III-3. ACCURATE PHASE LENGTH MEASUREMENTS OF LARGE MICROWAVE NETWORKS

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Electrical phase length measurements and adjustments on long waveguide or coaxial feed lines are necessary sometimes in antenna arrays and particle accelerators. Most of the commonly used phase length measurements schemes are of a transmission or reflection type and may use a modulated signal. Transmission methods usually require a precision reference length to reach both ends of the network, while reflection methods require only a stable reflector, but inherently cannot resolve $180^{\circ}\ \phi$ differences.

The Stanford two-mile linear 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, disc-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 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,[1] to obtain the correct phase relation between the bunched beam and the wave in one particular accelerator section driven by that klystron. The high-power waveguide feed network must be permanently adjusted so that when that one accelerator section is phased correctly, the other three sections driven by that same klystron are also phased correctly. This paper describes the measurement technique used in the permanent phase-adjustment of the rectangular waveguide.

Since the accelerator rf inputs are spaced by an integral number of wavelengths (29), the waveguide network branches only need to be equalized within an integral number of wavelengths. A modulated reflection method, similar to the one used by Swarup and Yang[2] to adjust an antenna array for radio astronomy, was used at SLAC with the single-input-port, four-output-port, waveguide network. A modulated reflector at each output port is switched on, one at a time, and the small return signal is compared in quadrature with a large reference signal at the phasing machine console. Figure 2 shows the vector relationships and Eqs. (1), (2), and (3) give the expressions for the reference signal, v_r , the square wave modulated reflected signal, v_m , and the resulting sum signal v_s . 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 reflection modulator and back to the detector is θ .

Two things should be pointed out about Eq. (3) and the resulting conditions for a null at ω_m. 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. 3 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 $90^{\circ}\ \phi$ ambiguity can be resolved by noting (see the vector diagram of Fig. 2) that when θ is increased by the phase shifter the resulting vector, v_s , increases in amplitude if $(\theta - \eta) \approx 270^{\circ}\phi$ and decreases when $(\theta - \eta) \approx 90^{\circ}\ \phi$. Whether v_s increases or decreases is easily determined by synchronous detection; that is, by triggering the oscilloscope with a 1 kHz signal from the modulator supply and observing whether the 1 kHz amplitude modulation of v_s from the crystal detector is consistently positive or negative (in phase or out of phase) for all the branches.

The 180° ϕ ambiguity can be resolved by using a coarse transmission method, as shown in Fig. 3. A microwave signal is fed through 75 feet of coaxial cable to a modulator flange (which acts like a very loosely coupled coaxial to waveguide adapter when its diode is reverse biased) and on through one network branch at a time to one end of the slotted line. Also, a sample of the generator signal is fed to the other end of the slotted line. An attenuator adjusts the amplitudes of the two waves to achieve a null, whose position is located by the slotted line probe. This is done for each branch, and the null positions are compared to resolve the 180° phase ambiguities. The slotted line is used to measure the reflection from the waveguide network, also. Figure 4 shows the physical arrangement of the circuit in Fig. 3.

The modulator flange (Fig. 5) consists of a diode switch mounted in a special flange. The diode is spring loaded on a post which, through a TNC fitting on the flange, connects to a 1kHz bias power supply. The post and the two vacuum tight, adjustable, matching screws are sealed with teflon O-rings. Forward biasing the diode causes the modulator flange to be a small shunt admittance across the waveguide, which results in little reflection. Reverse biasing creates a large shunt admittance, which results in a large reflection. During phase measurements each flange, one at a time, is switched at 1kHz, while the other three are forward biased for minimum reflection.

