Errata

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HP References in this Application Note

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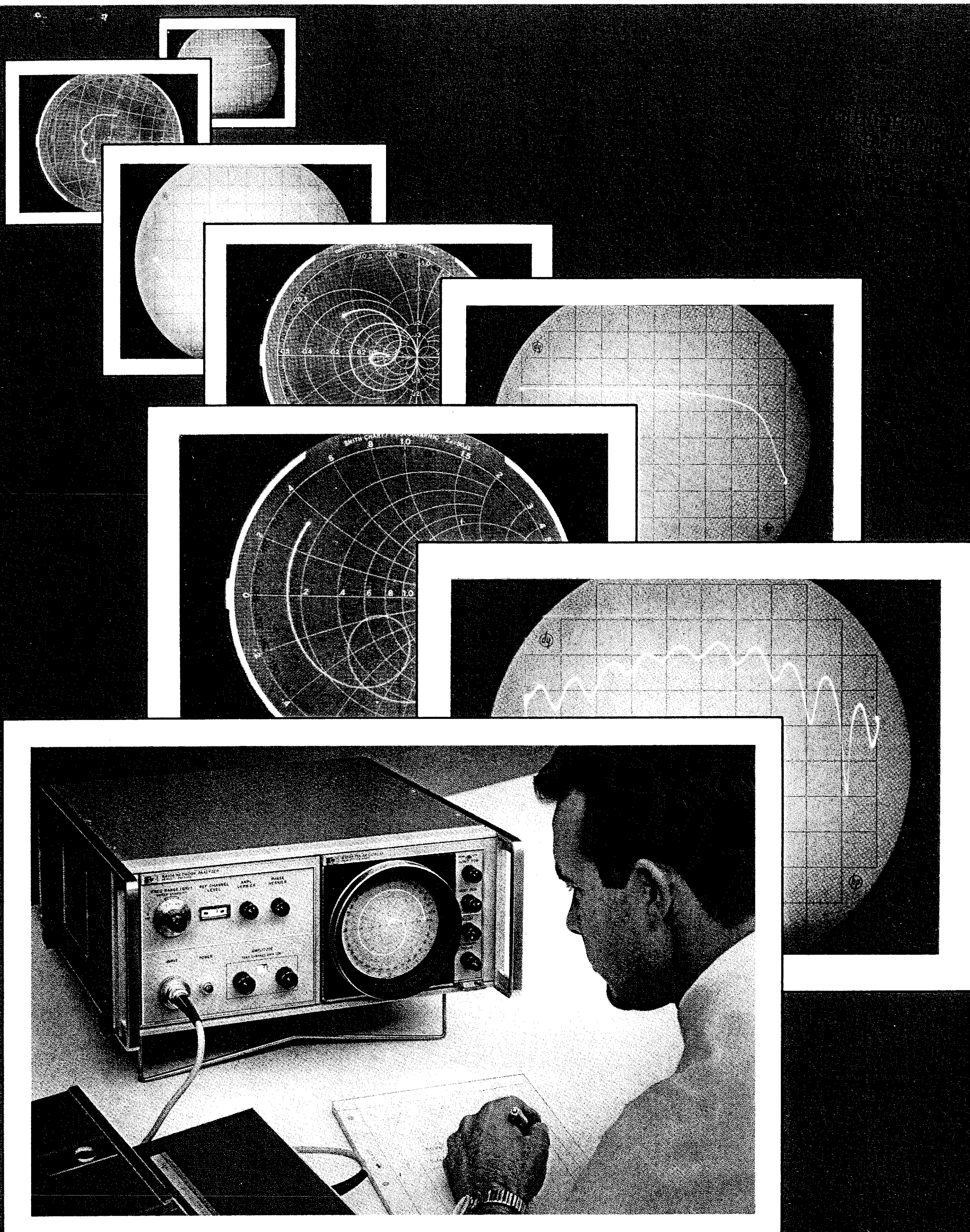

NETWORK ANALYSIS AT MICROWAVE FREQUENCIES

application note 92

HEWLETT PACKARD

NETWORK ANALYSIS AT MICROWAVE FREQUENCIES

THE NEED FOR PHASE INFORMATION

Complete characterization of microwave devices — the determination of transfer and driving point functions of active and passive components and networks — is becoming more and more necessary for the microwave engineer. Lighter and smaller microwave systems can be built by combining and cascading solid-state devices provided the designer knows exactly how each individual component or network behaves so system performance can be predicted.

Traditional measurements of frequency response (attenuation or gain) and standing-wave ratio (SWR) provide only scalar or magnitude information; they do not provide a full description of the device's performance. Since both the transfer function (transmission coefficient) and driving point function (reflection coefficient) are **vector** quantities having both magnitude and phase components, **phase shift** information is needed to characterize a network or component completely.

Phase Information in Radar Systems

The use of small, solid-state components in modern radar systems means that many of these elements can be phased for larger power outputs and thus reduce physical size and weight. It is necessary to know the phase characteristics of all these components — circulators, hybrids, modulators, etc.

Phased-array antennas that can be electronically scanned for rapid target acquisition require precise phasing of signal energy to elements in the array to provide maximum power output and to aim the beam precisely. These phased-array systems also require accurate phase matching of the transmission lines used so the signals are in the correct phase relationships needed for proper system operation.

Phase Information in Communication Systems

Many communication systems operating in the microwave spectrum are dependent on precise phase information. Phase nonlinearity in communication links causes distortion which degrades the quality of the transmitted information. Phase information is needed to maintain constant "group delay" ($d\phi/d\omega$) for all frequency components over the passband of interest.

Many sensitive receivers used in space communications use phase modulation. The phase sensitivity of receivers and related components to factors such as vibration, temperature, and time must be known.

Phase Information in Microwave Component Design

Correct phase information is absolutely necessary for successful design of microwave components. For example, consider the situation encountered in the designing of a tuned filter. It is desirable to determine the pole-zero plot of this filter. This information can be obtained from the filter's amplitude response if the dynamic range of the measuring instrument is large enough, but measuring steep skirt response is difficult in almost any application. However, the entire pole-zero response of the filter can be determined easily by measuring the phase slope through resonance of the filter. Phase-measuring capability can eliminate the need for extremely sensitive, narrow bandwidth instrumentation to characterize the response of high-Q devices over wide dynamic ranges.

IMPEDANCE

Measuring complex impedance is usually necessary in system design. Since magnitude and phase angle of the reflection coefficient are necessary to determine the reactive part of impedance, both must be determined to fully understand the device or network. For example, obtaining maximum power through a system requires a conjugate match of its cascaded or combined components. SWR measurements alone on an antenna do not determine the reactive part of its impedance and, therefore, do not provide enough information to design matching networks. Many devices perform satisfactorily only when presented with a particular impedance; consider bilateral semiconductor devices which can oscillate with certain reactive loads. The Smith Chart impedance plot can also show these regions of instability.

SCATTERING PARAMETERS

Generally, transmission and reflection coefficient measurements completely characterize any black box or network. Transmission and reflection parameters—attenuation (gain), phase shift, and complex impedance—can also be described in terms of a set of linear parameters called "scattering" or "s" parameters. Knowing these characteristic parameters, one can predict the response of cascaded or parallel networks accurately. Unlike y or h parameters, s parameters are determined with input and output ports terminated in the characteristic impedance of the transmission line (Z_0), a much more practical condition to obtain in RF and microwave work.

There are four scattering parameters for a two-port network: s_{21}, s_{12}, s_{11}, and s_{22}. The transmission coefficients, s_{21} and s_{12}, are the magnitude ratio and phase shift of the signal flow between ports when the opposite port is terminated in Z_0. s_{11} and s_{22} are reflection coefficients at the input and output ports with the opposite port terminated in Z_0. These are directly related to the impedance of the device by the equation

$$\frac{Z_{in}}{Z_o} = \frac{1 + s_{11}}{1 - s_{11}} .$$

This equation also defines the Smith Chart.

