TABLE I

	Intermodulation Rejection (dB)
Conventional Operation	50
New Technique Single Diode	80
New Technique Balanced Mixer	100

The measurements described were made in the 200 MHz region, but the technique is applicable at any frequency range where similar mixing devices are available.

ACKNOWLEDGMENT

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Accurate Phase-Length Measurements of Large Microwave Networks

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Abstract—The Stanford two-mile linear accelerator uses 240, single-input-port, four output-port, S-band, rectangular-waveguide networks to feed RF energy sixty to seventy feet from the klystrons above ground to ten-foot-long, disk-loaded, circular-waveguide, accelerator sections below ground. During installation it is necessary to permanently adjust the phase lengths of the four network branches to be within ±4.5 electrical degrees of the design lengths for RF wave and electron beam synchronization.

A modulated reflection phase-length comparison method is used, whereby a small signal is sent into each branch and reflected, in turn, by a diode switch, which is turned on and off at a 1 kHz rate. A null occurs in the amplitude modulation of the sum of a large reference signal and the small reflected signal, when the two signals are nearly in-phase quadrature. The reflectors are placed so that the network branches are properly adjusted when the nulls from all branches occur for the same setting of a variable phase shifter in the measurement line.

Small mismatches and multiple power divisions do not affect the accuracy of this method. Frequency, temperature, and air pressure are the main environmental conditions affecting the measurement and are discussed along with the design of the reflecting diode switch, which is mounted in a vacuum-sealed waveguide flange.

I. Introduction

OFTEN THE accurate measurement of electrical phase lengths is necessary in the design of microwave circuits. For numerous components and devices, a short length of transmission line is needed to act as a transformer to match two impedances. In other cases, where the load and the generator are matched already to the characteristic impedance of the transmission line, but the generator feeds several output ports, it may be important that the signals arrive at the respective ports with specific time delays or phase relationships. The system described below is designed to measure this latter type of network.

The method used to make an electrical phase-length measurement depends upon the type of network. Relative or comparison phase-length measurements generally are easier to make than absolute phase-length measurements, which may require detailed knowledge of the device's phase vs. frequency response. Thus, the phase length usually is determined by comparison with a similar, known or reference phase length in conjunction with an accurately calibrated, variable phase shifter.

The device to be measured may be inserted into the measurement circuit, so that the test signal either is transmitted through the device, or sent through the device and reflected back through it by a reflector on the output. A reflection method measures twice the length, and thus half-cycle phase length differences appear as full-cycle phase shifts. This phase ambiguity in large networks, with uncertain phase-frequency characteristics, is not resolvable simply by measuring the phase length at several frequencies. A transmission phase measurement method for a large network usually requires that a long phase-stable return path be provided from the output port of the network back to the phase measurement system. Often with both methods, effects due to small undesired CW reflections and unequal circuit attenuation can be minimized by modulating the signal. A more complete discussion of various phase-length measuring techniques is given by Lacy, and others [1]–[7].

