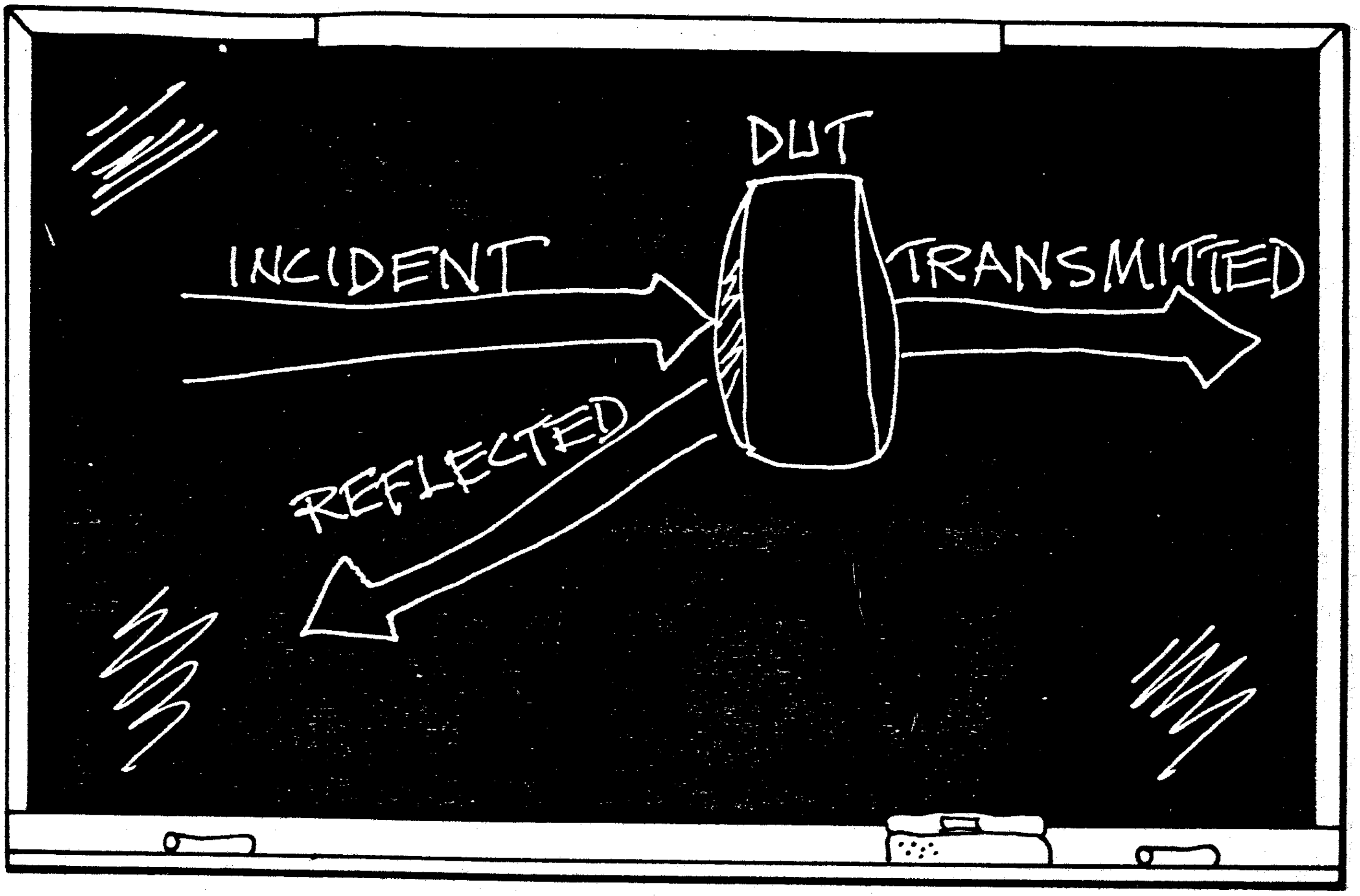
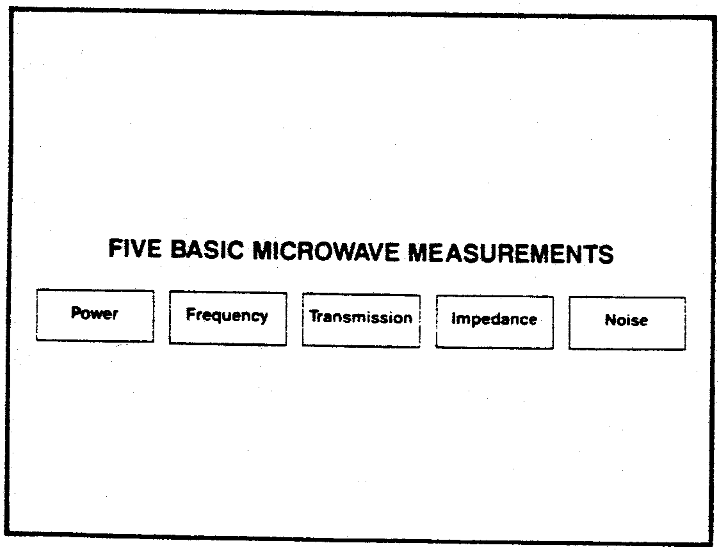
HEWLETT
PACKARD

