



HOMODYNE GENERATOR AND DETECTION SYSTEM

I. INTRODUCTION

A homodyne generator and detection system for measurements such as high values of attenuation, antenna patterns and rf signal leakage is shown in figure 1.

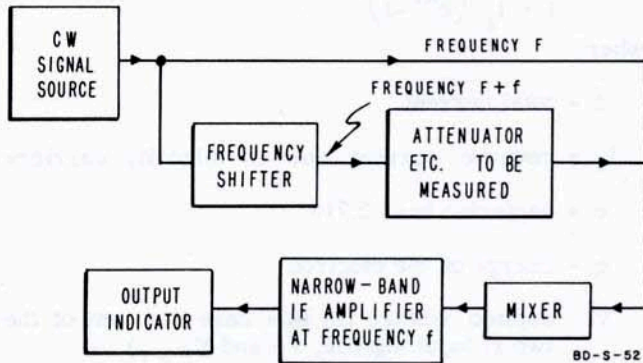


Figure 1. Homodyne Generator and Detection System

Briefly, the operation of the equipment is as follows:

The rf signal at frequency F , from the cw signal source, is split into two parts, one of which goes directly to the mixer and the other to an rf frequency shifter. The output from the latter is at a frequency $F + f$ (or $F - f$) where f is in the audio or low rf frequency range (1000 cps to 200,000 cps). This signal is fed to the attenuator which is to be calibrated and from there to the mixer. The mixer input therefore, consists of two signals - one at a frequency F from the signal source, and one at a frequency $F + f$ from the frequency shifter via the apparatus under test. The beat frequency output of the mixer at frequency f is amplified by a narrow band IF amplifier and the resulting signal amplitude displayed on an indicating meter. This output is linearly proportional to the amplitude of the mixer input signal at frequency $F + f$.

In the microwave frequency range, where well-isolated frequency shifters are available, the homodyne generator and detection system has several important advantages over other available systems:

A. In comparison with square-law detection, the most widely used method of microwave measurement, the homodyne system gives an improvement in sensitivity and dynamic range of about 40 to 50 db. One of the major advantages of the homodyne system is that a milliwatt of rf power can be used to measure

100 db of attenuation in comparison to the 100 watts of rf power which would be required to measure the attenuation in a square-law detection system. This makes the homodyne method ideally suited for use with low-level traveling-wave tube amplifiers as frequency shifters.

B. In comparison with the superheterodyne method of detection, the homodyne system avoids the necessity for accurate control of the frequency difference between the two microwave signals (from the external signal source and from the local oscillator in the receiver). The homodyne system shown in figure 2 (a 1000 cps intermediate frequency, 30 cps bandwidth and a crystal mixer) is comparable in sensitivity (-100 dbm) and dynamic range (80 db) to common pulse radar superheterodyne systems. If a barretter were used as the microwave mixer instead of a crystal, a 15 db improvement in dynamic range could be obtained by increasing the upper limit of linear detection to 0 dbm. System sensitivity would be about -95 dbm.

By working at a higher IF frequency, say 100 kc, and using the same 20 to 30 cps bandwidth, 20 db improvement in sensitivity over the above quoted figure can be obtained. This sensitivity is comparable to that of a cw radar superheterodyne system. The higher intermediate frequency and narrow bandwidth presents no particular problem because the frequency difference can be very easily controlled.

C. The homodyne system is a phase coherent system and thus relative rf phase measurements can be made in the offset signal channel by converting the rf phase changes to relative phase changes at the IF frequency.

