Correspondence

VHF Slow-Wave Slotted Line*

Introduction

A slotted line is generally used for voltage-standing-wave-ratio and impedance measurements in the microwave region. Because a slotted line operable down to 25 Mc would be from 10 to 15 feet long, it is impractical for general laboratory use, and is difficult and expensive to construct with the necessary mechanical tolerances.

A new VHF slotted line has been designed using a slow-wave structure technique; the line was developed to measure the VSWR and the impedance of components in the frequency range from 25 to 200 Mc. Because a slow-wave technique is used in the design, the new slotted line is only 3 feet long whereas a comparable conventional coaxial slotted line would be 24 feet long.

The VHF slow-wave slotted line consists of conductor bent into a serpentine configuration and placed over a conducting ground plane (Fig. 1). A fast fundamental wave travels along the serpentine line with the velocity of light and a slow wave travels along the axis of the line with a much reduced velocity. If the line is terminated with a load other than its characteristic impedance, a standing wave will exist along the line. A probe moving along the axial direction of the serpentine line will measure a voltage standing wave with a wavelength much shorter than that of free-space wavelength.

The ratio of the free-space wavelength, λ_0, and the axial serpentine line wavelength λ_s is termed the wave-retardation factor and is approximately

$$\frac{\lambda_0}{\lambda_s} = \sqrt{\epsilon} \csc \psi = \frac{2L}{P} \sqrt{\epsilon}.$$

The characteristic impedance of the serpentine line above a conducting plane can be calculated as the impedance of a straight wire over an infinite ground plane. If the spacing between the serpentine line and the plane is small compared to the spacing between adjacent legs of the line, the error due to increased distributed capacitance is negligible. Therefore, the characteristic impedance Z_0 is

$$Z_0 = \frac{60}{\sqrt{\epsilon}} \cosh^{-1} \frac{2h}{d},$$

where:

ϵ = composite dielectric constant of the materials surrounding the wire,
h = distance between the center of the wire and the ground plane,
d = diameter of the wire.

Experimental Model

The experimental model of the slow-wave slotted line is shown in Fig. 2. The serpentine line was formed with $\frac{3}{16}$-inch aluminum rod and is housed in an aluminum casing. A slow-wave with a wave retardation factor of about 8.3 was measured along the axial direction of the serpentine line. Measurements with a coaxial sliding short at the load end of the slotted line show that the slow-wave standing-wave minimum moves with that of the sliding short at the rate of the wave-retardation factor. The over-all length of the line is three feet. The specifications are as follows:

Frequency range	25 Mc to 200 Mc
Characteristic impedance	50±1 ohms
Wave-retardation factor	8.3
Residual VSWR	1.05 max
Short-circuit VSWR	100:1 min
Connectors	Type N
Detector element	1N21
Tuner	Lumped LC circuit
Dimensions	$3\frac{3}{4} \times 6 \times 36$ inches
Weight	30 pounds

A significant advantage of the VHF slow-wave slotted line is that it is amenable to manufacture by inexpensive laminating and etching processes. The critical dimensional tolerances are the uniformity of the serpentine line and its spacing from the ground plane. The dielectric surrounding the line must be of low-loss material such as teflon or rexolite. The laminating and etching processes make it possible to construct a serpentine line with the necessary mechanical tolerances. Such a slotted line can extend the operating frequency of a 3-foot line from 25 Mc to as much as 1000 Mc. Several sizes of serpentine lines used in the slow-wave slotted line are shown in Fig. 3.

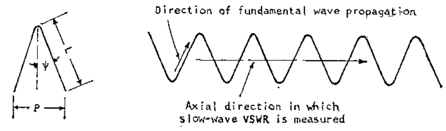

Fig. 1—Wave in a serpentine line.

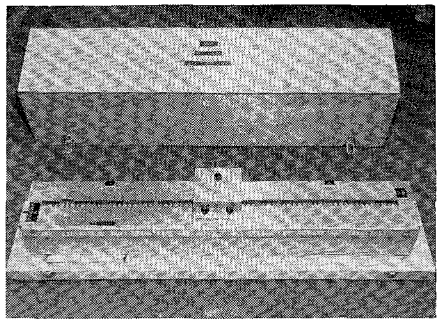

Fig. 2—VHF slow-wave slotted line.