Introduction to the Vector Network Analyser & HP 8753A

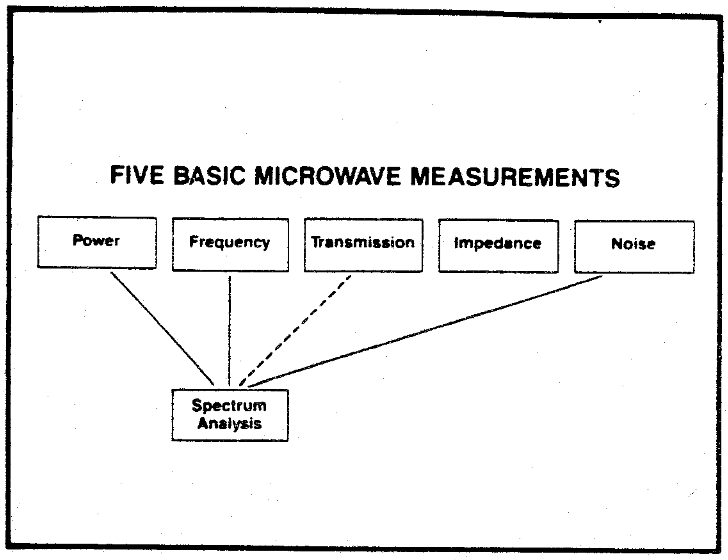



FIVE BASIC MICROWAVE MEASUREMENTS

Power | Frequency | Transmission | Impedance | Noise

There are five basic microwave measurements.

Spectrum analysis primarily measures power, frequency, and noise.

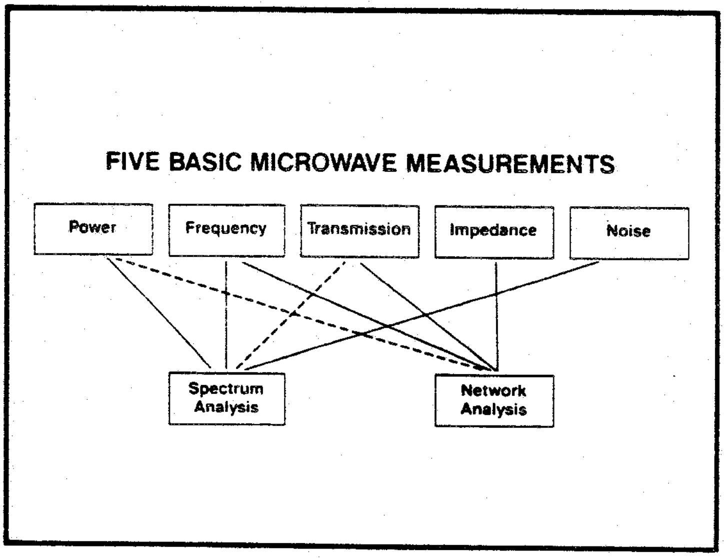


FIVE BASIC MICROWAVE MEASUREMENTS

Power
Frequency
Transmission
Impedance
Noise

Spectrum Analysis

Network analysis is primarily concerned with impedance, transmission, and frequency. Power can also measured in network analysis.