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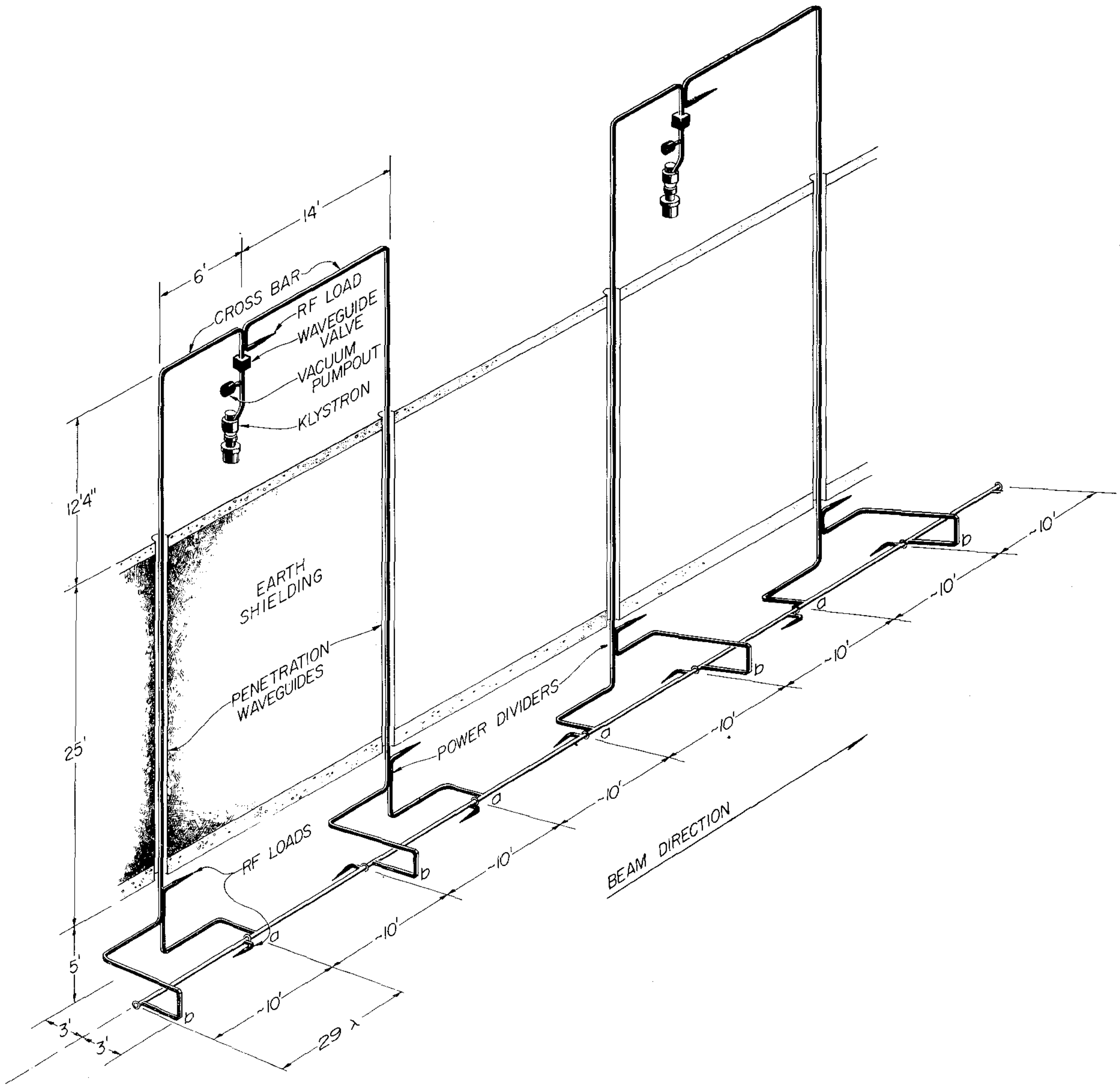


Fig. 1. Waveguide network.

The problem of adjusting the phase length of the 240 rectangular waveguide feed networks for Stanford's two-mile linear accelerator was solved by using a modulated reflection method [8], similar to one suggested by Schaeffer [2] and later further developed and used by Swarup and Yang [9] to adjust a radio astronomy antenna array. The accelerator utilizes a high power, S-band, rectangular-waveguide network to feed RF energy from each 25 megawatt klystron amplifier through 25 feet of earth shielding to four, independently fed and terminated, 10 foot, disk-loaded, circular-waveguide, accelerator sections (Fig. 1). In operation, an RF wave appears to move along the entire accelerator in a single coherent wave with a phase velocity equal to c. This condition requires that each accelerator section have an RF phase velocity equal to c at the operating frequency, 2856 MHz, and that the RF wave entering each section be phased correctly with respect to the bunched electron beam. Each klystron is individually phased by an automatic system [10] to obtain the correct phase relationship between the bunched beam and the wave in one particular accelerator section driven by that klystron. Thus, the high power waveguide feed network must be permanently adjusted so that when the wave is phased correctly with respect to the beam in that one accelerator section, it is phased correctly in the other three sections driven by that same klystron. Since the accelerator RF input ports are spaced by an integral number of wavelengths (29), the waveguide network branches must be adjusted to be equal in phase length, or to differ by only an integral number of wavelengths.