II. SYSTEM COMPONENTS

The homodyne system diagrammed in figure 1 can be very easily assembled using standard Hewlett-Packard components as shown in figure 2. The use of a traveling-wave tube amplifier as a frequency shifter has a number of advantages over other types of shifters such as crystals, ferrites, rotating dielectrics, gas tubes and klystrons. For example, a wide 2:1 bandwidth can be easily covered by the signal shifter with a minimum number of adjustments since the offset frequency and the IF frequency track automatically whenever the frequency of the signal source is changed. Another important advantage is the degree of isolation that can be provided between the input signal to the

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signal shifter and the output signal at the offset frequency. If a net gain is not required from the traveling-wave tube frequency shifter, an attenuator whose value is equal to the gain of the traveling-wave tube amplifier can be placed ahead of the TWT so that the net gain of the attenuator-traveling wave tube combination equals unity. This arrangement provides approximately 90 db isolation when \oplus traveling-wave tube amplifiers are used as frequency shifters (60 db backward loss of the traveling wave tube plus approximately 30 db attenuation). The isolation provided under these circumstances is comparable to that provided by a klystron.

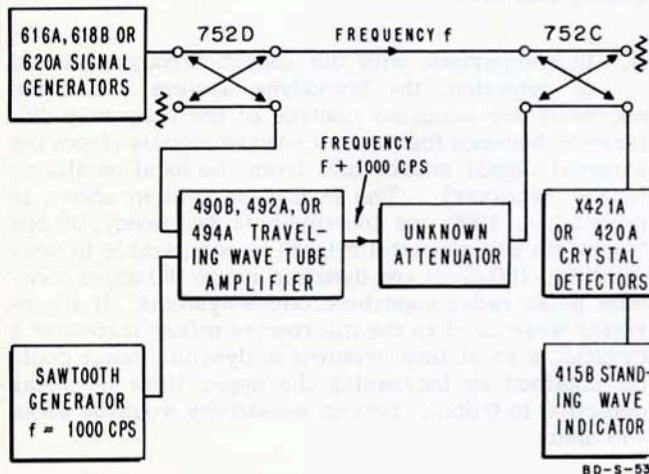


Figure 2. \oplus Homodyne System

The operation of most of the components shown in figure 2 is self-explanatory and the action of the sawtooth generator-traveling wave tube amplifier frequency shifter has been described in another application note.¹

III. SENSITIVITY & ACCURACY CONSIDERATIONS

With the general background, it is now possible to discuss the performance of the crystal mixer in limiting the sensitivity and dynamic range of the system. We will also discuss the approximate magnitude of the possible errors due to harmonics and other imperfections in the outputs of the sawtooth generator and the traveling-wave tube frequency shifter.

A. Crystal mixer limitations on the maximum and minimum signal levels.

1) Sensitivity. The sensitivity or minimum detectable signal is limited primarily by the signal-to-noise ratio of the crystal mixer. For a 1 kc IF frequency system similar to that shown in figure 2, the crystal has a high excess noise (proportional to $1/f$) which results in a noise figure of over 50 db. However, the

very narrow IF bandwidth of about 30 cps gives a sensitivity of about -100 dbm which is as good or better than that of a pulse radar receiver.

2) Dynamic Range. Since the minimum detectable signal is limited by the noise power generated in the crystal detector, the next question to consider is the maximum usable offset frequency level into the mixer.

The input-output characteristic of a microwave crystal detector can be approximated by the p-n semiconductor junction relationship:²

$$I = I_s \left(e^{\frac{qV}{kT}} - 1 \right)$$

where:

I = total current

I_s = reverse current due to minority carriers

e = napierian base 2.718

q = charge on the electron

V = applied voltage (in this case the sum of the two rf input signals, V_F and V_{F+f})

k = Boltzman's constant

T = absolute temperature.

It can be derived that the amplitude of the difference frequency output from a crystal is proportional to:

$$I_1 \left(\frac{qV_{F+f}}{k} \right)$$

where I_1 is a modified Bessel function of the first order. This is a linear function when $\frac{qV_{F+f}}{k}$ is small compared to kT/q but departs from linearity as V_{F+f} becomes large.

For the system shown in figure 2 and an assumed 50 ohm rf impedance at the crystal, this theoretical approximation and our experimental results show the mixer departs from linearity when the offset frequency signal is of the order of -20 dbm. Thus the maximum signal is limited to about -20 dbm and the dynamic range of the homodyne system is therefore 80 db (from -20 dbm to -100 dbm).