FIVE BASIC MICROWAVE MEASUREMENTS

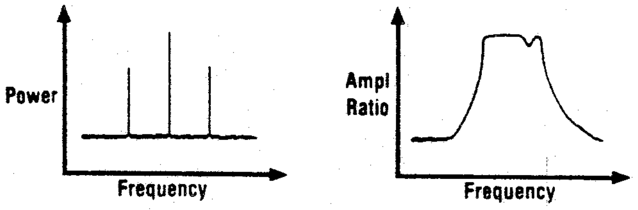
Power
Frequency
Transmission
Impedance
Noise

Spectrum Analysis
Network Analysis

The major difference between spectrum and network analysis is illustrated here.

Spectrum analysis is primarily concerned with characterizing a signal; i.e., its spurious and harmonic components, modulation, noise, etc. It indicates discrete frequencies where microwave energy exists.

In network analysis, we want to characterize a microwave component; that is, determine how efficiently energy is transferred into the network (or out of the network), or measure its transmission characteristics to determine how effectively energy is transferred through the network. Frequency is important since parameters are usually measured and displayed as a function of frequency.

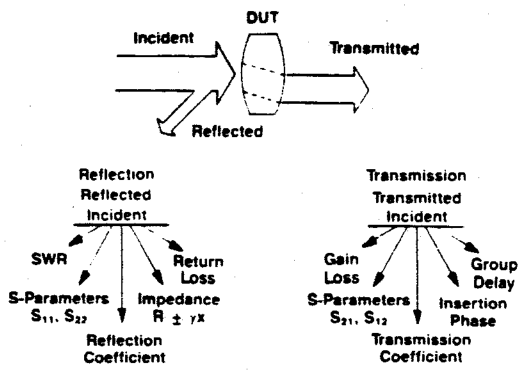


Typical Spectrum Analyzer Display

Typical Network Measuremants Display

SIGNAL CHARACTERISTICS
- Communications
- Receiver Testing
- Signal Generation Tests

NETWORK CHARACTERISTICS
- Network Test
- Component Test
- System Test



Incident
DUT
Transmitted
Reflected

Reflection
$\frac{\text{Reflected}}{\text{Incident}}$
SWR
S-Parameters
S_{11}, S_{22}
Reflection
Coefficient
Impedance
$R \pm jX$
Return
Loss

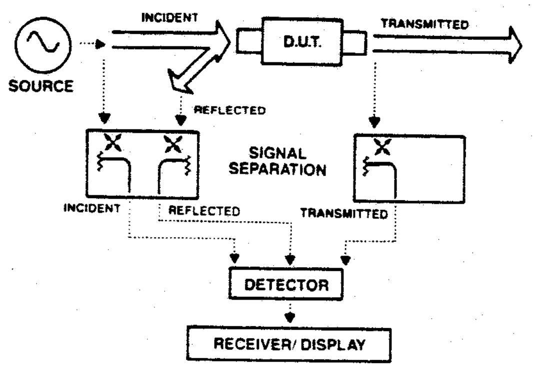
Transmission
$\frac{\text{Transmitted}}{\text{Incident}}$
Gain
Loss
S-Parameters
S_{21}, S_{12}
Transmission
Coefficient
Insertion
Phase
Group
Delay

Microwave energy can be likened to light energy. Throughout this seminar we will use this analogy. Three waves will be of interest to us: the incident, the reflected, and the transmitted waves. To characterize a network (or component) completely, both magnitude and phase information is necessary. For many applications, however, scalar (magnitude only) characterization is sufficient.