The modulated reflection method for phase adjustment uses an individually controlled reflector (modulator flange) at each of the four output ports. A 2856 MHz CW signal is fed to the input port of the waveguide network, then one reflector at a time is switched at a 1 kHz rate, and each return signal is compared in quadrature with a much larger unmodulated reference signal. The resultant signal will not exhibit a 1 kHz amplitude modulation when the reference signal and the carrier of the modulated reflected signal are out of phase by ninety degrees. Meanwhile, the diodes in the other three modulator flanges are not switched, but are dc biased to cause little reflection. This allows comparison of the phase lengths of the four branches, subject to the half-cycle ambiguity of the reflection method. The half-cycle ambiguity then is resolved by a transmission phase measurement, which uses the modulator flanges as coax-to-waveguide adapters and uses a long coaxial cable as the return path to the input port of the waveguide network.

The particular problems encountered with the networks (see Fig. 1) are: its large physical size, its 6 dB of power division per branch, and numerous, small mismatches of undetermined phase. The great length of the network branches requires that the phase-length dependent parameters (temperature, frequency, and dimensional tolerances) be stringently controlled. However, some of the temperature-dependent effects tend to cancel, as they are nearly the same for all the branches of a network. Unequal attenuation and random, small, network reflections produce only second-order phase measurement errors with the modulated reflection method.

II. Description of the Measurement System

The phase-length measurement system consists of four modulator flanges (Fig. 2), connecting cables, and the phasing machine console (Fig. 3). The circuit, including the network under measurement, is shown in Fig. 4. A vector diagram showing the relationship between the pertinent signals at the detector and the equations for those signals are given in Fig. 5. Equations (1), (2), and (3) give the expressions for the CW reference signal v_r, the square-wave modulated reflected signal v_m, and the resulting sum signal v_s, respectively. The microwave frequency is ω_c, and the reflector is switched at ω_m. The reference phase length from the signal generator to the detector is η, and the phase length from the generator through one branch of the network to the modulator flange and back to the detector is θ. The modulation index is m. Equation (4) gives the quadrature null conditions.

Two things should be pointed out about (4). First, the null conditions are independent of the type of amplitude detection (linear, square-law, etc.), and secondly, the phase length θ varies as twice the length of a branch. Thus, the phase shifter in Fig. 4 can be adjusted so that η is in quadrature with θ for any network branch, if its length is any multiple of quarter wavelengths long. The quarter-cycle ambiguity can be resolved by noting (see the vector diagram of Fig. 5) that when θ is increased by the phase shifter the resulting vector v_s increases in amplitude if $(\theta-\eta)\approx 270°\ \phi$ and decreases when $(\theta-\eta)\approx 90°\ \phi$. Whether v_s increases or decreases is easily determined by synchronous detection; that is, by triggering the oscilloscope (Fig. 4) with a 1 kHz signal from the modulator power supply and observing whether the 1 kHz amplitude modulation of v_s (at the crystal detector) is of the same phase as each of the four modulator flanges energized in turn.

The remaining half-cycle phase ambiguity is resolved by a transmission phase measurement, for which a 75-foot flexible coaxial cable serves as the return path from each modulator flange, one at a time, to the phasing machine console. The phase-length instability of the coaxial cable prevents the transmission measurement from being accurate to better than $10°\ \phi$. A reference signal and the transmitted signal are fed into opposite ends of a standing-wave detector. The variable attenuator equalizes the amplitudes of the reference and transmitted signals to produce a null pattern in the standing-wave detector, and the sliding probe indicates any shift in null position from branch to branch. The modulator power supply square-wave modulates the RF signal from the generator, reverse-biases one modulator flange (through the dc isolation T), and forward-biases the other modulator flanges.