It is necessary to phase adjust the network after its final installation because of its great length, its numerous joints, and its phase length sensitivity to temperature and vacuum conditions. The S-band waveguide branches are from 59 to 70 feet long (118 to 140 guide wavelengths), and there are seven, vacuum-tight, copper-gasket joints along a branch. The changes in phase length due to temperature, frequency, and vacuum changes are $+ 0.0144\ (^{\circ}\phi) \cdot (^{\circ}F \cdot ft)^{-1}$, $+ 0.532\ (^{\circ}\phi) \cdot (MHz \cdot ft)^{-1}$, and $- 0.8\ (^{\circ}\phi) \cdot (ft)^{-1}$, for a change from atmospheric pressure to less than 25×10^{-3} torr. The modulator flanges are installed between the accelerator sections and their rf loads rather than at the input to the sections, and thus after phase-adjustment only the load need be moved in order to remove the modulator flanges. Including the accelerator sections in the measurement is possible since their phase lengths are adjusted very accurately before installation. The total attenuation down and back through a network branch is 22 dB due to the two power dividers and the accelerator section. An additional 30 dB of attenuation in the phase measurement circuitry results in a $(\theta - \eta)$ of 90.07° ϕ or 269.93° ϕ for the null conditions given in Fig. 2. The input VSWR due to tuning and the various components is less than 1.2:1. The average temperature of the branches can be maintained within 0.75°F of each other by a water heating system which keeps the temperature of the accelerator and the network close to 113.0°F. The actual adjustment of the phase length is performed by squeezing the waveguide walls with a modified C-clamp. The modulator flanges are calibrated and maintain an accuracy better than $\pm$ 0.5° ϕ. Of the 240 networks measured and adjusted, two had phase errors of approximately 180° ϕ, while most of them had maximum differences from 30° ϕ to 60° ϕ. If all the errors due to temperature differences, modulator calibration, and built in measurement circuitry are considered the design phase error of $\pm$ 2° ϕ for the network was readily achieved on a production basis. The system, although used at a fixed frequency, is not limited to that and can readily be used to measure phase versus frequency characteristics or the absolute phase length of a network, whose phase versus frequency characteristics are known.

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References

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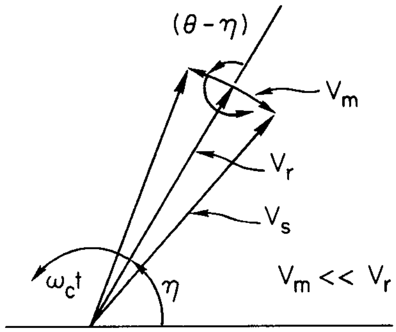
2. G. Swarup and K. S. Yang, "Phase Adjustment of Large Antennas," IRE Trans. on Antennas and Propagation, V.AP-9, p. 75-81, January 1961.



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

$$v_m = \frac{V_m}{2}\left[1 + \frac{4}{\pi}\sum_{n=o}^{\infty}\frac{\cos(2n+1)\,\omega_m t}{(2n+1)}\right]\cos(\omega_c t + \theta) \tag{2}$$

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

For $\frac{1}{2}V_m + V_r\cos(\theta - \eta) = 0$, the ω_m amplitude modulation is zero.

$\therefore\ (\theta-\eta) = \cos^{-1}(-V_m/2V_r) \approx \pi/2,\ 3\pi/2$ for a null in ω_m.

