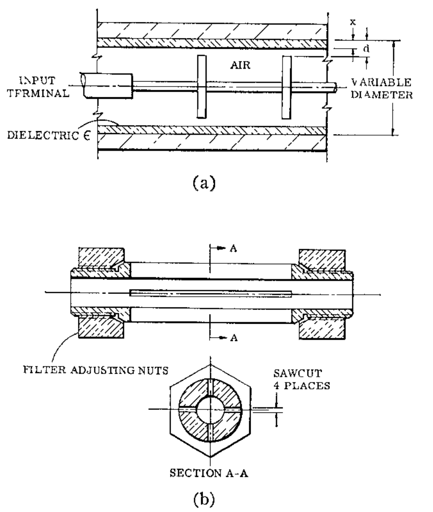

Fig. 2. (a) Open condition of variable cutoff low-pass coaxial filter. (b) Mechanical diagram showing the simplicity of the filter adjustment mechanism.

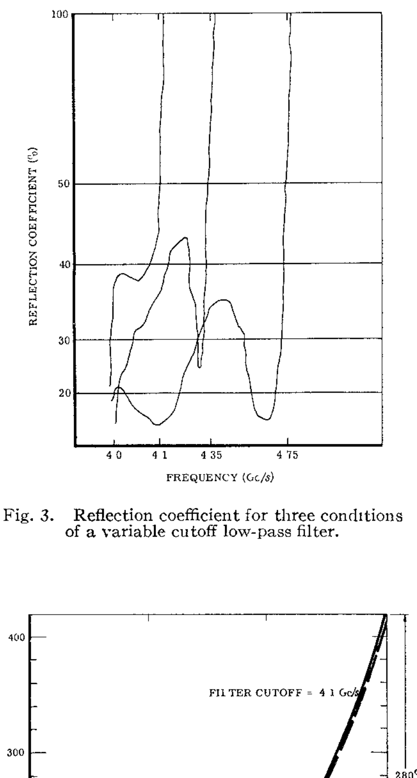

Fig. 3. Reflection coefficient for three conditions of a variable cutoff low-pass filter.

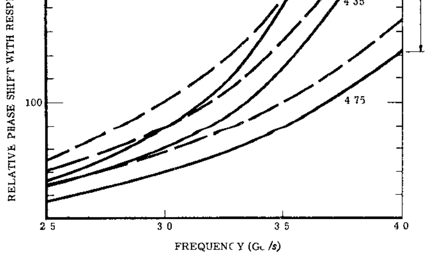

Fig. 4. Correlation of theoretical and measured phase shift data for three conditions of the variable cutoff low-pass filter.

$$\phi = 114.6n\left[\text{Arcsin}\left(\frac{f}{f_c}\right) - \frac{f}{f_c}\right] \text{degrees.} \quad (1)$$

For the case $f=2f_0$ in (1), the change in ϕ at f_0 is assumed negligible because of the frequency separation of f_0 with respect to either f_{c1} or f_{c2}. In almost any application, it must be noted that the frequency of interest must not be selected too close to the filter's cutoff frequency if the reflection coefficient is to be acceptable.

Figure 2(a) exhibits the practical coaxial low-pass network. The smaller center conductor high impedance line sections represent the required series inductance whereas the low impedance ring sections produce the desired amount of shunt capacitance. Upon cutting four equally spaced axial slots in the outer conductor, as shown in Fig. 2(b), this outer diameter was made quite flexible by applying concentric external pressure through the use of adjusting nuts. The dielectric shown was a sheet wrapped inside the outer coax cylinder to partially fill up the gap between the center ring conductors and the outer coaxial shell. Thus, by varying the designated air gap $(d-x)$, the shunt capacitance is made adjustable with a consequent change in the filter cutoff frequency. The ratio of the cutoff frequencies for the two limiting conditions (fully open and fully closed) can be approximately equated as,

$$\frac{f_c(\text{open})}{f_c(\text{closed})} \cong \sqrt{\frac{\epsilon}{1+\frac{x}{d}(\epsilon-1)}}. \quad (2)$$

Normally, the lowest cutoff frequency f_c (closed) is initially determined when designing the filter network. Then the upper cutoff frequency f_c (open) may be obtained from (2) which should be good for ratios of f_c (open) to f_c (closed) up to at least 1.25.

An S-band harmonic phase delay network was designed to exhibit a maximum cutoff variation from about 4.0 to 4.8 Gc/s through the use of (2). The dielectric material used in this ten section filter was Aclar, which has a relative dielectric constant of approximately 2.5. Three of the cutoff conditions are shown in Fig. 3 as a function of the reflection coefficient characteristics from 4 to 5 Gc/s. The reflection coefficient was under 15 percent at all frequencies below 3.9 Gc/s for each cutoff condition shown.