In the fine-phase (modulated reflection) measurement, the modulator power supply switches one of the four modulator flanges at a 1 kHz rate, forward-biases the other three, and triggers the oscilloscope sweep. The calibrated dielectric-slab phase-shifter is adjusted to produce a null in the 1 kHz amplitude modulation of the sum signal v_s at the detector. The difference in phase shifter settings for the four branches of the network are direct measures of the phase length differences. The standing-wave indicator conveniently doubles as a 1 kHz tuned preamplifier for the detected amplitude modulation of the sum signal. The oscilloscope displays this amplitude modulation, and serves also as the synchronous detector which resolves the quarter-cycle ambiguity. The standing-wave detector is used also in a separate measurement to measure the input reflection of the network after tuning. Additional facilities permit calibration of the reflection phases of the modulator flanges and monitoring of drift in the phasing machine.

The modulator flange (Fig. 2) consists of a diode switch mounted in a special, stainless steel, vacuum sealing, S-band waveguide flange. The point-contact germanium diode is spring loaded on the end of a post across the waveguide. Two adjustable tuning screws in the plane of the diode are used for matching. The tuning screws and the diode post, which connects to the center conductor of a TNC fitting on the flange circumference, are vacuum sealed with teflon O-rings. Thin gold plating on top of copper plating improves the calibration accuracy and the shelf life of the stainless steel flanges. When the flange is properly tuned, reverse biasing the diode creates a large shunt admittance across the waveguide, which causes almost complete reflection. Forward biasing creates only a small shunt admittance, which causes

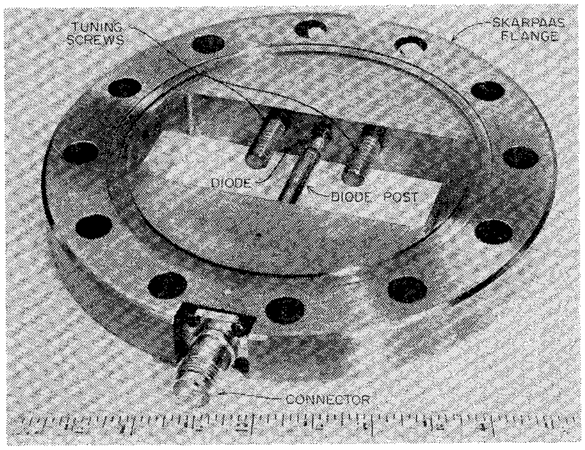


Fig. 2. Modulator flange.

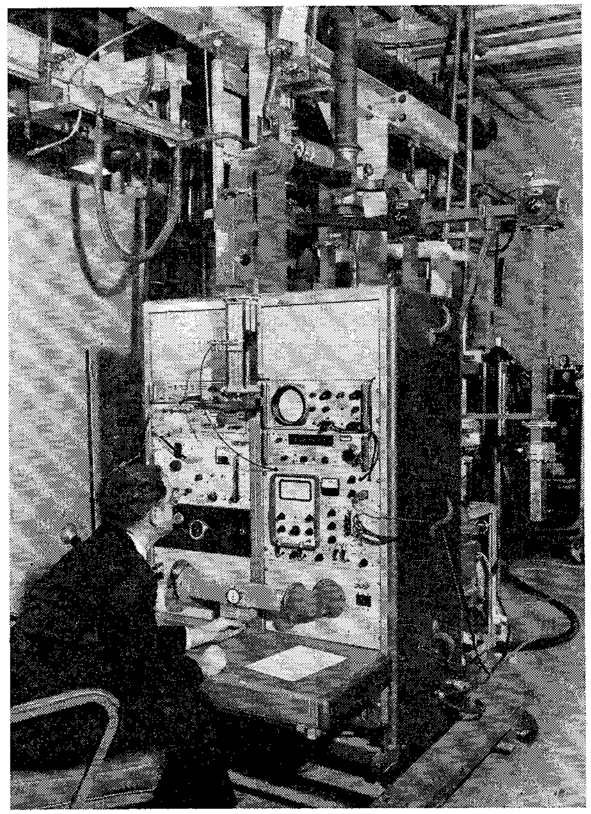


Fig. 3. Phasing machine console.