WHY MAKE SCALAR MEASUREMENTS INSTEAD OF VECTOR MEASUREMENTS???

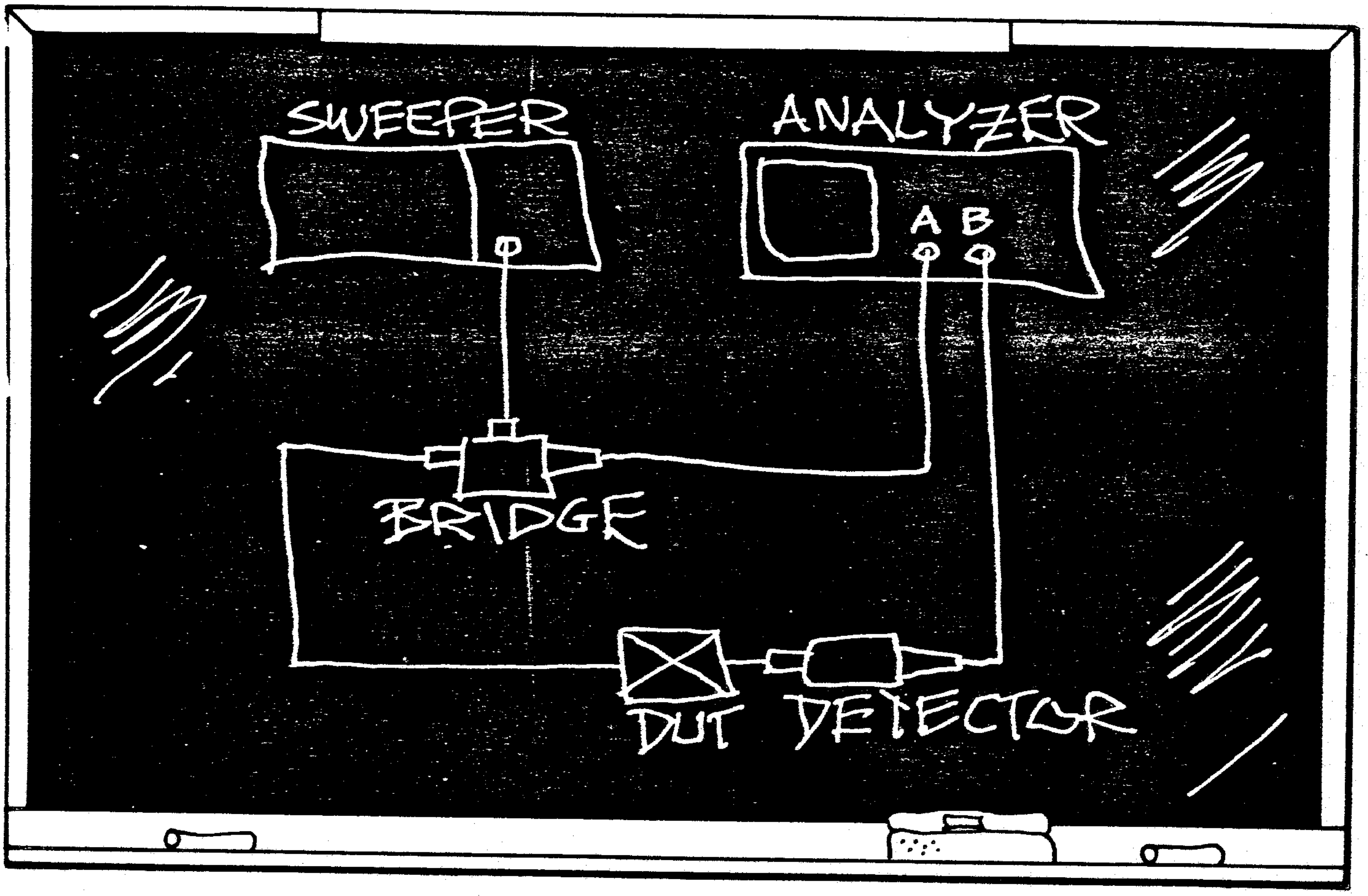
Why are we interested in making scalar (magnitude only) measurements instead of making vector (magnitude and phase) measurements?