Scattering parameters are very useful in characterizing linear active devices such as transistors at RF and microwave frequencies. Using scattering parameters simplifies obtaining such figures of merit as maximum available gain, transducer gain, and frequency of unit gain.

PHASE AND AMPLITUDE TEST SET — VHF TO X-BAND

The Model 8410A Network Analyzer was designed to meet the engineer's need to define the complex parameters of both active and passive networks completely. The instrument has these characteristics:

1. It provides magnitude and phase information simultaneously between any two ports over a signal amplitude range of 60 dB.
2. It provides complex impedance information looking into any port of the device being tested.
3. It has the accuracy and resolution suitable for the characterization of high-performance networks.
4. It automatically tunes to swept signals over any octave band in the frequency range of 110 MHz to 12.4 GHz.
5. Its readout bandwidth permits swept displays of phase and magnitude variations to be observed without flicker on an oscilloscope. Changes in parameters can be observed as test devices are adjusted.
6. It has plug-in readout modules which present results in the form of attenuation/gain (dB) and phase data or combined into polar coordinates. This allows future readout flexibility.
7. It is small enough for convenient lab use.

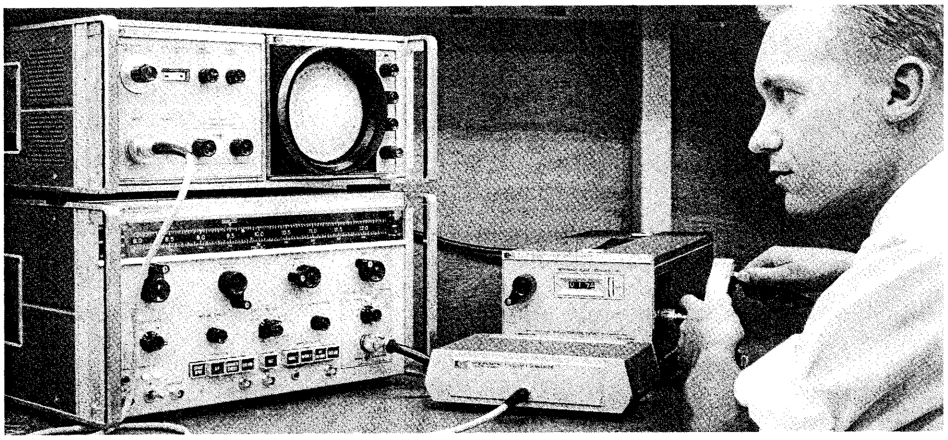

Figure 1. A swept-frequency impedance measurement being made on an X-band filter with the network analyzer. Impedance is read directly from Smith Chart overlays on the polar display unit.

This instrument system, shown in Figure 1 being used to make a swept-impedance measurement on an X-band filter, is based on the use of two wideband samplers[1] to convert the input frequencies to a constant IF frequency. RF-to-IF conversion takes place entirely in the harmonic frequency converter, (see Figure 2) which converts input frequencies over a range of 0.11 to 12.4 GHz to 20-MHz IF signals. The phase and amplitude of the two RF input signals are maintained in the IF signals. The network analyzer mainframe provides the phase-lock circuitry to maintain the 20-MHz IF frequency while frequency is being swept, takes the ratio of the reference and test channels by use of identical AGC amplifiers, and then converts down to a second IF of 278 kHz. It also has a precision 0- to 69-dB IF attenuator with 10- and 1-dB steps for accurate IF substitution measurements of gain or attenuation.

A plug-in for the 8410A mainframe compares the amplitudes of the two IF signals and provides a meter readout of their ratio directly in dB with 0.1-dB resolution. It also compares phase in degrees over a 360° unambiguous range with 0.2° resolution on the meter. Phase difference is presented on the same meter when the appropriate function button is depressed. Separate calibrated outputs (dB/mV and deg/mV) provide direct-coupled signals for simultaneous display of phase and amplitude versus frequency on an oscilloscope or X-Y recorder. In addition, higher resolution can be obtained through the use of sensitive oscilloscopes, X-Y recorders, or other voltage-indicating devices. The polar display plug-in is used in place of the phase/gain indicator to display amplitude and phase components in polar coordinates on a cathode-ray tube. This is especially convenient for reflection coefficient measurements. CRT overlays are provided for Smith Chart readout of normalized impedance or admittance. Outputs proportional to the horizontal and vertical deflections are available for X-Y recorder Smith Chart plots for highest resolution.

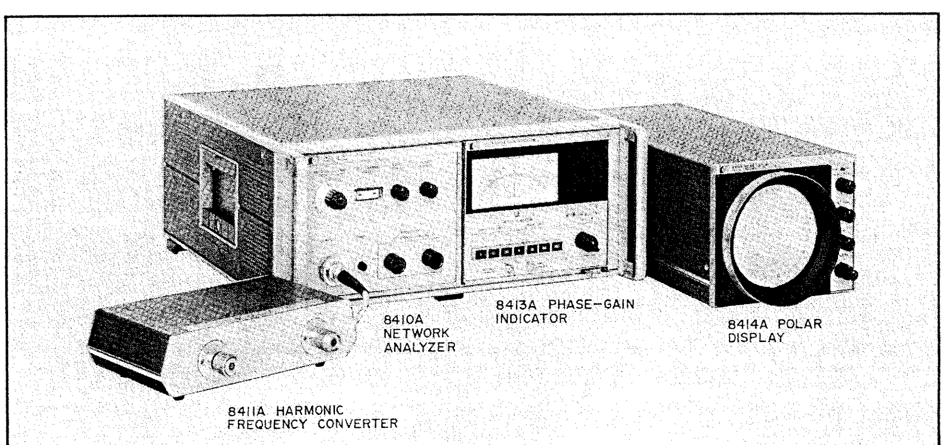

Figure 2. Hewlett-Packard Model 8410A Network Analyzer.

NETWORK ANALYZER APPLICATIONS

DETERMINING TRANSMISSION AMPLITUDE AND PHASE

Figure 3 shows a transmission test setup for measuring phase shift and attenuation or gain through a device.

Operation of the system is as follows. The RF signal source is a sweep oscillator which provides either a single frequency or sweeping input signal to the transmission unit. Due to the AGC amplifiers in the network analyzer, leveled power is unnecessary provided the source power variation is less than 20 dB over the band, as it generally is from swept-frequency oscillators.

The transmission test unit is a combination power splitter and line stretcher that has an operating frequency range from dc to 12.4 GHz. The line stretcher allows **electrical** length adjustments of up to 30 cm for phase balancing any difference in electrical length between the test and reference channels when a device is inserted in the test channel. Ten centimeters of **mechanical** extension in the transmission unit reference channel is provided to compensate for the physical length of the test device. To compensate for physical length of an unknown beyond the ten centimeters range of the mechanical extension, sections of coaxial air line or cable are inserted in the reference channel.

Transmission tests usually fall into one of three categories—insertion, incremental, or comparative. Insertion tests measure the difference in magnitude and phase before and after the test device is inserted into the measurement system. Incremental tests are made with the test device already inserted in the system and measure phase shift versus such factors as frequency, temperature, or time. Comparative tests determine the relative phase and magnitude difference between the test device and a standard inserted in the reference channel.

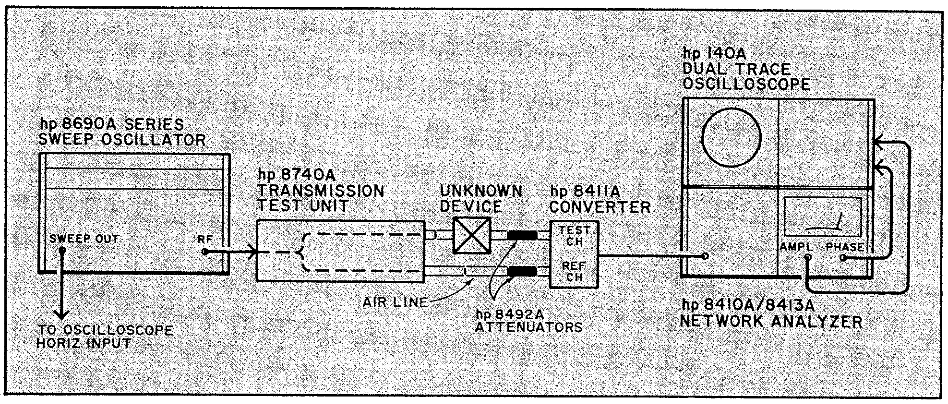

Figure 3. Typical test setup for measuring both the phase shift and attenuation or gain through a device.