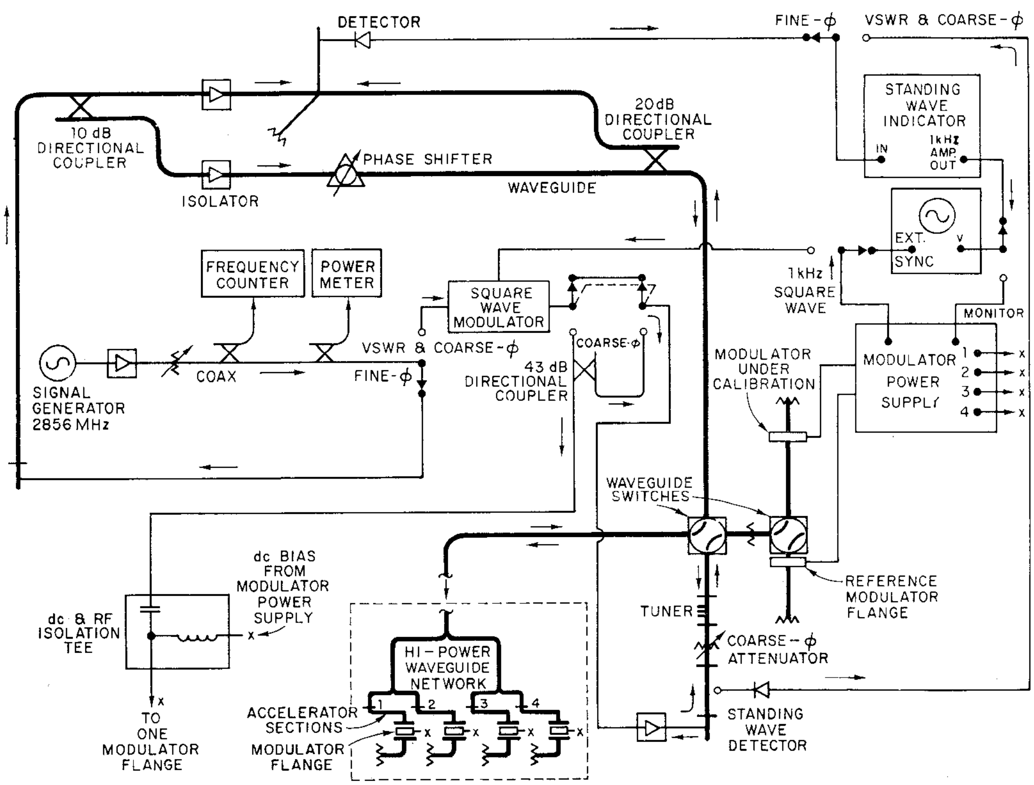
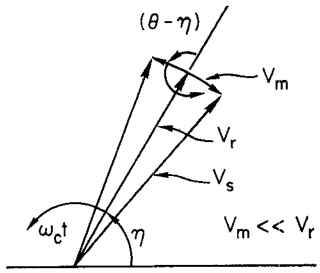


Fig. 4. Block diagram of phasing machine.



$$v_r = V_r \cos(\omega_c t + \eta) \tag{1}$$

$$v_m = V_m \left[\left(\frac{2-m}{2}\right) + \frac{2m}{\pi} \sum_{n=0}^{\infty} \frac{\cos(2n+1)\omega_m t}{(2n+1)}\right] \cos(\omega_c t + \theta) \tag{2}$$

$$v_s = \left\{\text{dc terms} + \frac{4mV_m}{\pi}\left[\left(\frac{2-m}{2}\right) V_m + V_r \cos(\theta-\eta)\right]\cos\omega_m t + \text{harmonic terms}\right\}^{1/2} \cos\left[\omega_c t + \alpha(\omega_m t, \eta, \theta)\right] \tag{3}$$

