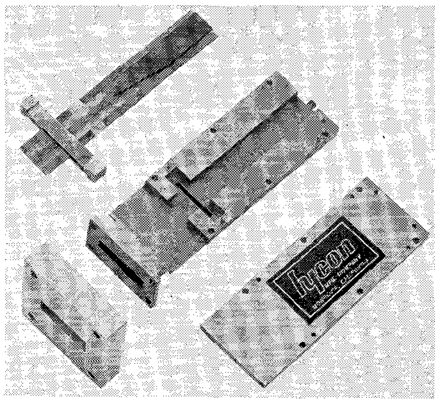

Fig. 1—Miniature strip-line to waveguide slot adapter—inline.

TABLE

Center Frequency (GC)	Bandwidth for VSWR ≤1.20 (Mc)	(per cent)	Data Plotted on Fig. 2
9.0	960	10	Curve 1
9.4	780	8	Curve 2
9.8	1180	12	Curve 3
10.2	700	7	Curve 4

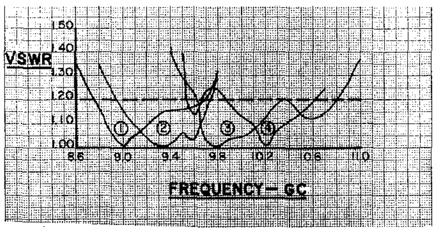

Fig. 2—Input VSWR vs frequency.

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Addendum to "Discontinuities in the Center Conductor of Symmetric Strip Transmission Line"*

It has been pointed out to the authors that the formulas contained in their paper[1] refer only to air-filled and not to dielectric-filled strip transmission lines. This is indeed the case and constitutes an omission which is now rectified by sketching the simple extension to the dielectric-filled case.

All formulas in the paper, save that for the characteristic impedance, remain completely unaltered in the dielectric-filled case provided λ is understood to denote the wavelength in the dielectric-filled strip-line; *i.e.*, $\lambda=\lambda_0/\sqrt{\epsilon'}$, where λ_0 is in air-filled strip-line and $\sqrt{\epsilon'}$ is the relative dielectric constant. In dielectric-filled line κ then is $2\pi/\lambda$. In the dielectric case, the characteristic impedance Z_0 must be computed from

$$Z_0=\frac{1}{\sqrt{\epsilon'}}\,\frac{30\left(1-\frac{t}{b}\right)}{D/b}.$$

It should be kept in mind that the results for normalized reactance and susceptance network elements, such as X_a' or B_b', are now normalized with respect to Z_0 (or $Y_0=1/Z_0$) as just defined. The equivalent strip width D and other equivalent dimensions do not depend on the dielectric constant.

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* Received by the PGMTT, November 20, 1961.

[1] H. M. Altschuler and A. A. Oliner, "Discontinuities in the center conductor of symmetric transmission line," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 328–339; May, 1960.

A Simple Method for Measuring the Phase Shift and Attenuation through Active Microwave Networks*

Most methods used to measure the phase shift and attenuation through a microwave network require that the network be matched to the waveguide or the VSWR must be measured. In passive networks, these are simple and straightforward operations. However, in some active networks, principally networks which are time/temperature dependent, these operations become very tedious and time consuming. The method illustrated here can measure the phase shift and attenuation simultaneously, directly, independently, eliminates the inconvenience of measuring the VSWR, or matching the network to the waveguide and consists of generally available microwave equipment.

The physical setup consists of a conventional phase-shift loop (interferometer) and a loosely coupled "attenuation" loop. The loops are illustrated in Fig. 1.

The function of the attenuation loop is to take into account the reflected power so that the attenuation of the unknown network may be read directly.

The loop accomplishes this by adding, in phase, part of the power output and part of the power reflected by the unknown network. The sum of these powers will remain constant unless there is attenuation by the network. When attenuation occurs, it will appear as a decrease in the sum of the reflected power and the power output of the network. The attenuation of the network may then be directly read by increasing the power input to the attenuation loop until the original sum is reached. The reflections may occur at any discontinuity or set of discontinuities in the network; therefore, an

* Received by the PGMTT, November 27, 1961.