Insertion Tests

Figure 4A shows the insertion phase and attenuation versus frequency of a PIN diode RF modulator. Amplitude and phase references are first set on the network analyzer and the PIN line is inserted in the test channel. If the unknown device is electrically longer than the reference channel, there will be a phase lag as the test frequency increases as shown in the bottom trace of Figure 4A. Phase lag can be measured directly in degrees from the oscilloscope display, or the calibrated line stretcher of the power splitter can be adjusted to cancel the linear phase slope. This feature provides two capabilities. First, once the linear slope is removed, small phase variations can be resolved by expanding the phase scale with the oscilloscope sensitivity controls. This is possible since the calibrated outputs from the 8413A Phase/Gain Indicator read directly in units of dB/mV and degrees/mV. Thus, the oscilloscope range switch converts the display to a desired scale in units of dB or degrees per centimeter. Secondly, the excess electrical length of the PIN modulator can then be read directly in centimeters on the transmission unit as shown in Figure 4.

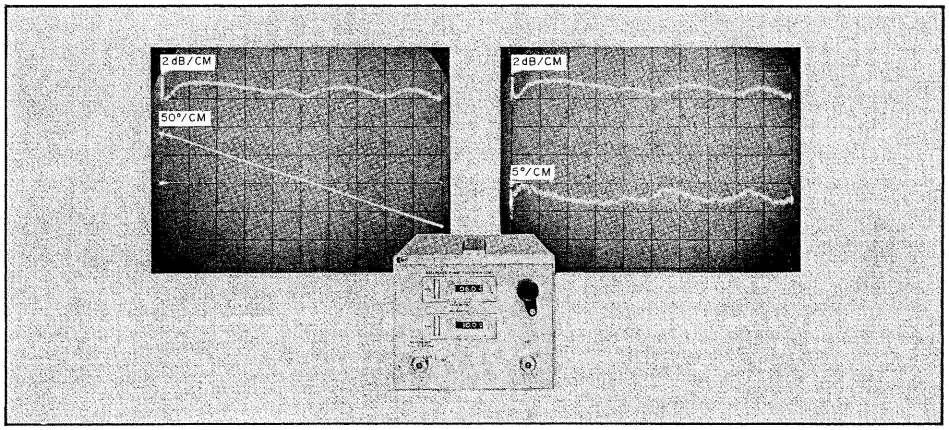

Figure 4A. Insertion phase and attenuation of a PIN modulator showing both amplitude frequency response and the linear phase shift due to excess electrical length.

Figure 4B. The excess electrical length of the PIN modulator of Figure 4A has been compensated for by the transmission unit line stretcher. The phase display now shows the "nonlinear" portion of the phase transmission. Note the phase scale change to 5°/cm. The excess electrical length over coaxial air line is read from the transmission unit directly as 6.0 cm.

The line stretcher can remove enough line from the test channel to compensate for the excess electrical length in the unknown device. The total electrical length of the system under test is the sum of the physical length of air line in the reference channel and the equivalent extension of the line stretcher in the test channel that gives constant phase versus frequency on the oscilloscope display. The remaining phase ripple displayed on the scope is the nonlinear phase shift. Stable and accurate readings of phase and amplitude are assured by the use of rigid coaxial air lines in the reference channel.

Group delay is the time by which signal energy is delayed when traveling through a device. This parameter is an important one to control in communications systems where any change in time delay over a frequency band of interest introduces distortion. Time delay information is necessary to determine delays from components in radar systems that might degrade overall performance of the system.

Group delay is related to the phase shift—frequency ratio by the relationship

$$\text{Group delay} = t_D = \frac{1}{360°}\,\frac{\Delta\phi}{\Delta f}\,.$$

An easy way to use the network analyzer to determine group delay for an electrically long device is to measure the frequency band (Δf) that results in a $\Delta\phi$ of 360°. Therefore, from the above equation, $t_D = \frac{1}{\Delta f}$.

Shown in Figure 5 is a traveling-wave tube amplifier tested for delay. A 360° phase shift over a bandwidth of 118 MHz indicates that the TWT has a delay of $10^6/118 = 8.5$ ns. In general, group delay can be measured in terms of time units by measuring the slope of phase versus frequency. The use of an operational amplifier for taking the derivative of phase speeds the measurement.

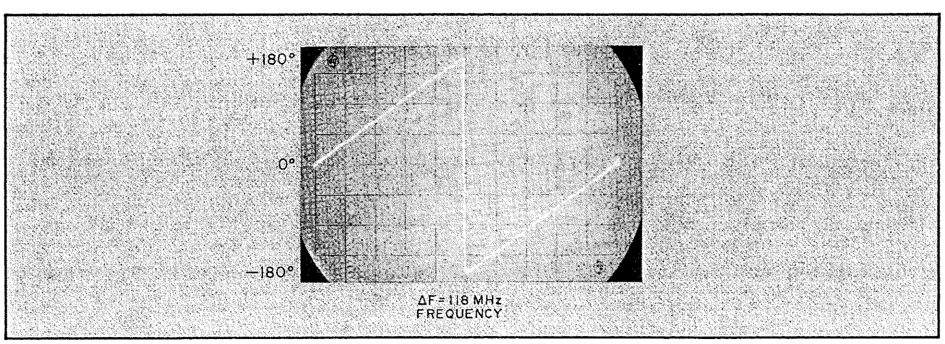

Figure 5. TWT amplifier measures total group delay. The sweep oscillator bandwidth is adjusted for 360° of linear phase shift. Group delay then reduces to the equation: $t_D = 1/\Delta f$. Here $\Delta f = 118$ MHz, indicating 8.5 nanoseconds of delay.

Incremental Tests

Many applications require incremental measurements of phase shift as a function of bias current, temperature, vibration, input power, and the like. Some typical examples are: phase linearity of paramps, phase stability of amplifiers, and cables. This type of test ignores the residual loss and phase shift of the device; thus, only arbitrary 0 dB attenuation and 0 degree phase references need be established with the test device connected in the setup.

A typical test result is shown for the phase and attenuation characteristics of the PIN modulator versus bias current shown in Figure 6. Note that phase information is obtainable over the full 60 dB dynamic range of the network analyzer. Figure 7 shows the phase characteristics of an S-band TWT versus frequency for various anode voltages. Minimum phase nonlinearity, in this case, can be obtained by the proper selection of anode voltage.

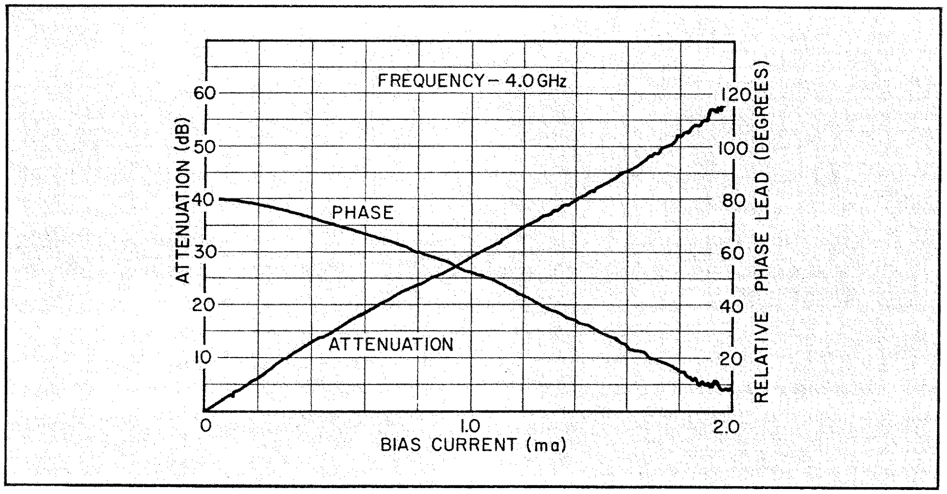

Figure 6. Attenuation and relative phase lead versus bias current for PIN modulator (4.0 GHz).