A null in the amplitude modulation of v_s at the frequency ω_m occurs for

$$\left(\frac{2-m}{2}\right) V_m + V_r \cos(\theta-\eta) = 0 \quad \text{or}$$

$$(\theta-\eta) = \cos^{-1}\left[\frac{-V_m(2-m)}{2V_r}\right] \approx \frac{\pi}{2} \quad \text{and} \quad \frac{3\pi}{2}, \tag{4}$$

where $0 < m < 1$ and $V_m \ll V_r$.

Fig. 5. Equations and vector diagram showing quadrature null conditions.

little reflection [11]. The tuning screws are adjusted so that the phase of the sum of the small and large reflections are the same in all four modulators. Since the diodes exhibit a nonlinear phase vs. voltage characteristic, square-wave modulation makes for easier understanding and analysis of the circuit. The modulator power supply switches the diode bias signal at a 1 kHz rate from −20 volts to +100 mA. When the diode is reverse biased, the modulator flange serves as a coax-to-waveguide adapter with stable phase characteristics and a transmission loss of about 20 dB. Between the diode support post and the flange body, there is a capacitive reactance of about 1.4 ohms at 2856 MHz, which provides some isolation between the bias circuitry and the microwave reflector circuitry. Since the modulator flange is primarily a precision reflector, and only incidently a coax-to-waveguide adapter, the 20 dB transmission loss is preferable to less isolation. Frequent recalibrations locate the position of the equivalent plane of reflection to better than $\pm 0.3° \phi$. The major source of calibration error is the small movement of the internal parts of the flange modulator during its installation. The modulator flange is definitely the most critical component of the system.

Before the waveguide components are assembled, they are matched to a VSWR of better than 1.05. Ground loops, frequency drift, temperature fluctuations, and modulator calibration accuracy are the major sources of phase adjustment error. If this type of system is to be used over a range of frequencies, and for absolute phase length measurements, a good deal of care would be required in its design and calibration. However, many of the mismatches and component phase characteristics are not detrimental when the system is operated at a single frequency and as a null comparison meter.

III. Results and Limitations of the Measurements

Phase-length measurements and adjustments were made on 240 networks. The branches of any single network are from 59 to 70 feet long (118 to 140 guide wavelengths), and there are seven, vacuum tight, copper gasket joints along each branch. The waveguide components were tuned for minimum VSWR before installation, but it was necessary to adjust the phase of the networks after final installation because of their great length, their numerous joints, and their sensitivity to temperature and vacuum conditions. The ten-foot accelerator sections were included between the modulator flanges and the output ports of the waveguide network (Fig. 4), so the network would not have to be moved or connected after its adjustment. Including the accelerator sections in the measurement was possible, since their phase lengths were adjusted to within $\pm 2.5° \phi$ before installation. The biggest problem was maintaining their temperature. Figure 6 shows the phase length dependency of a network and an accelerator section. The phase length of an accelerator section is seen to be eight to ten times more sensitive to temperature and frequency changes than a waveguide network branch. A water heating system that sets the accelerator sections and waveguide network temperatures nominally at 113° F is capable of maintaining a temperature difference between network branches of less than 0.75° F and between accelerator sections of less than 0.2° F for conditions of no RF power, such as during phase measurement. Phase-length stability under accelerator operating conditions is more complex, and is treated elsewhere [12].

The measurement frequency is monitored with a frequency counter and maintained within 500 Hz of 2856 MHz. The network and accelerator sections are evacuated to less than 25×10^{-3} (torr).