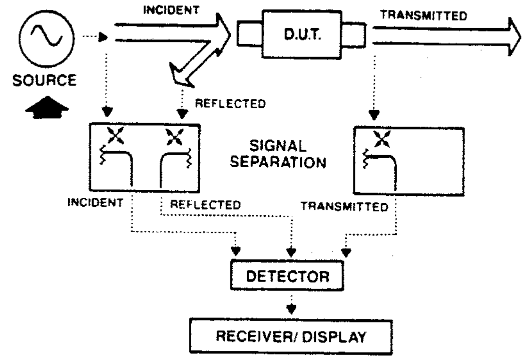
This set-up shows the typical network measurement configuration. Note the device under test is analogous to the "lens" of our previous example with light, the component which we want to characterize.



SOURCE
INCIDENT
D.U.T.
TRANSMITTED
REFLECTED
SIGNAL
SEPARATION
INCIDENT
REFLECTED
TRANSMITTED
DETECTOR
RECEIVER/DISPLAY

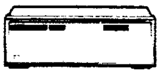
Measurement Systems



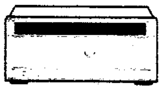


The microwave source provides the swept frequency stimulus for the measurement system.

RECOMMENDED SOURCES



SWEEP OSCILLATOR
- Broadband
- Analog Sweep

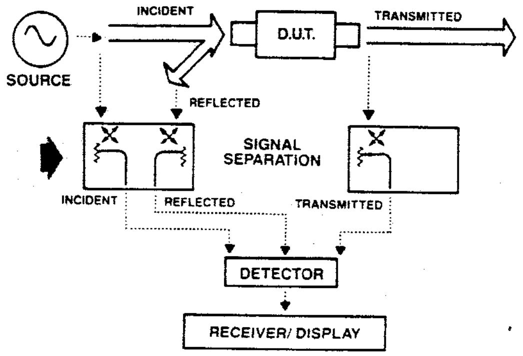


SYNTHESIZED SWEEPER
- Broadband
- Analog Sweep
- Synthesized Frequency Accuracy

The two major sources we will discuss are the sweep oscillator (more commonly called a "sweeper"), and the synthesized sweep oscillator.

The basic sweeper provides a signal (frequency) that is swept over a broad band of frequencies with settings to select the range over which the source is swept. Additional features such as internal and external power leveling, modulation (AM, FM, Pulse), and programmability are important for more accurate and cost effective systems as we'll see later.

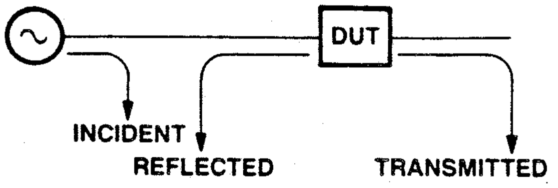
A synthesized sweeper has the additional capabilities of providing very accurate synthesized stepped CW sweeps as well as the standard broadband analog sweep. Narrowband, highly accurate, phase-locked sweeps are also possible with a synthesized sweeper.



Let's discuss signal separation devices and how they are used in network measurements.

The signal separation device samples the test signal in one direction only. For example, a directional coupler used for reflection measurements only samples the test signal reflected from the input of the DUT but not the incident signal.

SIGNAL SEPARATION DEVICES



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The three devices used for sampling the signal of interest are:(1) directional couplers, (2) directional bridges, and (3) the two resistor power splitter.