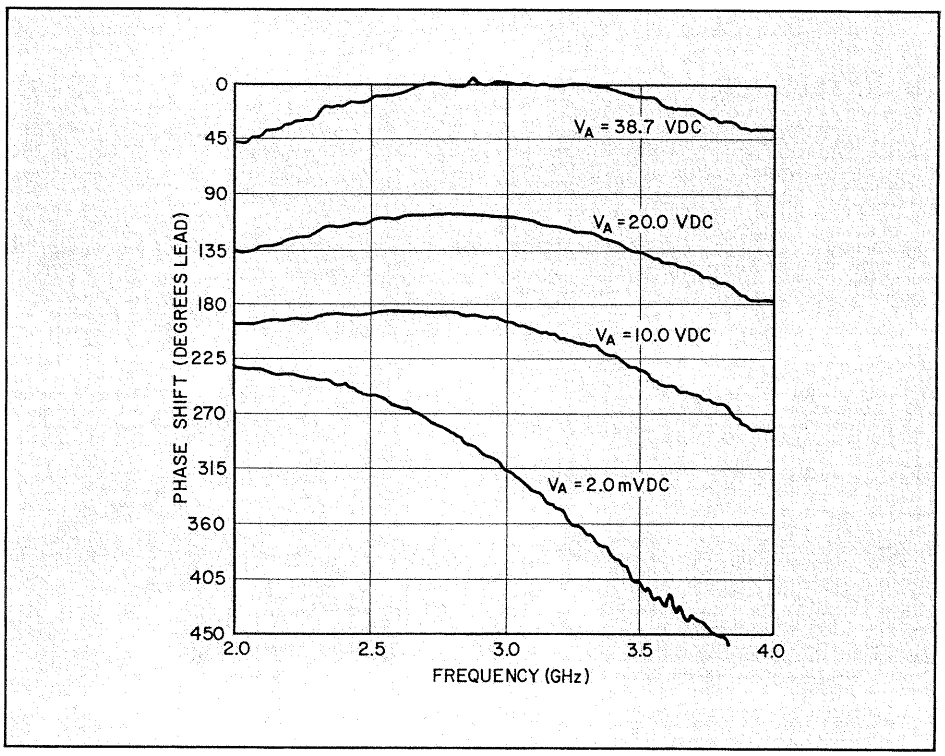

Figure 7. TWT amplifier phase shift versus frequency at four values of anode voltage.

Response time of the network analyzer is fast enough to measure amplitude and phase variations of up to about 10 kHz. This feature is an important one in many applications requiring phase or amplitude data that vary with time. An interesting use to which the network analyzer's fast response time has been applied is the measurement of solid propellant burning rate using a Doppler shift principle.[2] As shown in Figure 8, if an incident RF signal of frequency f_i is applied to a moving surface, the reflected signal undergoes a frequency shift that depends upon the rate at which the surface is moving away from the incident signal. Measuring this frequency shift by observing the beat-frequency difference limits the resolution of the measurement since at least a half cycle of the beat frequency must be recorded before the frequency shift can be determined. This problem is overcome, however, by continuously measuring the phase angle between the incident and reflected signal. From the Doppler effect, the burning rate is equal to the rate of change of phase with respect to time by the relationship

$$r = \frac{dx}{dt} = \frac{\Delta\phi\ \lambda\rho}{\Delta t\ 720°} \bullet$$

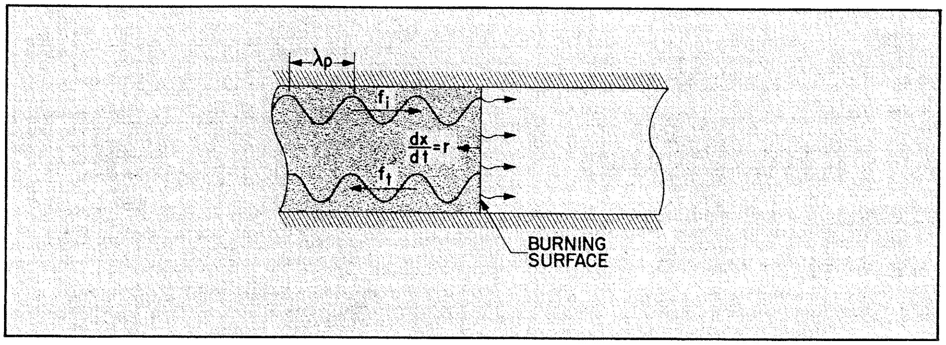

Figure 8. Continuous phase monitoring allows the direct measurement of solid propellant burning rate by using the Doppler shift principle.

Figure 9 shows a block diagram of the system for this measurement. The output of the phase meter is differentiated by an operational amplifier to obtain $d\phi/dt$ and applied to an analog tape recorder during the propellant burning run.

Phase information is also necessary in the design of a tuned filter where it is necessary to determine the pole-zero plot of the filter. This information can be obtained from the filter's amplitude response if the dynamic range of the measuring instrument is large enough, but measuring steep skirt response is difficult in almost any application. However, the entire pole-zero response of the filter can be determined easily by measuring the phase slope through resonance of the filter. Phase-measuring capability can eliminate the need for extremely sensitive instrumentation, with its inherent narrow bandwidth, to characterize the response of high-Q devices over wide dynamic ranges.

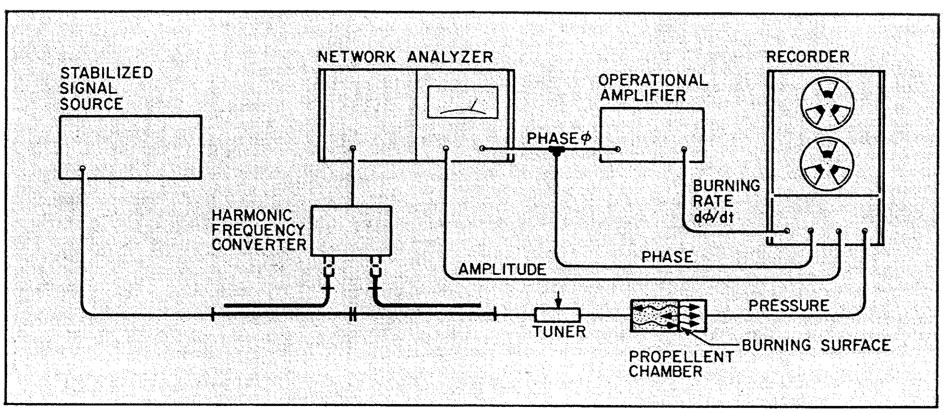

Figure 9. Block diagram showing system for measuring burning rate of a solid propellant using the Doppler shift principle.

An application for phase-versus-time measurement is in the testing of receivers, amplifiers, and other components of high performance communications systems requiring low phase sensitivity to environmental conditions. Figure 10 shows the phase "jitter" of a UHF power amplifier as it is being vibrated on a shake table. The vertical scale is calibrated directly in degrees per centimeter. The horizontal scale is time and is related directly to the frequency of vibration by synchronizing the sweep time of the oscilloscope to the vibration rate. This multiple exposure shows phase jitter of about 0.5° occurring, as expected, when the shake table reverses in direction and applies the greatest amount of acceleration.