At this internal pressure, changes in internal pressure do not affect appreciably the phase length, either by change in dielectric constant or by elastic deformation of the waveguide walls. However, elastic deformation, due to changes in external atmospheric (barometric) pressure, can cause measurable changes in phase length. These changes will be less than $0.014(°\phi)\cdot(\text{ft})^{-1}$ for 5

Phase Coefficient	$\theta_F \frac{(^\circ\phi)}{(\text{kHz})}$	$\theta_T \frac{(^\circ\phi)}{(^\circ F)}$
70 feet Waveguide Network Branch	+0.037	+1.0
10 foot Accelerator Section	+0.30	+7.9

Fig. 6. Phase length coefficients.

percent changes in barometric pressure. It is interesting to note, too, that the total effect of evacuating the system from 760 torr of dry N_2 to less than 25×10^{-3} torr is -0.68 ($^\circ\phi$)$\cdot$(ft)$^{-1}$ for the rectangular waveguide and -280 ($^\circ\phi$) for the ten-foot accelerator section [12].

The total attenuation of the reflected signal v_m is about 52 dB greater than the reference signal v_r; therefore, the quadrature null condition of (4) results in a $(\theta-\eta)$ of not exactly 90 ($^\circ\phi$) or 270 ($^\circ\phi$), but 90.07 ($^\circ\phi$) or 269.93 ($^\circ\phi$). Thus, small variations in attenuation of less than a dB from branch to branch cause less than 0.01 ($^\circ\phi$) error. The waveguide network input VSWR after tuning is less than 1.2; this contributes a negligible error to this kind of modulated-reflection, null-comparison method. The actual phase adjustments are performed by permanently indenting the waveguide walls with special C-clamps with twelve-inch jaws. Smooth indentations over a length of several feet easily produce phase shifts up to the usually maximum variation between branches of 60 ($^\circ\phi$). A few networks were out of adjustment by almost 180$^\circ\phi$ and required clamping over longer lengths to prevent significant reflections. Two operators in telephone communication easily made the remote adjustments. Bowing in of the narrow wall of the waveguide decreases its phase length, and bowing in of the broad wall (with consequent bowing out effects on the narrow wall) increases its phase length.

The measurement system is capable of reproducing measurements within ±0.1 ($^\circ\phi$). The modulator flanges have a calibration accuracy of better than ±0.3 ($^\circ\phi$). The networks have a phase stability of better than ±0.5 ($^\circ\phi$). The accelerator sections are within ±2.5 ($^\circ\phi$) of their design lengths. Thus, the overall accuracy of phase adjustment is better than ±4.5 ($^\circ\phi$), allowing ±1.0 ($^\circ\phi$) for temperature instabilities for the accelerator sections.

Figure 7 shows the distribution of phase unbalance of the waveguide components before tuning. Measurements were made, at most klystron stations of the phase unbalance from the klystron port to the input port of the lower power divider, and at all stations of the S-assemblies (the quarter-power portions of the rectangular waveguide network, including the lower power divider). Separate data are shown for S-assemblies at odd- and even-numbered klystron stations because of the alternating feed arrangement shown in Fig. 1. The data show that the rectangular waveguide crossover branch