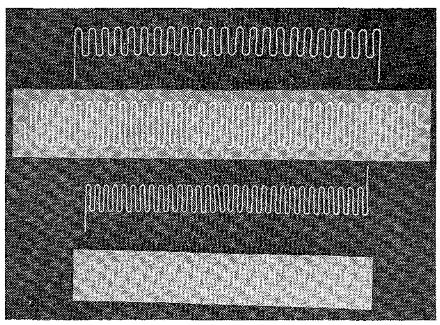

Fig. 3—Serpentine lines.

Conclusion

The operation of slotted lines is simple and straightforward, so that even unskilled technicians can make measurements without difficulties. With the new, compact slow-wave slotted line, VSWR and impedance measurements can now be extended down to 25 Mc. Although impedance measurements at frequencies as low as 25 Mc can be made with an impedance bridge, such null-detecting measurements are usually cumbersome and time consuming.

Luis L. Oh
C. D. Lunden
Transport Div.
Boeing Airplane Co.
Renton, Wash.

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Miniature Strip-Line to Waveguide Slot Adapter—Co-Linear*

An inline adapter from miniature air dielectric strip transmission line with a ground plane spacing of 0.130 in to half height X-band waveguide has been developed by the Microwave Design Group at Hycon Manufacturing Company, Monrovia, Calif. The upper wall of the waveguide was used as the lower ground plane for the stripline. Unity coupling was accomplished through a transverse slot in the common wall. The slot was 0.066-in wide by 0.750-in long surrounded by a metal cavity 0.250 in by 1.000 in built into the ground planes. (See Fig. 1.) The slot junction was matched by use of tunable shorts both in the stripline and the waveguide. The output of the 50-ohm stripline was terminated by a built-in strip-line load whose VSWR was 1.04.

An intermediate height waveguide section was attached to the unit since it was tested on a standard X-band waveguide slotted line. This intermediate section was 0.283-in high, 0.900-in wide, by 0.401-in long.

The slot adapter was tuned for optimum VSWR at four frequencies with the moveable shorts. Test results are tabulated in Table I and plotted in Fig. 2.

While only an X-band slot adapter has been built, similar designs for higher or lower frequency bands are feasible.

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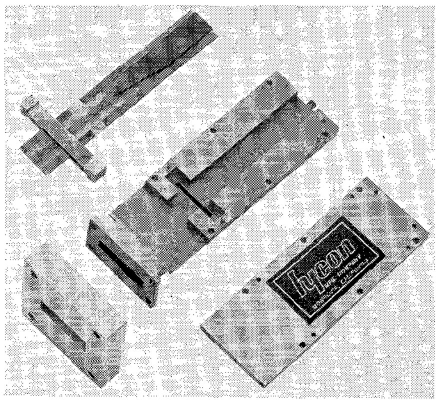

Fig. 1—Miniature strip-line to waveguide slot adapter—inline.

TABLE

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

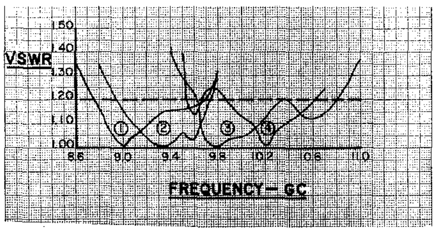

Fig. 2—Input VSWR vs frequency.

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Onnig Shahan
Microwave Design Group
Hycon Manufacturing Co.
Monrovia, Calif.

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.

H. M. Altschuler
A. A. Oliner
Microwave Research Institute
Brooklyn, N. Y.

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[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 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.

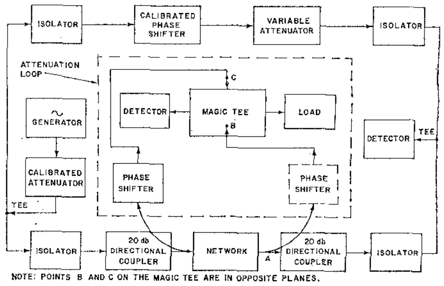

Fig. 1—Phase shift and attenuation loops.

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.

John R. Alday
U. S. Naval Ordnance Lab
Corona, Calif

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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.