B. Errors introduced by the TWT and sawtooth generator characteristics.

1) Source of Errors. Besides the limits on the sensitivity and dynamic range determined by noise in the crystal and by the crystal input-output characteristics at high levels, there are other effects that produce errors at certain offset frequency signal levels. These errors are mainly due to harmonics and other imperfections in the outputs of the sawtooth generator

¹ \oplus Application Note 14: Traveling Wave Tube Amplifiers.

² \oplus Application Note 18: "Introduction to Solid State Devices" Bernard M. Oliver

and TWT signal shifter. As far as the traveling-wave tube output is concerned, these errors may be classified as:

- a. Errors at all levels due to image frequency.
- b. High signal level errors due to signals at F and $F + 2f$.
- c. Low signal level errors due to the leakage of the $F + f$ signal into F channel.

Before we consider these errors in detail, we will spend a little time reviewing the various harmonics and imperfections that are present in the traveling-wave tube output.

2) Harmonic Considerations. The output signal from the sawtooth modulated TWT is not a single frequency signal but contains a band of frequencies offset from the input signal frequency by harmonics of the sawtooth generator frequency (see figure 3).

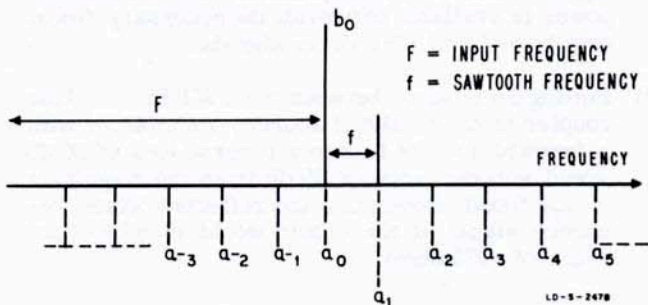


Figure 3. TWT Frequency Spectrum

In figure 3, the input signal is a frequency F , and of amplitude b_0 , and the main output signal is a frequency $F + f$, of amplitude a_1 . The doppler harmonics of the output signal are shown at frequencies $---F-3f, F-2f, F-f, F, F + 2f, F + 3f, F + 4f ---$ etc. and of amplitudes $---a_{-3}, a_{-2}, a_{-1}, a_0, a_2, a_3, a_4, ---$ etc. The amplitude of these doppler harmonics is normally 30-40 db below the main output signal (a_1).

These harmonics are caused by:

- a. Amplitude modulation of the rf signal as the helix voltage is sawtooth modulated to produce the 2π phase shift in the TWT.
- b. Non-linearity in the phase shift - helix voltage characteristic of the TWT, producing a slightly non-linear phase shift vs time for the rf signal.
- c. Sawtooth deficiencies such as a non-linear slope, or a relatively long flyback time.

When this spectrum of signals is mixed with the "local oscillator" signal in the crystal, a series of IF signals at frequencies $f, 2f, 3f ---$ etc. will be produced. The narrow-band amplifier eliminates from necessary consideration all but those IF signals at frequency f . It is easily shown that the first and second terms of the output IF signal at frequency f are of the form:

$$V_{out} \text{ (at } f) = b_0 a_1 \left[\cos (2\pi ft + \phi_1 - \theta_0) + \frac{a_{-1}}{a_1} \cos (2\pi ft - \phi_{-1} + \theta_0) + \frac{a_0 \cdot a_1}{a_1 b_0} \cos (2\pi ft + \phi_1 - \phi_0) + \frac{a_2 a_1}{a_1 b_0} \cos (2\pi ft + \phi_2 - \phi_1) \right]$$

where $\phi_{-1}, \phi_0, \phi_1, \phi_2$ are relative rf phases of the respective signals, and θ_0 is the rf phase of the "local oscillator" signal.

3) Error Calculations.