Correlation of the theoretical with measured relative phase shift data for the three cutoff conditions is graphically related in Fig. 4. The theoretical phase shift was determined by (1). Considering that small mismatches occurred over the band, 2.5 to 4.0 Gc/s, each correlation is quite good. Thus, a continuously controllable difference in phase shift of about 280°, at 4 Gc/s with respect to 2 Gc/s, was observed in this simple coaxial configuration which proved to be adequate for a particular circuit application. However, it should be observed that a 360° variation could have readily been realized by increasing the number of filter sections. In addition, it should also be noted that the frequencies of concern do not necessarily have to be harmonically related. However, the closer the frequency separation, the smaller will be the relative phase difference in any given filter. If both frequencies of concern were closely spaced near the filter's cutoff frequency, then, the actual phase shift or phase delay would be the relative angle difference $\phi_2-\phi_1$.

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A Semi-Automatic Technique for Tuning a Reflectometer

Introduction

Tuned microwave reflectometers are used in single-frequency high-precision microwave measurement applications. The tuners are adjusted such that a constant detector output is achieved when the reflectometer is terminated by loads of constant reflection coefficient magnitude and variable phase. At present, the phase variation is achieved by sliding the loads inside a terminating section of uniform waveguide.

The tuning procedure is accomplished in two steps: 1) tuner x, as shown in Fig. 1, is adjusted for no variation in the detected output as a low reflection load is moved inside the terminating waveguide, and 2) tuner y is adjusted for no output variation as a short-circuit is moved inside this waveguide [1]–[3].

This correspondence describes a method which substantially shortens the time required for tuning the reflectometer by automatically driving the load inside the uniform waveguide and synchronously displaying the tuning response on an oscilloscope. The application of this method to existing microwave systems employing the reflectometer technique requires the utilization of additional equipment, which is commercially available, except for one item, the carriage.

Description of Equipment Used

Figure 1 shows the simplified diagram of the reflectometer in the left part and the added equipment required for the application of this technique in the right part.

In the tuning process two types of loads are used, a large reflection and a low reflection termination, as in Fig. 2. Teflon spacers are used to support the loads inside the waveguide and to reduce wear of the guide.

The reciprocating motion of the load is accomplished by attaching the load to a driven carriage. The carriage, shown in Fig. 3, was made in the National Bureau of Standards instrument shop.[1] It converts the oscillating rotational motion of a commercially available "sweep drive" output into reciprocating longitudinal motion. Adjustments are provided on the carriage for alignment.

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[1] This carriage is more elaborate than required but was available and satisfactory for demonstrating the technique.

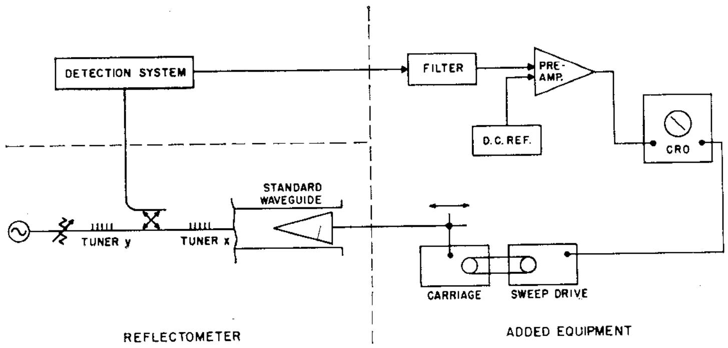

Fig. 1. Block diagram of the reflectometer system using the semi-automatic technique.

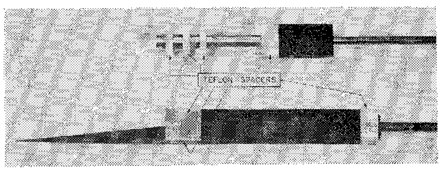

Fig. 2. Large reflection termination (upper) and low reflection termination (lower) used in tuning the reflectometer. Teflon spacers are indicated by arrows.

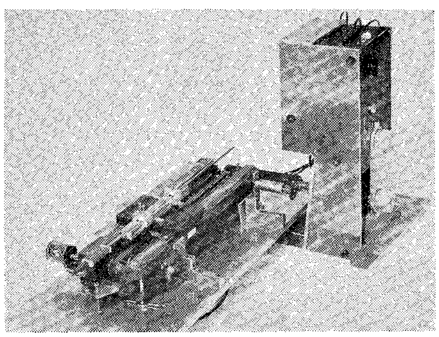

Fig. 3. "Sweep drive" unit (upper right) and carriage (lower left) used to drive the load with reciprocating longitudinal motion inside the waveguide.