SIGNAL SEPARATION DEVICES

- Directional Coupler
- Directional Bridge
- Power Splitter

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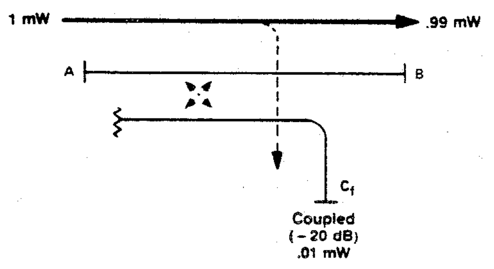
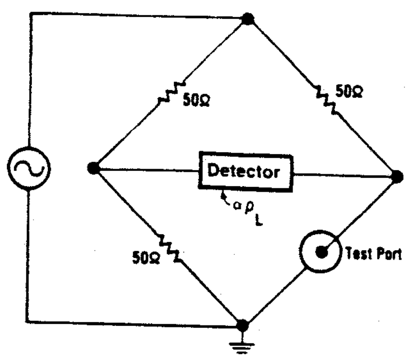
The coupled arm of a directional coupler samples a signal traveling in one direction only. The coupled signal is at a reduced level and the amount of reduced level is called the coupling factor. Notice that in this example of a 20 dB directional coupler that the coupled port is 20 dB below the input. A 20 dB reduction means that the coupled arm is $0.01xP_{in}$ or 1% of the input power. The remainder of the signal (99%) travels through the main arm. There is also a frequency response or coupling variation associated with couplers, expressed in ± dB.

The coupler schematic represents the direction that signals will be coupled, depending on which way the arrows are pointing.

DIRECTIONAL COUPLER

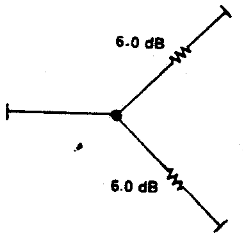


DIRECTIONAL BRIDGE



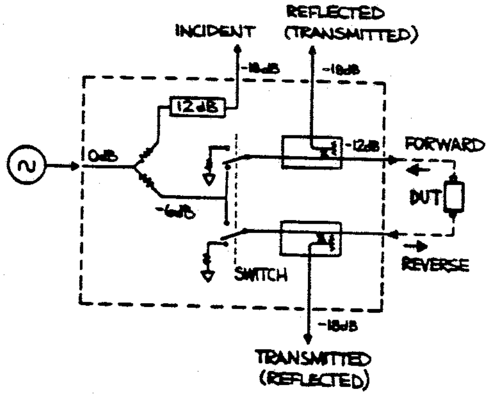
The next signal separation device used in measuring reflected signals is the directional bridge. Its operation is similar to the simple Wheatstone bridge. If all four arms are equal in resistance (i.e., test port = 50 ohms) a voltage null is measured. If the test port load is not 50 ohms, then the voltage across the bridge is proportional to the mismatch (deviation from 50 ohms) of the DUT.

POWER SPLITTER

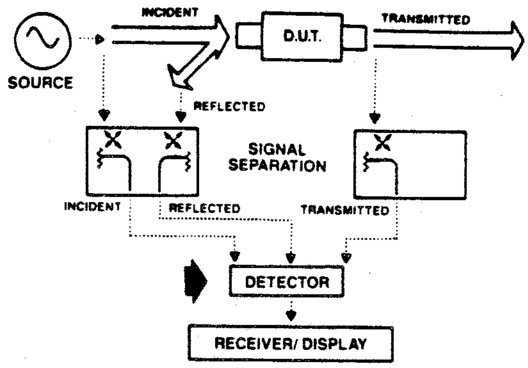


The two resistor power splitter is used to sample either the incident signal or the transmitted signal. The input signal is split equally between the two arms with the output signal (power) from each arm being 6 dB below the input. The typical microwave power splitter is broadband operating over a frequency range from DC to 26.5 GHz.

S-PARAMETER TEST SET CONFIGURATION

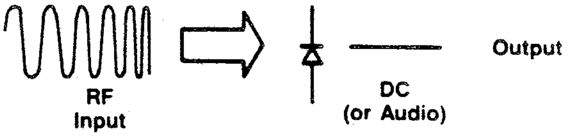


Test devices like transistors and isolators have different forward and reverse transmission functions. Characterization of these devices is greatly expedited by an S-Parameter test set which can automatically switch or reverse the power flow without disconnecting the device. This test set was designed for a three channel receiver and permits the simultaneous display of reflection and transmission. A single DPDT switch reverses power flow to the test device while maintaining a 50 ohm source and load match. Automatic measurements are possible since the switching can be dome remotely.