- a. Error due to the signal at the image frequency (a_{-1}).

Now, in general, the terms a_0, a_{-1}, a_2 are $\ll a_1$, hence if a_1 is much less than b_0 , we can neglect the

terms $\frac{a_0 a_1}{a_1 b_0}$ and $\frac{a_2 a_1}{a_1 b_0}$ for medium to small values

of $\frac{a_1}{b_0}$ (i.e. $20 \log \frac{b_0}{a_1} > 20$ db). Practical power

levels would be in the order of -10dbm for the "local oscillator" and -30 dbm or less for the offset power.

The significant terms of the crystal mixer output at frequency f will then be equal to:

$$V_{out} = b_0 a_1 \left[\cos (2\pi ft + \phi_1 - \theta_0) + \frac{a_{-1}}{a_1} \cos (2\pi ft - \phi_{-1} + \theta_0) \right].$$

Depending on the relative rf phase of the two mixer input signals, these two terms in the output signal can add or subtract, thus giving:

$$20 \log V_{out} = 20 \log b_0 a_1 + 20 \log \left(1 \pm \frac{a_{-1}}{a_1} \right)$$

Therefore, the maximum error introduced by the image frequency signal is equal to:

$$e_1 = 20 \log \left(1 \pm \frac{a_{-1}}{a_1} \right)$$

For example, if $a_{-1} = 0.032 a_1$ (doppler harmonics 30 db down), this will amount to a maximum possible error of:

$$20 \log (1 \pm 0.032) = \pm .28 \text{ db in the indicated output}$$

independent of the relative level of the two rf signals at the mixer input.

- b. Error due to the signal at the original frequency (a_0) and the frequency offset by the second harmonic of the sawtooth generator frequency (a_2).

These two spectral components produce an error that is most serious when the offset frequency level and local oscillator level at the crystal mixer are nearly equal ($a_1 \approx b_0$). Assuming that the image frequency error is small compared with this error at these signal levels, the expression for the significant terms in the mixer output signal becomes:

$$V_{\text{out}} = b_0 a_1 \left[\cos(2\pi ft + \phi_1 - \theta_0) + \frac{a_0 a_1}{a_1 b_0} \cos(2\pi ft + \phi_1 - \phi_0) + \frac{a_2 a_1}{a_1 b_0} \cos(2\pi ft + \phi_2 - \phi_1) \right]$$

Under the worst possible phase conditions the maximum error is:

$$e_2 = 20 \log \left| 1 \pm \frac{a_0 a_1}{a_1 b_0} \pm \frac{a_2 a_1}{a_1 b_0} \right|$$

For example, if:

$$\frac{a_1}{b_0} = 1 \text{ and } \frac{a_0}{a_1} = \frac{a_2}{a_1} = 0.032$$

this error is equal to: $20 \log \left| 1 \pm 0.064 \right| = \begin{matrix} +.54 \text{ db} \\ -.60 \text{ db} \end{matrix}$

This is, therefore, a fairly sizeable error when the amplitude of the two mixer input signals are nearly equal.

- c. Error due to reflected signal from output of TWT.

There is still one other source of error that is significant at very low offset frequency levels into the crystal mixer.

Due to internal reflections in the TWT and external reflections from the output coupler and load, coupled with the finite backward propagation loss of the TWT, some of the offset frequency signal appears at the TWT input terminals. (This has been measured in all three low level TWT amplifiers and found to be of the order of 60 db below the output signal.) Referring to figure 2, this signal will be transmitted through the coupler between the signal source and the TWT, and, depending on the directivity of the coupler and the impedance match looking back into the signal source, will arrive at the mixer via the same transmission path as the local oscillator signal. E.g., if the coupler directivity is infinite and the swr looking into the signal source is 1.5 to 1, then the magnitude of the reflected offset frequency signal finding its way into the mixer will be:

TWT output -60 db -20 db (coupler loss) -14 db (reflection loss) -10 db (2nd coupler loss) =
TWT output -104 db.