FIG. 2 - EQUATIONS AND VECTOR DIAGRAM SHOWING NULL CONDITIONS

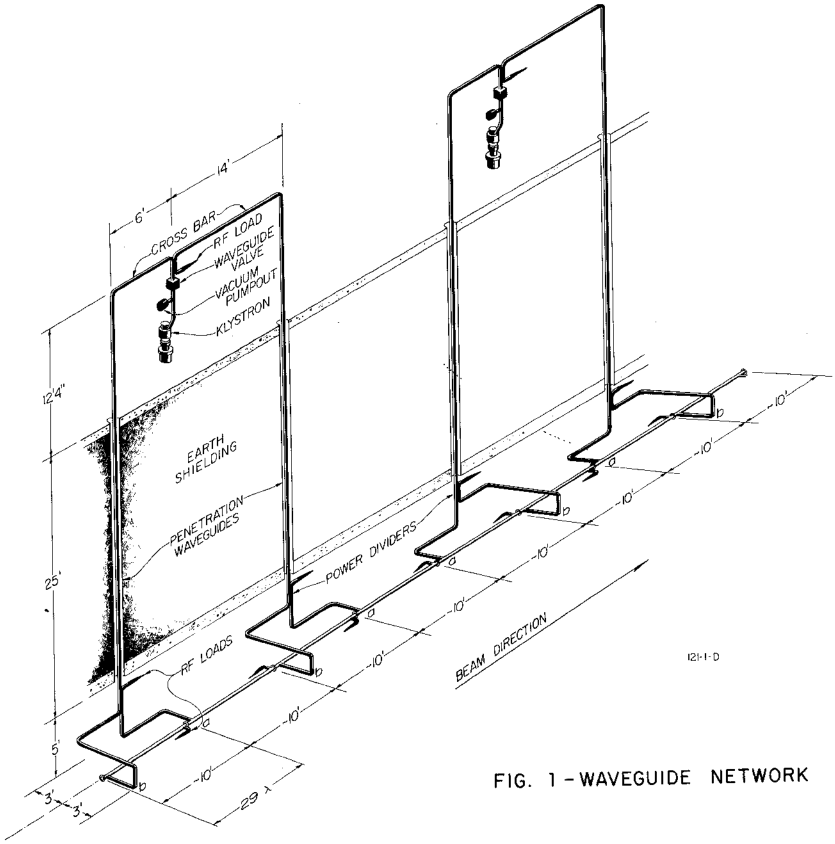


FIG. 1 - WAVEGUIDE NETWORK

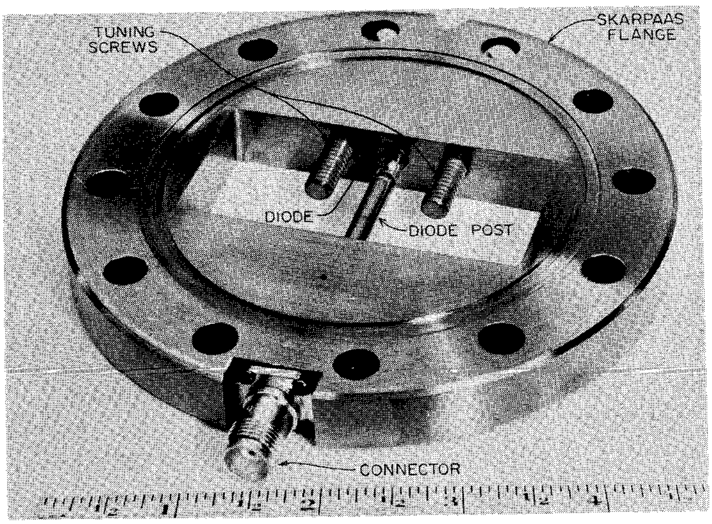