Diode detectors convert the RF signal to a proportional DC voltage. If the signal is amplitude modulated, the diode strips the modulation. Diode detectors can be very broadband (10 MHz to >26.5 GHz), have fast response times, and have a dynamic measurement range of up to 76 dB.

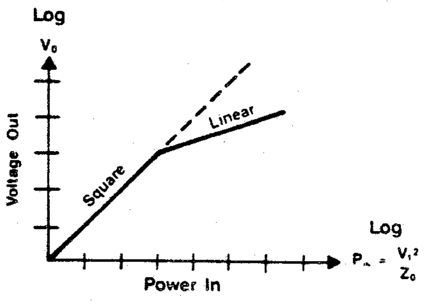
DIODE DETECTION



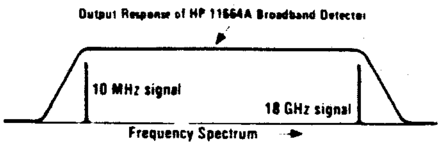
- Broadband Detection
- –60 dBm
- Economical

Diode detectors have a square law region over which the voltage out is proportional to the square of the power in. Above a certain power level the response becomes linear. When a diode detector is used with an oscilloscope to display some detected response, its measurement dynamic range is limited to the square law range of the diode. Since the "knee" is predictable and repeatable with certain diodes (the type of diode used with scalar analyzers being one), scalar analyzers can compensate for this characteristic and hence have the ability to measure responses over a larger dynamic range.

DIODE DETECTION



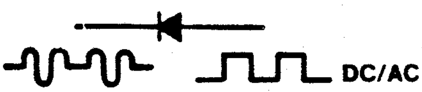
BROADBAND DETECTION AND FILTER REQUIREMENTS



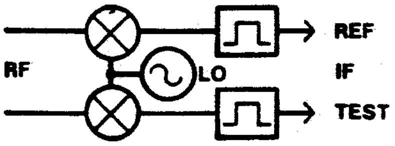
Broadband detectors will respond to any signal or signals present at the input port in the frequency range of the detector. For example, if we are interested in a network's response at 18 GHz, and in addition to 18 GHz there is also a 10 MHz signal present, the system will respond to the composite response. Therefore, in order to observe the response of the 18 GHz signal only, the unwanted signal must be removed in some way.

DETECTION SCHEMES

- **SCALAR**



- **VECTOR**

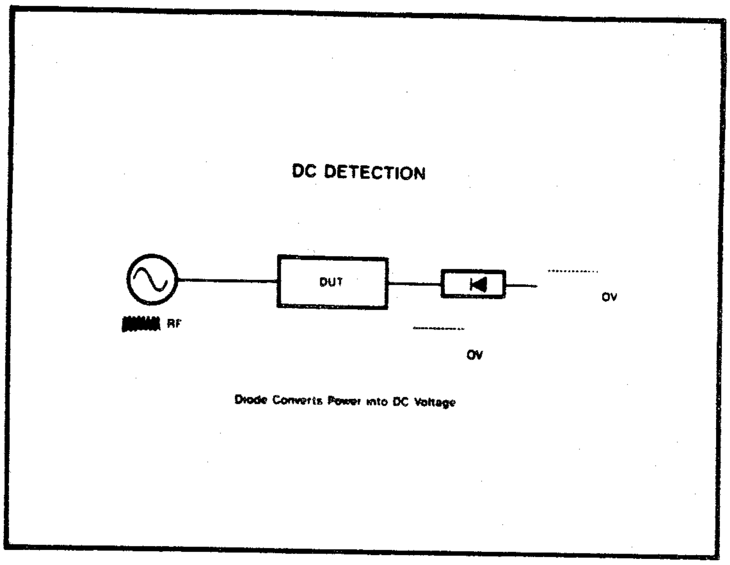


The most significant distinction betwe network analyzers is in the receiver techniq used. The two schemes are Scalar (magnitu only) and Vector (magnitude and phas capabilities). Scalar systems rely on dio technology to convert the RF to either DC or lo frequency AC signals. Vector systems use du channel mixers to downconvert from RF to maintaining phase information.