If the TWT output is +10 dbm, this term will be of the order of -94 dbm which is significant when the offset frequency signal approaches the minimum detectable level.

This reflected offset frequency signal (a_{r1}) can add either in phase or out of phase with the desired offset frequency signal (a_1) and produce a maximum possible error of:

$$e_r = 20 \log \left| 1 \pm \frac{a_{r1}}{a_1} \right| \text{ db}$$

which would be serious if the ratio $\frac{a_{r1}}{a_1} \rightarrow 1$

The situation can be improved somewhat by:

- (1) Using a nearly reflectionless pad between the signal source and the first coupler, thus improving the impedance match looking back into the signal source. This is possible only if sufficient power is available to provide the necessary "local oscillator" and TWT drive signals.
- (2) Putting an isolator between the TWT input and the coupler from the signal source. An isolator with a forward loss of 1 db to a reverse loss of 30 db would subtract another 30 db from the magnitude of a_{r1} found above (i.e., the reflected offset frequency signal at the mixer would then be of the order of -125 dbm).

IV. TYPICAL EQUIPMENT SETUP AND ALIGNMENT PROCEDURE

A typical equipment setup for attenuation measurements in X-band is shown in figure 4. The following alignment procedure will aid you in adjusting the equipment to obtain optimum operation.

- 1) Connect the equipment as shown in figure 4. Do not connect the sawtooth generator yet. Arrange the hp 620A for cw operation at desired frequency.
- 2) Adjust the rf output from the hp 620A to -5 dbm, thus supplying -15 dbm local oscillator power into the X485B Detector Mount.
- 3) Simultaneously adjust the hp X375A Flap Attenuator and optimize the hp 494A Traveling Wave Tube Amplifier helix voltage to give a TWT output of 0 dbm. The hp 430C Power Meter will indicate -10 dbm.
- 4) Connect the sawtooth generator to the hp 494A and adjust the hp X382A Attenuator to introduce a 40 db loss into the circuit. Vary the frequency and amplitude of the sawtooth generator output until a maximum reading is obtained on the hp 415B Standing Wave Indicator. Observe the resulting sine-wave output from the hp X485B on the oscilloscope and adjust the amplitude of the sawtooth generator output and TWT helix voltage until minimum distortion of the beat