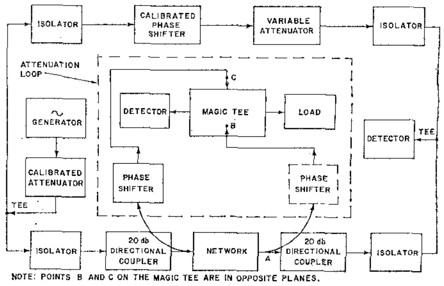

Fig. 1—Phase shift and attenuation loops.

adjustable phase shifter capable of 360° phase shift must be employed (in either arm of the attenuation loop) and tuned for detector maximum in the attenuation loop for each attenuation measurement. The directional couplers (20 db or more) must be balanced or some attenuation must be added to the more tightly coupled coupler in order to maintain the accuracy of the system. The system may be checked out by inserting an adjustable short in the output (at point A) of the network and insuring that there is only very small variation in the output at the attenuation loop detector. Care must be taken to insure that the network is not radiating as this method will not take radiation into account. The attenuation of the network is read directly by the calibrated attenuator external to the loops.

With an X-band power source of 10 mw the dynamic range is approximately 20 db. If greater range is needed, low-noise microwave amplifiers may be used in place of the crystal detectors or the power source may be increased.

This method is currently being used to measure phase shift and attenuation through gas discharge tubes at X-band and it has greatly simplified these measurements.

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Design of Interstage Coupling Apertures for Narrow-Band Tunable Coaxial Band-Pass Filters*

The design technique to be described in this note is applicable to tunable coaxial band-pass filters having narrow bandwidths (*i.e.*, less than 10 per cent). In the frequency range of 1500 Mc to about 10,000 Mc, coaxial band-pass filters usually employ coupled $\lambda/4$ resonant cavities. Unlike direct coupled waveguide band-pass filters which are often amenable to a complete paper design,[1] these coaxial band-pass filters require

* Received by the PGMTT, November 27, 1961.

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

empirical developmental methods to supplement the filter theory customarily employed. The purpose of this note is to present a technique for simplifying the empirical development.

The electrical performance of a fixed tuned coaxial band-pass filter is completely determined by the resonant frequency of the cavities, the unloaded Q's of the cavities, the input/output couplings to specified impedance levels, and the interstage couplings.[2] For multiresonator filters having unloaded Q's an order of magnitude greater than the minimum unloaded Q's required to realize the desired filter response shape, the bandwidth of the filter is primarily determined by the interstage couplings. To design a tunable band-pass filter with constant bandwidth, the frequency sensitivities of the interstage couplings must be controlled.

The electrical performance of an interstage coupling mechanism can be described quantitatively by a coefficient of coupling. This coefficient of coupling can be conveniently measured by known techniques.[3] It can be shown that

$$K \cong \frac{\Delta f}{f_0} \quad \text{at any frequency } f_0,$$

where

K = coefficient of coupling,
Δf = coupling bandwidth.

If the absolute bandwidth (not per cent bandwidth) of the tunable filter must be kept constant at all frequencies, then K must vary inversely with frequency and Δf must be independent of frequency.

A convenient mechanism for achieving interstage coupling at microwave frequencies is an aperture. Apertures are often preferable to probes, loops, or direct connections since no soldering is required and apertures can be readily machined to precision tolerances. The most convenient aperture is circular since it is easiest to fabricate. It can be shown that for small lossless circular apertures in thin walls

$$\Delta f \cong Cf^2 \left[2\cos^2\left(\frac{2\pi fL}{v}\right) - \sin^2\left(\frac{2\pi fL}{v}\right)\right],$$

where

Δf = coupling bandwidth
f = frequency and C is a constant,
L = distance from centerline of aperture to plane of short-circuit,
v = velocity of propagation in free space.

Taking $d(\Delta f)/df$ and equating to zero,

$$\cos^2\left(\frac{2\pi f_m L}{v}\right) - \left(\frac{2\pi f_m L}{v}\right)\cos\left(\frac{2\pi f_m L}{v}\right) \cdot \sin\left(\frac{2\pi f_m L}{v}\right) = \frac{1}{3},$$

where f_m = frequency of maximum coupling. Letting

$$\theta = \frac{2\pi f_m L}{v}, \qquad \cos^2\theta - \theta\cos\theta\sin\theta = \tfrac{1}{3}.$$

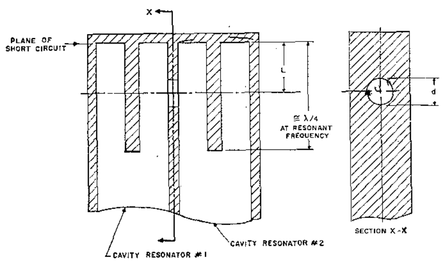

Fig. 1—Coaxial band-pass filter interstage coupling.