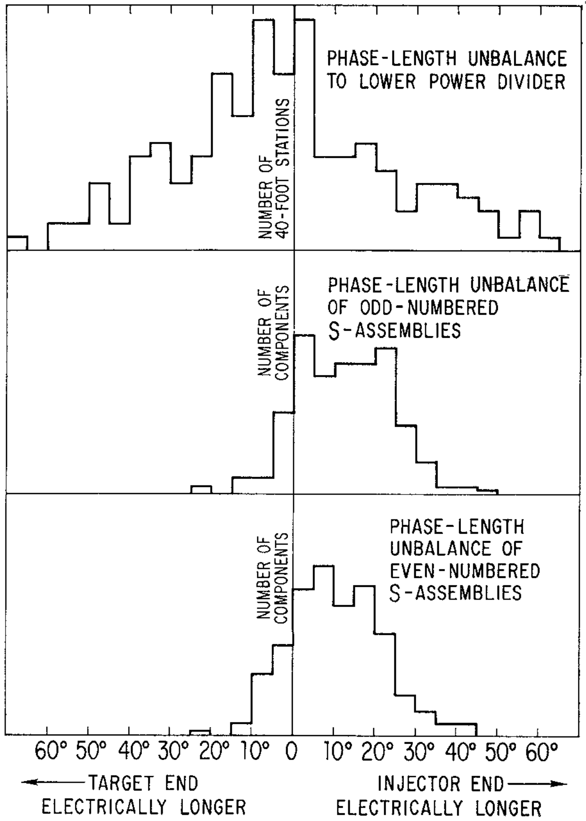


Fig. 7. Graphs of phase length errors due to manufacturing tolerances.

is correctly designed, as its location in the network does not affect the mean phase unbalance. The 15$^\circ\phi$ mean unbalance did not warrant adjustment of the design of the rectangular waveguide network during manufacture.

In conclusion, the modulated reflection scheme has been found to work very well on the large, single-input port, multi-output port, waveguide networks. The greatest errors come from limited control over the environmental parameters. The use of a diode for the modulated reflector is very satisfactory as long as the microwave power level at the diode was kept low (a few milliwatts). If this type of scheme is used at different frequencies, either to make absolute phase length measurements or to measure the phase vs. frequency characteristics of a network, the reflector and the microwave comparison circuit must be broadbanded or calibrated as a function of frequency.

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Sampling for Oscilloscopes and Other RF Systems: Dc Through X-Band

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***Abstract*—Sampling techniques as used in wideband oscilloscopes have, in the past, yielded bandwidths up to 4000 MHz. This approach has now been employed to achieve bandwidths in excess of 15 GHz. The design requirements necessary for this extended bandwidth are presented along with a detailed description of one solution to the design problem.**

The device is basically a two-diode sampler located at the center of a dielectric filled, biconical cavity containing the RF transmission line. The RF line is perpendicular to the axis of the biconical cavity. The sampling pulse is introduced into the cavity by applying it directly between the centers of the opposite faces of the cavity. This establishes a potential difference between two points on the ground conductor of the RF transmission line being sampled. This technique is basic to the operation of the device and plays a key role in the reduction of sampling loop inductance, which would limit the bandwidth. The equivalent circuits are presented along with the appropriate defining equations. The relationship between bandwidth, input VSWR, and step response overshoot, are presented, along with the typical measured results.

Introduction

SAMPLING TECHNIQUES have long been used on periodic waveforms to achieve wide bandwidths in oscilloscopes [1]–[12]. The same technique has been used in phase-locked loops [13], in random sampling voltage detectors, and other RF systems. Although the applications of sampling are quite different, the basic requirements for a broadband sampling device are nearly the same regardless of the application. This technique has now been extended to X-band frequencies and above by the development of a sampling device with a bandwidth in excess of 12.4 GHz.

There are other performance characteristics to be considered when evaluating a sampling device. Several of these are input signal dynamic range, sensitivity, input voltage standing-wave ratio, phase response, and mechanical configuration. The bandwidth of such a device, however, is the most important single performance characteristic.

There are many electronic systems presently using sampling devices and these devices vary widely in performance and general configuration. The state-of-the-art in wideband sampling devices up to now has been the 4 GHz bandwidth presently available in sampling oscilloscopes [14].

The bandwidth of a sampling device employing semiconductor diodes is determined entirely by the diodes, by the sampling pulse and by the method used to connect them to the RF transmission line being sampled. This paper will discuss these basic elements and the design requirements for each. A unique sampling circuit will be presented along with its mechanical realization. Typical measured performance data will then be presented, completing the design cycle.

A. Basic Sampling Requirements

Figure 1 shows an idealized sampling circuit. Switch S is closed for a short period of time, allowing the sampling capacitor C to charge to some fraction of the voltage

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