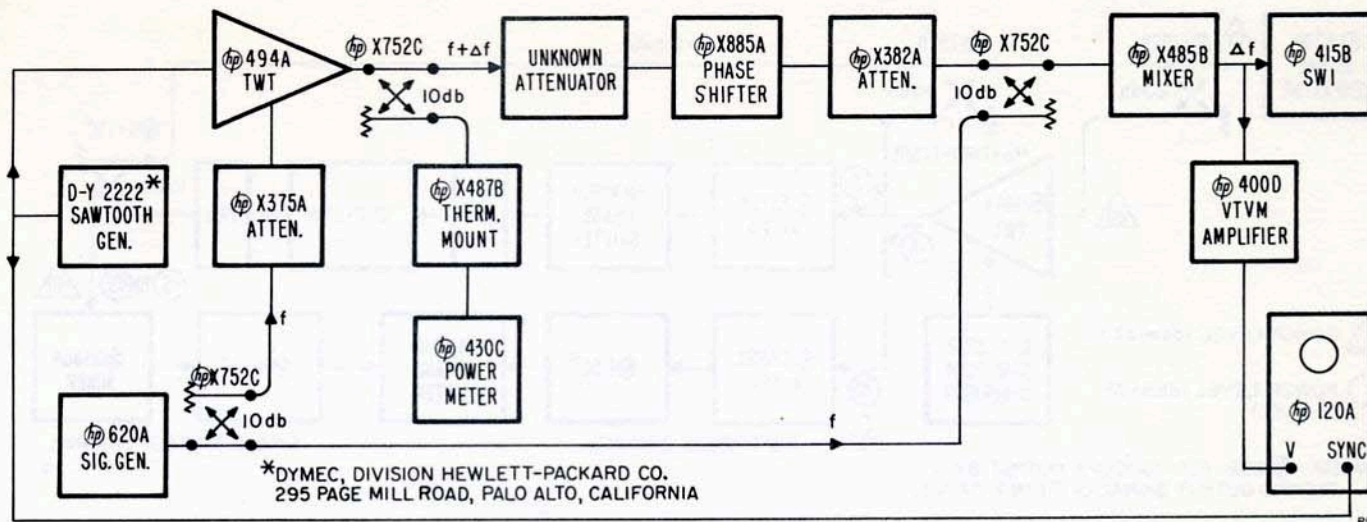


Figure 4. X-Band Homodyne System

sine wave is obtained for all rf phases. To obtain minimum distortion:

- a. Vary the hp X885A Phase shifter until the discontinuity (flyback) in the resulting 1000 cps sine wave appears at the peak of the sine wave.
- b. At that point, adjust the TWT helix voltage control until no discontinuity occurs at the peak of the sine wave. As this adjustment is made, the phase of the sine wave may change so that the phase shifter will have to be readjusted to keep the discontinuity at the peak of the sine wave.
- c. Phase the sine wave until the discontinuity appears at the axis crossing, using the X885A.
- d. Adjust the sawtooth generator output voltage for a minimum amount of discontinuity of the resulting sine wave at the axis crossing.
- e. Refine adjustments 1 through 4, to allow for interaction.
- f. Readjust the hp X375A Flap Attenuator to give 0 dbm output from the traveling wave tube amplifier as indicated on the hp 430C.

Functionally, when the discontinuity is set at the peak of the beat frequency sine wave both sides of the discontinuity represent two power outputs as determined by two different helix voltages. The idea there, of course, is to adjust the average helix voltage with the helix control until the power at each point on the two end points of the helix sawtooth swing is equal. At the zero axis crossing any discontinuity of the resulting sine wave is caused by not having an integral 2π radians of phase shift per sawtooth cycle. This, of course, is adjusted by applying a larger or smaller sawtooth voltage to the helix input.

- 5) Insert unknown attenuations and proceed with measurement.

CAUTION

Remember that the detector is now operating as a linear detector and 415B readings must be multiplied by 2 for attenuation.

During the above measurements, the output from the traveling-wave tube and the output from the hp 620A should remain constant. Therefore, special care should be taken to see that the hp 430C reading and the output of the hp 620A as shown on its internal power meter do not vary.

V. HIGHER SENSITIVITY DETECTION

The hp 302A Wave Analyzer offers increased sensitivity in the detection section of the homodyne for two reasons. First, its bandwidth of 6 cps is appreciably narrower than the 415B bandwidth of 40 cps. Secondly, its bfo output capability permits it to operate as a tuned voltmeter-oscillator combination up to 50 kc. Thus the Dymec 2222 Sawtooth Generator is driven at 50 kc by the hp 302A which is a 6 cps bandwidth tuned to the same offset frequency. This procedure improves considerably on $1/f$ crystal noise. Since the hp 302A input sensitivity is only $30\ \mu\text{v}$, additional sensitivity may be obtained by placing an hp 466A Broadband Amplifier between the detector and 302A, for an additional 40 db of gain.

VI. MICROWAVE PHASE SHIFT MEASUREMENTS

A homodyne system, since it is phase coherent, provides an excellent basis for measuring phase shift at microwave frequencies. It is necessary only to add another offset frequency channel to provide a phase reference. A typical setup for X-band is shown in figure 5, where phase shift versus attenuation of a rotary-vane attenuator is being determined.