FIG. 5 - MODULATOR FLANGE

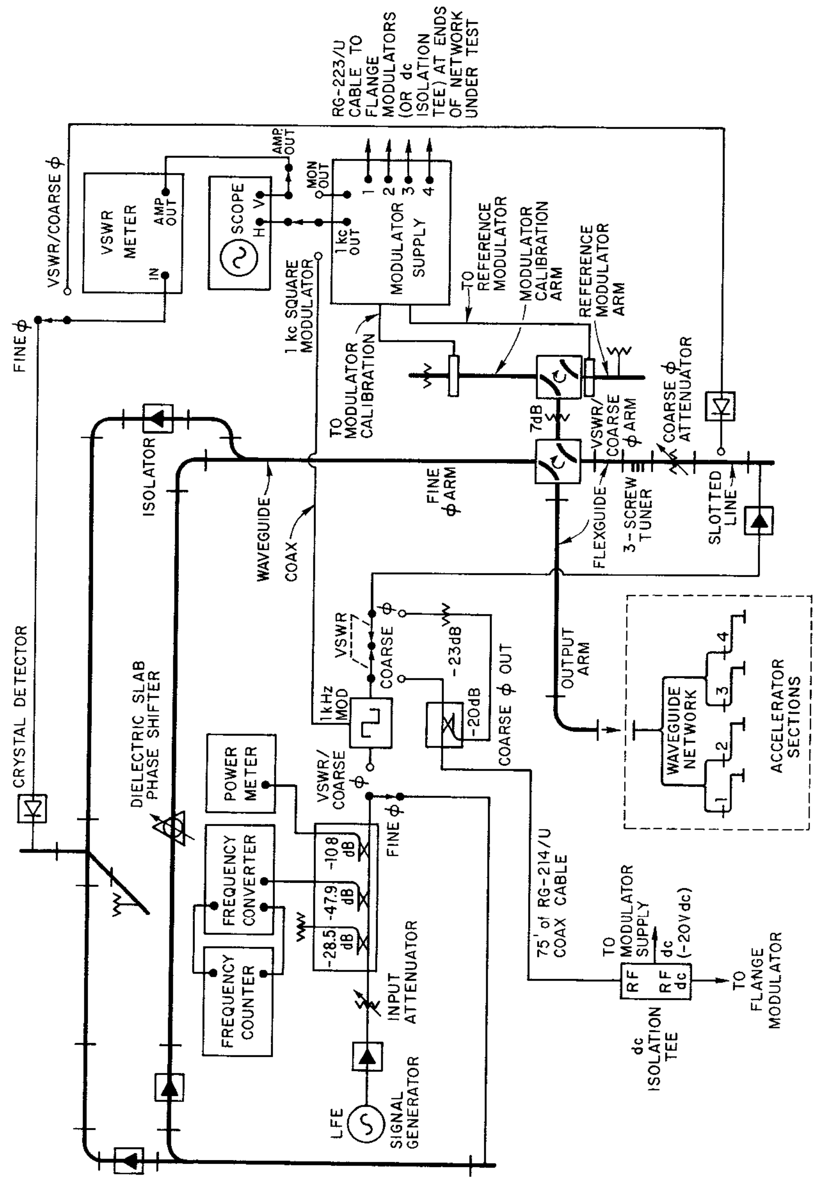


FIG 3 - BLOCK DIAGRAM OF PHASING MACHINE

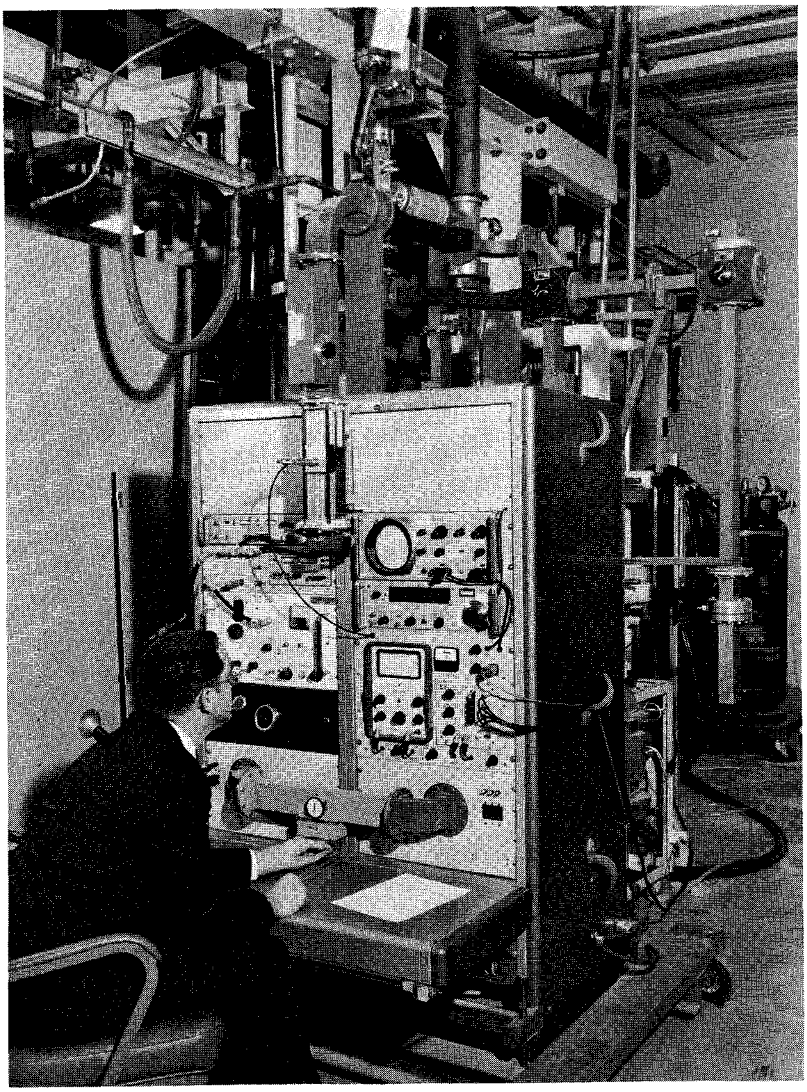


FIG. 4 - PHASING MACHINE CONSOLE