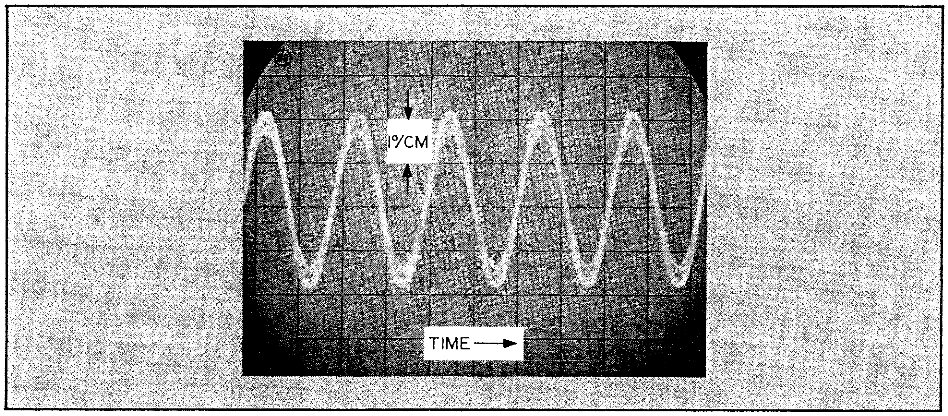

Figure 10. Phase jitter of a UHF amplifier under vibration test. The amplifier frequency is 450 MHz. Vertical scale: 1 degree/cm. Horizontal Scale: .01 second/cm. At a vibration rate of 50 Hz, a 0.5° peak phase jitter is measured at the peak of the vibration cycle.

Comparative Tests

An example of this type of test is shown in Figure 11. Here two S-band amplifiers are being matched for phase and gain tracking over this frequency band. One amplifier is placed in the reference channel and the other in the test channel. Note that phase and gain differences can be read directly off the oscilloscope. This result is typical of the many phase and gain matching applications in monopulse radar development. Comparisons to hundredths of a dB and a few tenths of a degree are possible.

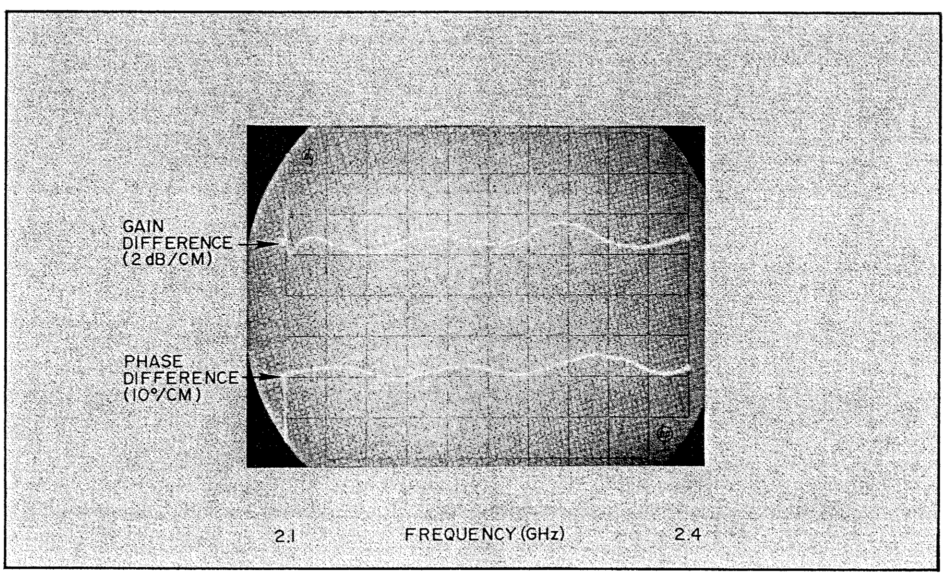

Figure 11. Phase shift and gain difference between two S-band amplifiers used in monopulse tracking system. Differential measurement provides for phase and gain matching between units on a dual trace oscilloscope.

IMPEDANCE MEASUREMENTS

Impedance is such an important design parameter for the RF and microwave design engineer that it is probably one of the most common measurements made. Knowing just the standing-wave ratio is not always sufficient; many times the design engineer must know both the magnitude **and** phase angle of the impedance so that the reactive part can be determined and compensated for with matching networks to improve circuit performance.

Determining complex impedance with a slotted line at each single frequency is a slow and tedious process. However, the network analyzer becomes a swept-impedance plotter by changing to a reflection unit and plugging in the polar display unit. This combination provides a direct Smith Chart display of impedance or admittance.

Figure 12 shows the equipment setup for this measurement. The power splitter shown earlier is replaced by a reflection unit containing a broadband dual directional coupler and line stretcher. Two of these coaxial reflectometer units allow the system to cover frequencies from 110 MHz through X-band. Incident and reflected signals from the reflection unit are measured in a manner similar to that used for the reference and test signals for transmission tests. The only difference is that now the amplitude ratio and phase angle between the reflected and incident signals that are measured are directly related to the impedance of the unknown device. The polar display unit plugged into the analyzer mainframe gives direct readings of reflection coefficient and angle or of normalized impedance with a Smith Chart overlay.

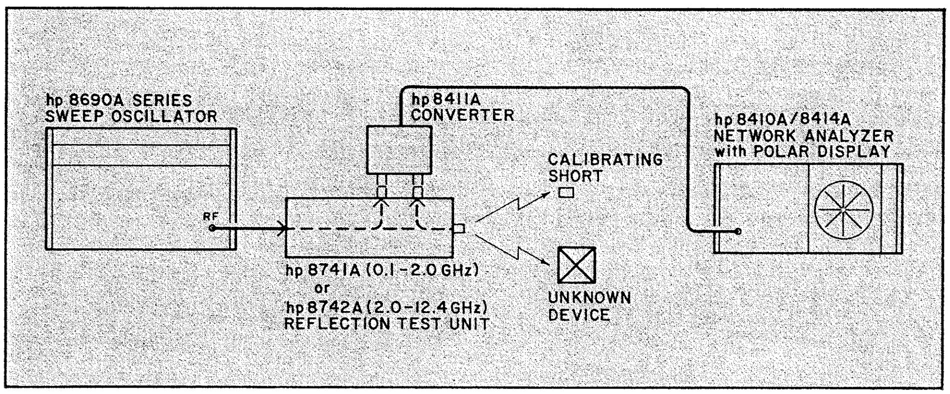

Figure 12. Reflectometer setup for Smith Chart impedance measurements. Readout of both the reflection coefficient magnitude and angle can also be made directly from polar display unit.

The system is calibrated with a short connected to the test port of the reflectometer. Test channel gain and phase vernier controls on the analyzer are then adjusted to give a full-scale indication of reflection coefficient of a short, which is equal to $1\angle 180^\circ$. The reflection unit line stretcher is adjusted to balance out the difference in electrical length between the incident and reflected signals when calibrating with the short. The reflection coefficient scale can then be expanded or compressed using the calibrated IF gain control. Frequency markers generated by the sweep oscillator appear on the polar display as bright dots, giving an indication of frequency. Recorder outputs make high resolution Smith Chart plots possible with a standard X-Y recorder.

Rapid Matching of Input Impedance

Swept-impedance tests can be made with the network analyzer on both active and passive devices. Figure 13A shows the input impedance of a TWT amplifier when swept about a 2.0-GHz center frequency. Adjustable frequency markers in the sweep oscillator intensify the trace to provide a marker at 2 GHz. The continuous swept display speeds testing, provides continuous data versus frequency, and allows impedance changes to be observed as other parameters, such as the amplifier input drive for example, are changed.

Rapid impedance matching of this TWT amplifier is possible with a swept-impedance display. In this application it was desired to match the amplifier to 50Ω at 2.0 GHz. With a tuner in front of the TWT, a quick adjustment of the tuner moves the 2-GHz marker to the center of the Smith Chart and provides a 50Ω match. The display then appears as in Figure 13B. Besides saving a great amount of time over conventional techniques using VSWR meters and slotted lines, the effect of the tuner on input impedance at other frequencies within an octave range is also immediately observable with a swept technique.