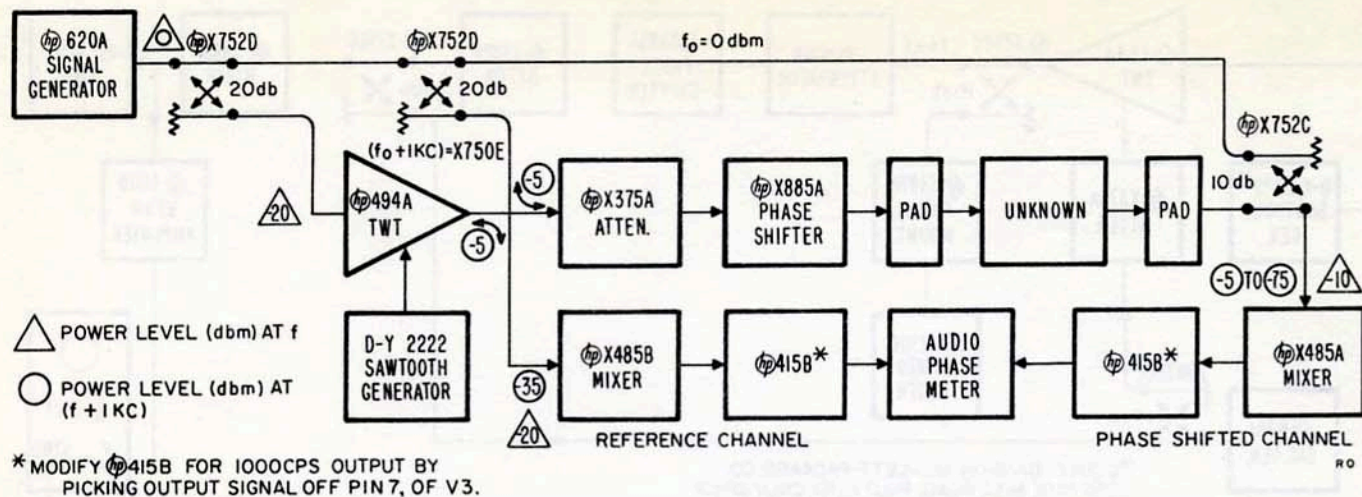


Figure 5. Typical Setup (X-Band) for Microwave Phase Measurements

Two different methods may then be used to measure phase shift:

- 1) Reading phase shift directly on the audio phase meter or,
- 2) by balancing phase shift with the 885A and using the phase meter strictly as a nulling device.

In the first method, the unknown attenuator is set for 0 db and the X885A is adjusted for a 0° reading on the phase meter. Then, the audio phase meter indicates the phase change as the attenuation is varied. In general, the amplitude of the 1000 cycle signals to the phase meter is not critical. However, when the attenuation is varied over a wide range, it may be necessary to increase the gain of the X415B in the measurement channel.

The second method uses the phase meter as a null detector. After a null is obtained and a reference shift noted on the 885A, the unknown phase shift is rebalanced with the 885A. Difference in the 885A readings is then the unknown phase shift.

Overall accuracy is affected by several factors. First is the image frequency from the traveling-wave tube caused by the finite flyback time of the sawtooth generator. For instance, an image down 30 db from the

$F + 1$ kc offset frequency may cause measuring ambiguities of $\pm 0.5^\circ$. This effect is minimized by close attention to reducing the distortion of the 1000 cps beat waveform.

Secondly, ambiguities are present due to the reflections in the microwave paths at points in the line where reflections change with movements of the device. For instance, if the device to be measured for phase shift is a rotary attenuator (X382A) with a swr of 1.15 and the pad driving it also has a swr of 1.15, the maximum second order re-reflected vector could amount to a spurious phase shift of about 0.3° . This means that although the readability of the phase meter may be tenths of a degree, as soon as the rotary attenuator is moved to a new waveguide setup, it sees a different line match than the one it was calibrated at. For this reason, in critical measurements, well-matched pads such as an X372A in-line attenuator or, on a single frequency basis, a slide screw tuner should be used to provide good match.

REFERENCES

Barnett, 885A-1, Service Note

M. Magid, Precision Microwave Phase Shift Measurements "IRE Transactions, Professional Group on Instrumentation" pp 321-331, Dec., 1958.