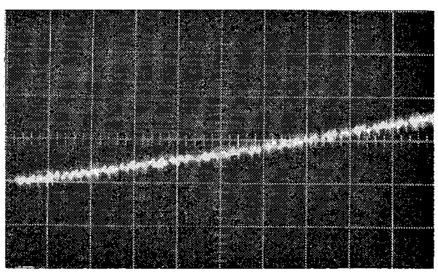

Fig. 4. Reflectometer output vs. position of the sliding short termination. Vertical sensitivity is 0.006 dB/unit and total horizontal displacement is in the neighborhood of $\lambda/2$ at 9.8 GHz.

The "sweep drive" has adjustments to control the center of the sweep, sweep speed (0.5–5 Hz), and sweep arc (30–300°). In addition, it provides the oscilloscope with a horizontal deflection voltage that is proportional to the angle (longitudinal position of load).

The output signal from the detection system should be filtered first and then applied to a "high gain dc differential pre-amplifier." It is especially convenient if this amplifier is a plug-in unit on the oscilloscope. A low-pass *RC* filter is used to restrict the ac variation of the detection system output to very low frequencies. The commercially available "high gain dc differential pre-amplifier" should have provision for use of either input, separately, or both together differentially, either ac- or dc-coupled.

The output response is centered on the oscilloscope by means of adjusting the variable dc reference voltage when the pre-amplifier is in the differential dc-coupled mode of operation.

Brief Description of Tuning Procedure

Preliminary Adjustment of Sweep Drive Unit

1) The driver unit is adjusted such that the reciprocating motion of the load scans a distance of more than one-half guide wavelength. This will assure that more than one cycle of ac variation will be displayed on the oscilloscope.

2) The frequency of the driver unit is adjusted to a sweep rate of about 1 Hz (cycle per second). One uses a slow sweep rate to avoid mechanical vibration, but it must be fast enough to avoid flicker of the scope trace.

Adjustments of the Tuners

1) Tuner x and the variable dc reference voltage are adjusted to decrease the dc level of the signal (as viewed on the scope) when a low reflection load is used and the pre-amplifier is in the differential dc-coupled mode of operation. This will assure that the directivity ratio of the reflectometer is increasing [2].

2) With the pre-amplifier switched to its ac mode and its sensitivity increased, tuner x is adjusted again to minimize the ac variation as viewed on the scope. Minimum variation indicates maximum directivity ratio.

3) Tuner y is adjusted next to reduce the ac variation after the low reflection load is replaced by a short.

Figure 4 shows the reflectometer output vs. position of the sliding short circuit after the system has been tuned at 9.8 GHz using this technique. The slope is an indication of the attenuation of the standard waveguide.

The average time taken for the tuning process is about ten minutes compared to a typical time of thirty minutes or more using the manual technique. In addition, this technique does not demand as much skill and knowledge of the system from the operator as the manual technique does.

Acknowledgment

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A Note on Wave Propagation in Periodic Media

The solution of problems involving wave propagation in longitudinally stratified media leads to two different equations which must be solved in order to ascertain the longitudinal dependence of the field quantities [1]–[3]. These are, for the TE and TM modes respectively,

$$\frac{d^2U^{(h)}}{dz^2} + [\omega^2\mu_0\epsilon(z) - \gamma_h{}^2]U^{(h)} = 0 \quad (1)$$

$$\frac{d^2U^{(e)}}{dz^2} - \frac{1}{\epsilon}\frac{d\epsilon}{dz}\frac{dU^{(e)}}{dz} + [\omega^2\mu_0\epsilon(z) - \gamma_e{}^2]U^{(e)} = 0 \quad (2)$$

in which ω denotes the frequency, μ_0 the (constant) permeability and ϵ the permittivity of the medium, and γ^2 the sum of the squares of the transverse separation constants. U is a function describing the longitudinal dependence of a field component or of a scalar potential function from which the field quantities may be derived, and z denotes the longitudinal coordinate.

It will be shown that when ϵ is a continuous even-periodic function of z with period p the solutions to (1) and (2) may be expressed in terms of solutions to Hill's equation [4], the method of solution of which is tedious but straightforward.

One makes the substitutions

$$\zeta = \frac{\pi z}{p} \quad (3)$$

$$U^{(h)}(z) = f^{(h)}(\zeta) \quad (4)$$

in (1) yielding

$$\frac{d^2f^{(h)}}{d\zeta^2} + \left(\frac{p}{\pi}\right)^2(\omega^2\mu_0\epsilon - \gamma^2)f^{(h)} = 0. \quad (5)$$

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