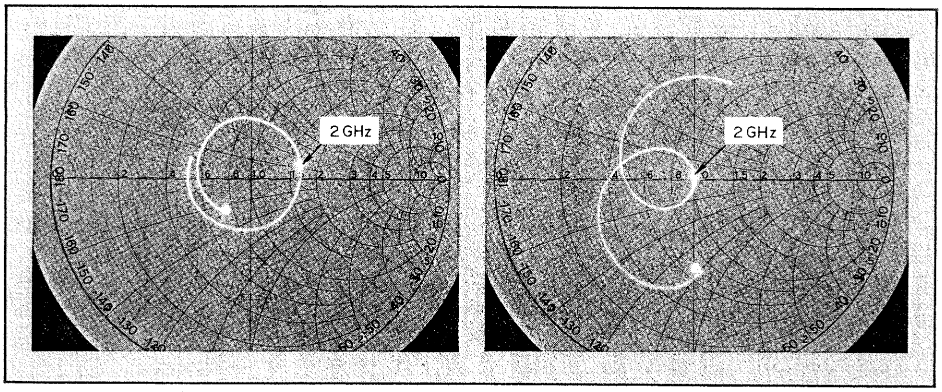

Figure 13A. Swept display of the input impedance of a TWT amplifier. The center frequency is 2.0 GHz (note bright marker on display). Full scale $\Gamma = 1$.

Figure 13B. Continuous swept display of the impedance of the TWT amplifier allows quick and easy adjustment of a tuner on the input of the amplifier to match it to the characteristic impedance of the transmission line at 2.0 GHz. Note that the effects of the tuner on the TWT's input impedance at other frequencies is readily apparent on the swept display.

Impedance Transformation

A common design problem is to minimize the mismatch of a device by a suitable compensating design. In many cases, however, the plane of the mismatch is not directly accessible to the test port of impedance instrumentation except through a length of transmission line. Normally, this requires correcting the measured impedance to reflect the electrical length of the transmission line between the test port and the desired impedance test plane at **each** frequency. This lengthy calculation can be avoided by use of the reflection unit line stretcher. stretcher.

Figure 14 shows the impedance, at the input connector, of a thin resistive film coaxial attenuator. It is desired to improve the mismatch by appropriately compensating the resistive element of the attenuator. The impedance appears primarily capacitive at the input to the device, but the attenuator film is actually located 3.16 centimeters beyond the input connector. It is at this point that the impedance should be determined since this is where design adjustments will be made.

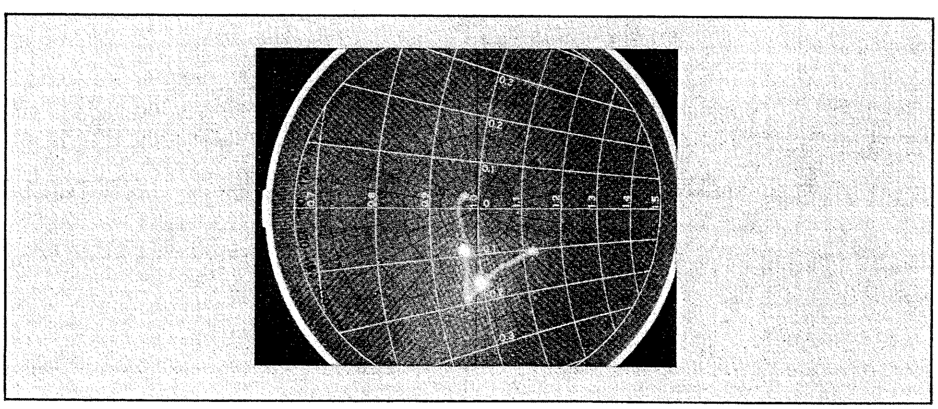

Figure 14. Swept display of input impedance of a coaxial attenuator at the plane of the connector. An expanded Smith Chart overlay for direct readout of impedance to a full scale $\Gamma = 0.2$ (VSWR = 1.5) is used. Frequency range: 10.0 – 12.0 GHz.

The length of the line stretcher in the incident arm of the reflection unit is now reduced, which effectively extends the test plane of the unknown beyond the test port of the reflection unit. The unit is calibrated to read directly in centimeters of reference plane extension. Figure 15 shows the impedance of the attenuator of Figure 14 when the reference plane was extended 3.16 centimeters by the line stretcher. Note that only the phase angle of the reflection coefficient has changed by the appropriate value of electrical phase shift at each particular frequency. The magnitude of the reflection coefficient does not change. The display now shows the true impedance at the actual plane of the resistive element. Thus, an appropriate compensating network can be designed to reduce the attenuator's mismatch. Alternately, impedance can be determined at any point in a line so that compensating tuning stubs or networks can be inserted where convenient.

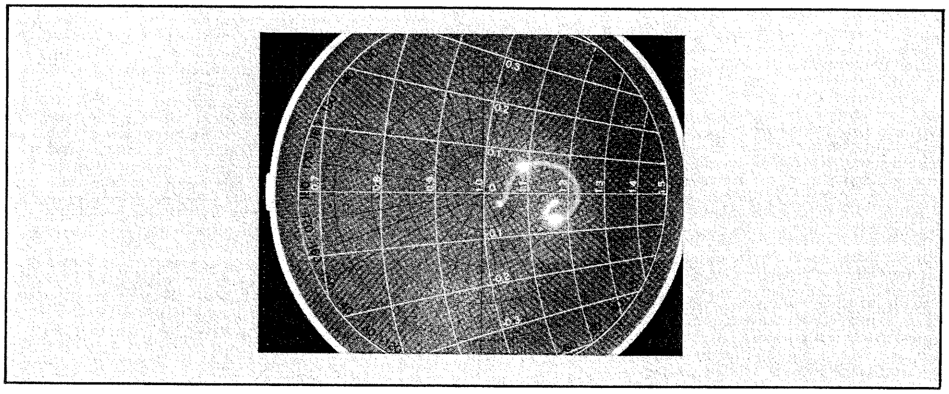

Figure 15. The same impedance measurement of the coaxial attenuator of Figure 14 but with the plane of measurement extended 3.16 cm into the attenuator where the resistive element is located. Impedance can be determined anywhere in the attenuator for design compensation considerations, for example, by using the line stretcher in the reflection unit to shift the plane of measurement the desired amount. Frequency range: 10.0-12.0 GHz.

Scale Expansion.

For impedance measurements near 50 Ω, as in the last example, it is possible to expand the polar display to obtain better resolution. The procedure is as follows. After calibration for a full-scale reflection coefficient of $1\underline{/180°}$, the scale is expanded using the network analyzer test channel gain control. Thus, for an expanded Smith Chart overlay corresponding to a full-scale reflection coefficient of 0.2 (or 1.5 SWR), the IF test channel gain is increased by $-20 \log (0.2)$ or 14 dB from the calibration setting. The expanded Smith Chart is now laid over the polar display unit to provide more accurate readings near 50 Ω.

Tunnel Diodes and Negative Impedance.

Engineers who design oscillators with tunnel diodes are vitally concerned with the impedance characteristics of this very useful device. The resistive part of the impedance presented by the tunnel diode must be negative to sustain oscillations.

Negative impedance measurements can be made with the network analyzer in two ways. Figure 16 shows a Smith Chart overlay designed for negative im-

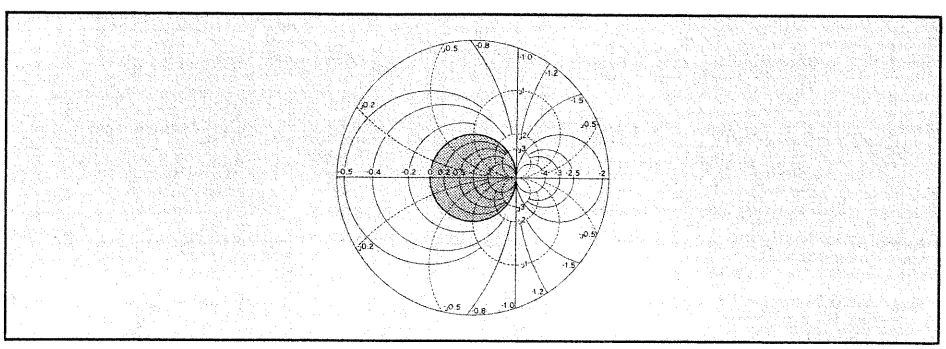

Figure 16. Compressed Smith Chart overlay available for direct readout of negative impedance. Full scale $\Gamma = 3.16$. ($\Gamma > 1$ indicates negative real part of impedance).

pedances for a full scale Γ of 3.16. (Γ's greater than 1 represent negative impedance). The shaded area in this Smith Chart represents the normal or positive values of resistance. The area outside this area represents $\Gamma > 1$ and, thus, negative resistance. This overlay is used in a fashion similar to that of the expanded Smith Chart except that now the IF test channel gain is decreased by $-20 \log (3.16)$, or -10 dB from the calibration setting for a reflection coefficient of 1.0.

