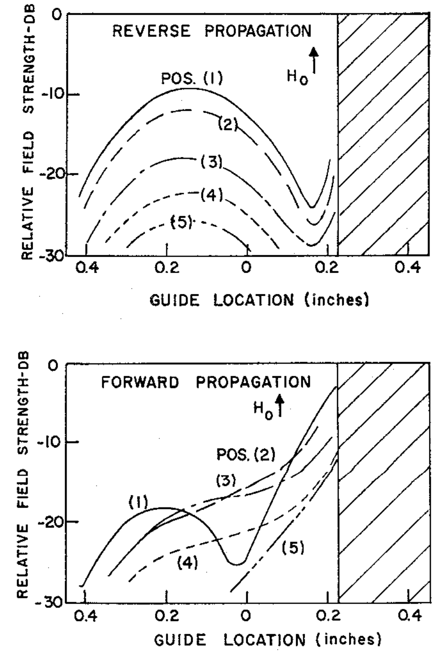

Fig. 2—Electric field distribution at 0.250 inch intervals along ferrite slab 0.227 inch thick with $H_0=2000$ gauss.

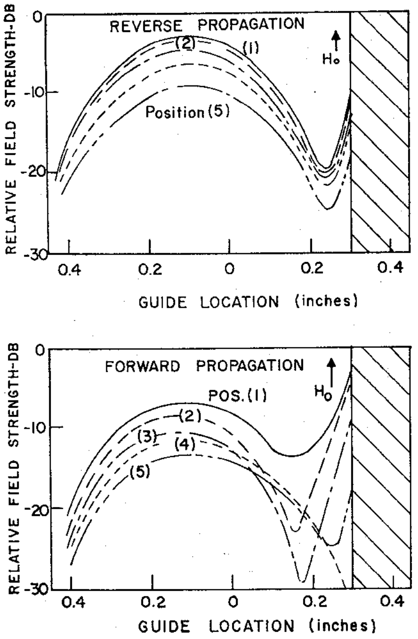

Fig. 3—Electric field distribution at 0.250 inch intervals along ferrite slab 0.145 inch thick with $H_0=2000$ gauss.

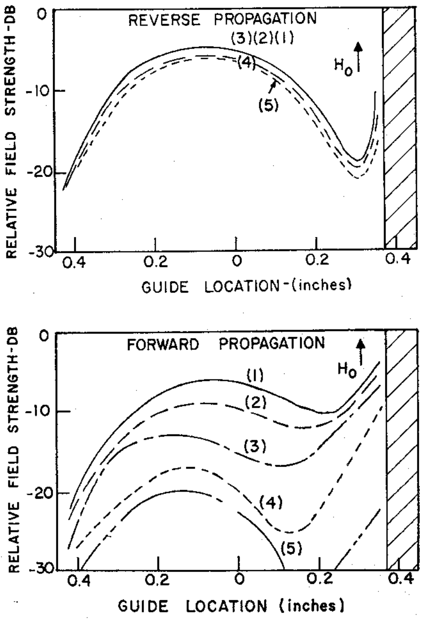

Fig. 4—Electric field distribution at 0.250 inch interval along ferrite slab 0.081 inch thick with $H_0=2000$ gauss.

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[1] R. L. Comstock, D. J. Angelakos and A. Johnson. "Determination of Fields in a Ferrite-Loaded Waveguide," Elec. Res. Lab., Univ. of California, Series 60, Issue 186; 1957.

[2] T. M. Straus, 1958 IRE Wescon Convention Record, pt. I.

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.

D. J. Angelakos
Elec. Engrg. Dept.
Univ. of Calif.
Berkeley 4, Calif.

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 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).

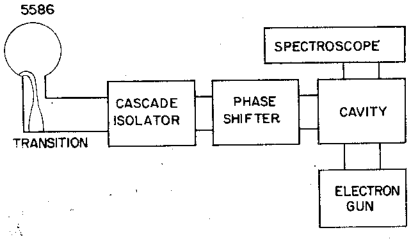

Fig. 1.

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.

H. A. Spuhler
R. J. Kenyon
P. D. Coleman
Dept. of Elec. Engrg.
Univ. of Illinois
Urbana, Ill.

* Received by the PGMTT, February 27, 1959.

[1] G. B. Collins, "Microwave Magnetrons," Rad. Lab. Series, vol. 6, pp. 638–639; McGraw-Hill Book Co., Inc., New York, N. Y.; 1958.

[2] I. Kaufman and P. D. Coleman, *J. Appl. Phys.*, vol. 28, pp. 936–944; September, 1957.

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).

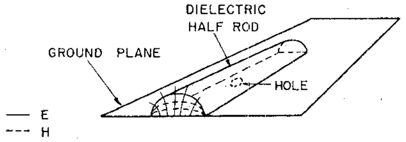

Fig. 1—Dielectric image line.

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:

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

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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[1] D. D. King, "Circuit components in dielectric image lines," IRE Trans. on Microwave Theory and Techniques," vol. MTT-3, pp. 35–39; December, 1955.

[2] D. D. King, "Properties of dielectric image lines," IRE Trans. on Microwave Theory and Techniques," vol. MTT-3, pp. 75–81; March, 1955.

[3] S. P. Schlesinger and D. D. King, "Dielectric image lines," IRE Trans. on Microwave Theory and Techniques, vol. MTT-6, pp. 291–299; July, 1958.

[4] H. W. Cooper, M. Hoffman, and S. Isacson, "Image line surface wave antenna," 1958 IRE National Convention Record, pt. 1.

[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.