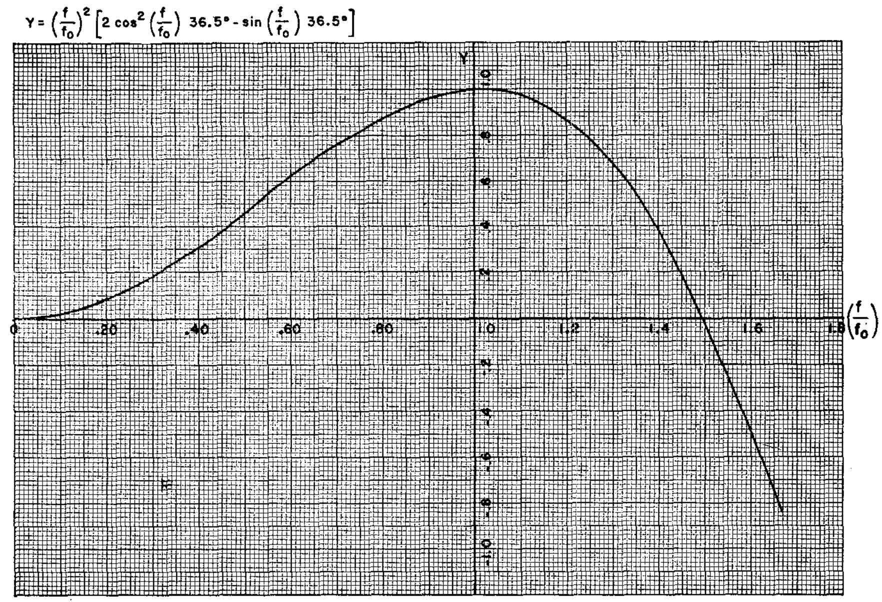

Fig. 2—Frequency sensitivity of circular aperture coupling bandwidth.

This is a transcendental equation which must be solved graphically. It can be shown that for maximum coupling bandwidth $\theta \cong 36°$.

To design a tunable coaxial band-pass filter using circular apertures as interstage coupling mechanisms, it is necessary to specify both the aperture location (L) and the aperture diameters (d) to determine the coupling bandwidth. (See Fig. 1.) Using the theory previously developed, L is chosen to make $\theta = 36$ electrical degrees at the center frequency of the tuning range. This reduces the design problem to that of determining the aperture diameter. d is best determined empirically and it is a relatively simple matter to obtain a plot of Δf vs d by measuring the coefficients of coupling at the center of the tuning range. Aperture diameters are then chosen to realize the coupling bandwidths required for the particular filter response shape that is desired.

The technique of locating circular interstage coupling apertures in tunable coaxial band-pass filters for maximum coupling bandwidths (*i.e.*, minimum frequency sensitivity of coupling bandwidth) has been successfully implemented for various filters. These filters have employed $\lambda/4$ resonators with cylindrical center conductors coaxial to either cylindrical, square, or rectangular outer conductors. The technique has worked for filters using both contacting fingers and noncontacting shorts.

A plot of $Y = \left(\frac{f}{f_0}\right)^2 \left[2\cos^2\left(\frac{f}{f_0}\right)36.5° - \sin^2\left(\frac{f}{f_0}\right)36.5°\right]$ is shown in Fig. 2. In practice, the aperture centerline is usually located between 36.5 and 40.5 electrical degrees from the plane of the short circuit. It can be seen that Y is not a symmetrical function of (f/f_0) and that the high end of the tuning range becomes more frequency sensitive when tuning over wide ranges. The simple theory presented also neglects open-end fringing capacitance and the fact that contacting fingers make the resonators compound transmission lines. The theory is most valid when the coupling bandwidth varies as the cube of the aperture diameters. It is less applicable for very small apertures and/or thick walls which cause appreciable attenuation through the apertures. It also breaks down when the aperture diameter approaches the size of the cavity outer conductor.

The theory presented can also be extended to noncircular apertures, loop couplings, and/or design for constant percentage bandwidth filters. It is highly recommended as a means of substantially reducing the design/development cycle required for tunable coaxial band-pass filters.

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[2] M. Dishal, "Dissipative band-pass filters," PROC. IRE, vol. 37, pp. 1050–1069; September, 1949.
[3] M. Dishal, "Alignment and adjustment of synchronously tuned multiple-resonator-circuit filters," PROC. IRE, vol. 39, pp. 1448–1455; November, 1951.