Another technique for displaying negative impedance offers more expansion for Γ's ranging from 1 to 10. Smith[3] has shown that a plot of $1/\Gamma$ can be represented on the normal Smith Chart provided the positive real axis is interpreted as **negative** real values. The network analyzer can be made to plot $1/\Gamma$ simply by reversing the inputs of the harmonic frequency converter so that the reflected signal, which is now greater than the incident signal for a negative resistance, is applied to the **reference** channel while the incident signal is connected to the **test** channel. A slight modification to the polar display unit, which simply reverses the voltage to the vertical deflection plates, displays the correct polarity of the reflection coefficient. Range of Γ is limited to a maximum of about 10 due to the reference channel dynamic range of 20 dB. Other methods can be used to expand this range but a Γ of 10 or less is typical of most negative resistance devices.

Figure 17 is a swept-impedance plot of a tunnel diode over the 110-MHz to 1.15-GHz frequency range. The frequency at which the impedance becomes positive real and, thus, where the tunnel diode can no longer oscillate, is readily measured as 1.08 GHz. It is that point on the trace crossing the outer perimeter of the Smith Chart.

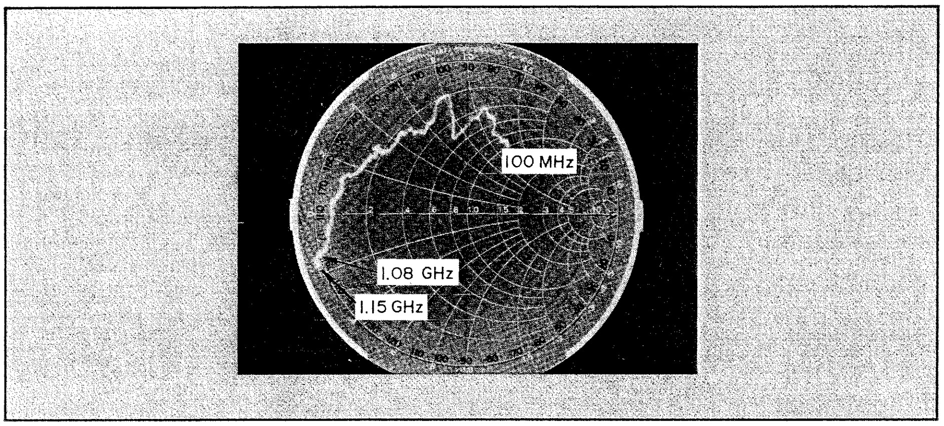

Figure 17. Swept-impedance plot of a tunnel diode obtained with the harmonic frequency converter inputs reversed to give negative Smith Chart plot of $1/\Gamma$. Resistive coordinates are now read as negative values. Frequency range is 110 MHz – 1.15 GHz. The frequency at which the resistance becomes positive is where the trace leaves the Smith Chart (Γ becomes less than 1) and, in this case, is 1.08 GHz. The tunnel diode will not sustain oscillations above that frequency.

Correcting Impedance Measurements for Directivity Error.

Coupler directivity is one of the limiting errors in accurate reflectometer impedance or SWR measurements, particularly near 50 Ω. This is because there is usually no information about the **phase** of the directivity vector with respect to the phase of the unknown's reflection vector.

However, it is possible to eliminate this error effectively with the network analyzer. This can be cone by connecting a coaxial sliding load to the reflectometer output. If the load is moved continuously back and forth at least one wavelength at a given frequency, a circle will be traced on the polar display as shown in Figure 18. The distance from the center of the polar display to the center of the circle being traced represents the magnitude of the directivity vector error. The angle of this vector is read directly from the radial calibration. A convenient way to correct for the directivity error directly on the polar display is to adjust the vertical and horizontal positioning controls on the polar display while moving the sliding load until the circle being traced is exactly in the center of the display unit. Directivity is then effectively cancelled from the readout, the sliding load can be removed, and the unknown measured.

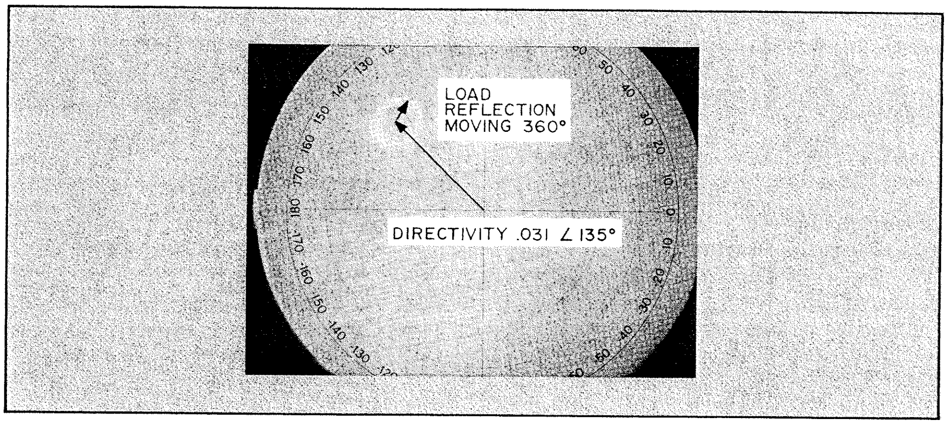

Figure 18. Coupler directivity calibration obtained by phasing a sliding load through one wavelength at a single frequency. The directivity vector magnitude is the distance from the center of the display to the center of the small circle obtained when the load is phased through 360°. Full scale $\Gamma = 0.05$. Directivity reading is $.031\underline{/135°}$. Frequency is 10.0 GHz.

This error correction technique is very effective when the network analyzer system is controlled by a computer. The computerized system can automatically and rapidly calibrate out directivity error at many fixed frequencies across the band. This can lead to reflectometer system directivities of about 70 dB, resulting in very accurate impedance measurements near 50 Ω.

WAVEGUIDE MEASUREMENTS

The wide frequency range of the network analyzer makes it compatible with the broad bandwidth of coaxial transmission lines. Precision 7-mm connectors were chosen to reduce mismatch ambiguities at all critical points in the RF measuring circuits as well as to define reference planes accurately. However, the network analyzer is not limited to coaxial measurements only; two methods have evolved for using it in waveguide tests.

The first method will be mentioned only briefly; it has evolved from the use of an instrumentation computer with the network analyzer. In this method, a waveguide-to-coax adapter is connected to the unknown port of the coaxial dual directional coupler to convert to waveguide tests. A calibration routine automatically measures the error in magnitude and angle due to the reflection from the adapter and stores it in the computer's memory. All subsequent measurements are corrected automatically by the computer to eliminate the adapter ambiguity. It is beyond the scope of this article to describe in detail the computer techniques used with the network analyzer. Further information is available from the Hewlett-Packard Company, Microwave Division, Palo Alto, California.

Figure 19 shows the system arranged for waveguide impedance tests in X-band. Power out of the sweep oscillator is divided into two equal channels

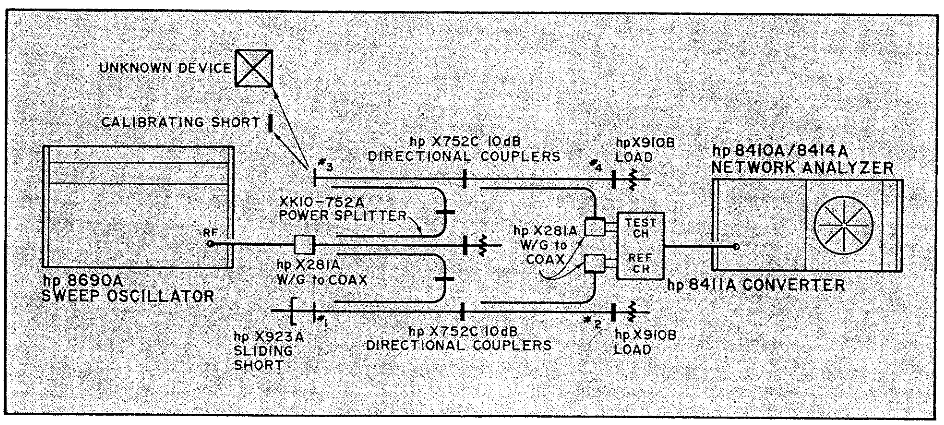

Figure 19. X-band waveguide system for measuring impedance on CW and swept-frequency basis. The sliding short acts as a line stretcher and the added couplers, because of their padding effect, effectively reduce the mismatch ambiguity of the waveguide-to-coaxial adapters.

by a special waveguide directional coupler having two auxiliary arms. The reference channel signal feeds through the auxiliary arm of a 10-dB directional coupler to a movable short. The short reflects all of the signal down the main line of the directional coupler into a second coupler and load. The electrical length of the reference channel can be adjusted with the movable short. This adjustment provides a variable reference plane and enables calibrating to a constant phase reference for swept-frequency tests. The second coupler effectively "pads" any mismatch caused by the waveguide-to-coax adapter connected to the harmonic frequency converter reference channel input.

The test channel signal feeds through the auxiliary arm of a third directional coupler (as shown in the diagram) to the unknown port. A waveguide shorting plate is connected to this port for initial calibration, reflecting all of the signal down to the fourth coupler which is attached to the test channel input of the converter.

Small tracking variations and source mismatch of the system cause the trace to move about as frequency is swept from 8.2 to 12.4 GHz. These variations appear during rapid sweeps as a larger spot size, which introduces a reflection coefficient ambiguity of about 0.02 (for swept-frequency measurements only). In terms of SWR, this is only 1.04. Angle ambiguity at a 0.2 reflection coefficient is about 6°. These ambiguities, however, can be calibrated out at single frequencies by manually adjusting the calibration reference as exactly $1\underline{/180^\circ}$ at each frequency of interest.

A similar setup can be used for waveguide transmission tests to obtain reference and test channel signals. A waveguide line stretcher is usually needed when testing individual components, but may be omitted when matching two identical components for amplitude and phase characteristics.

ANTENNA TESTING

The network analyzer can be used for antenna pattern and frequency response tests in an anechoic chamber as it has enough sensitivity to qualify as a range receiver over the short distances involved. In one test, with approximately 1 watt radiated a distance of 10 meters to a standard-gain pickup horn, the signal level at the analyzer test channel input was about-24 dBm, well above the -78 dBm noise level of the analyzer. The instrument can measure the antenna's phase characteristics, and it can be used to measure antenna impedance as well.

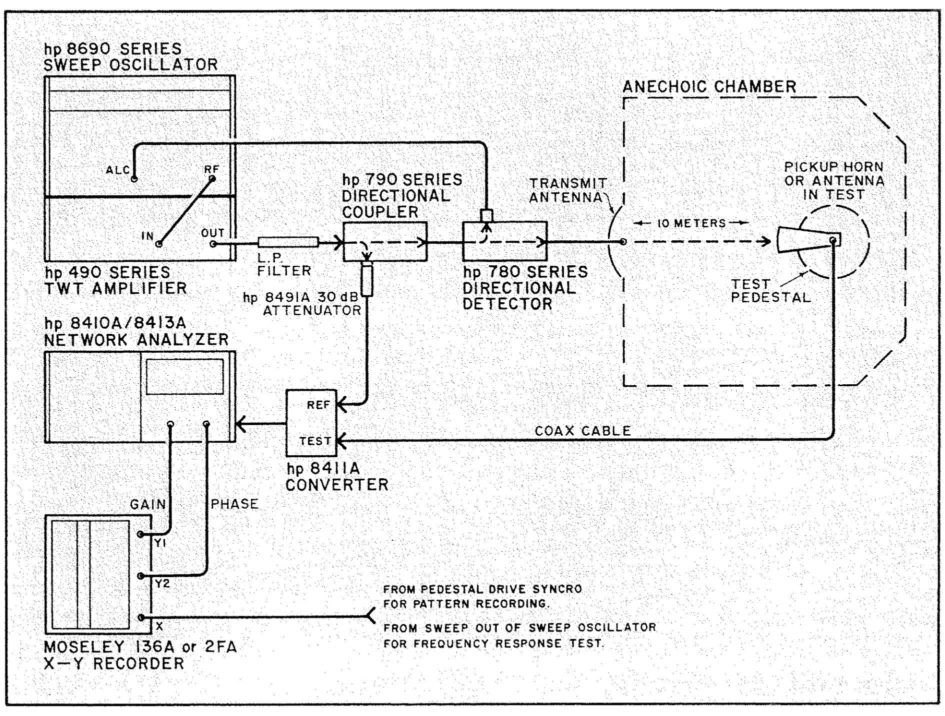

Figure 20. Network analyzer system for making CW and swept-frequency antenna tests in an anechoic chamber.

Figure 20 shows how the network analyzer can be coupled with normal range equipment for antenna tests. A rotation test of a standard gain horn at 3 GHz resulted in the phase-gain pattern shown in Figure 21. Similar swept-frequency techniques can be used to evaluate anechoic chamber characteristics.

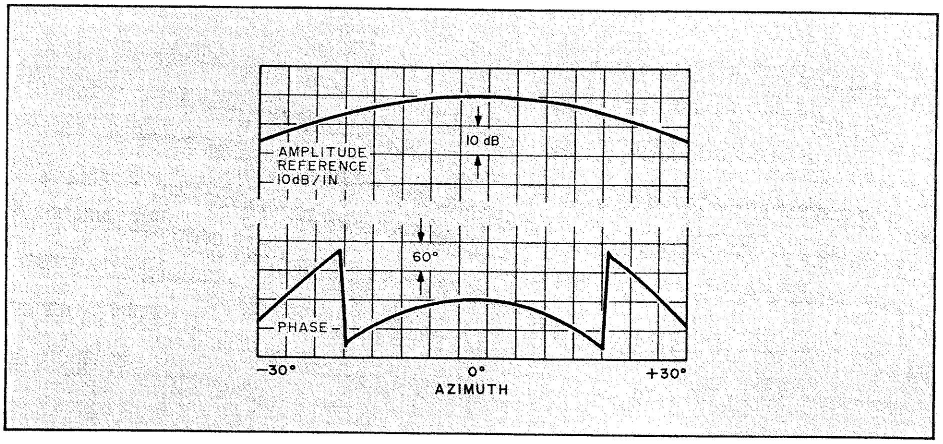

Figure 21. Plot of gain and phase response of a standard horn antenna as a function of azimuth. Scale: Amplitude — 10 dB/division, Phase — 60 degrees/division, Frequency — 3 GHz.

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1. W. Grove, "Sampling for Oscilloscopes and Other RF Systems DC Through X-band", International Microwave Symposium, 1966, Palo Alto, California.
2. V. Shelton, "A Technique for Measurement of Solid Propellant Burning Rates During Rapid Pressure Transients", Jet Propulsion Laboratory, California Institute of Technology, Pasadena, California, Internal Report.
3. P. H. Smith, "A New Negative Resistance Smith Chart," **Microwave Journal**, June 1965.