AC DETECTION
DC DETECTION

Diode detection schemes use either "DC detection" or "AC detection".

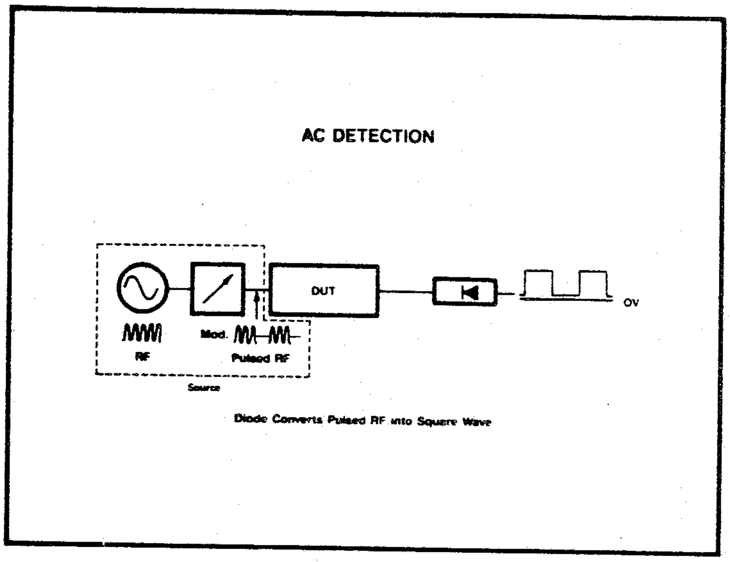
DC detection produces a DC signal that is proportional to the power incident upon the diode. The diode's output is read directly by the analyzer making the analyzer a fancy voltmeter with a logarithmic response.



DC DETECTION

Diode Converts Power into DC Voltage

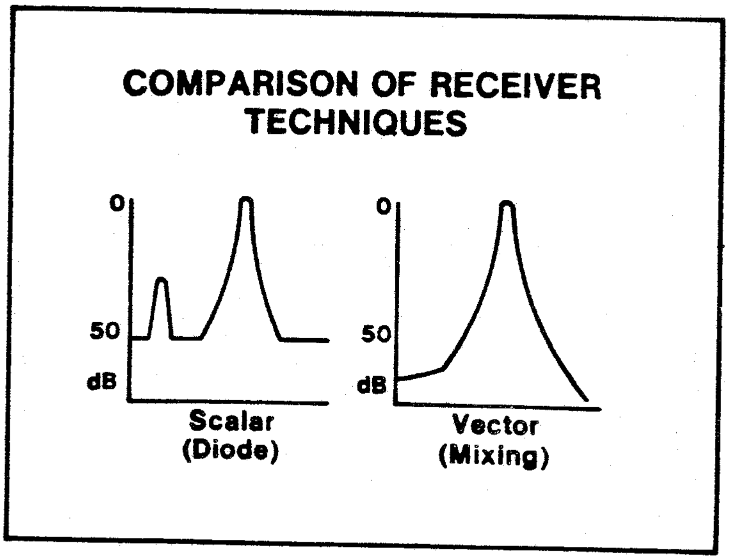
AC detection also produces a signal proportional to the power incident upon the diode. However, the RF power is modulated with a square wave signal. The pulsed RF travels through the DUT and stimulates the diode detector. The pulsed RF signal is turned into a square wave by the detector... the pulse being high when the RF is on and being low when the RF is off.



AC DETECTION

Diode Converts Pulsed RF into Square Wave

The diode detection scheme, while being broadband and inexpensive, is also susceptible to spurious signals and harmonics.



COMPARISON OF RECEIVER TECHNIQUES

DC ADVANTAGES

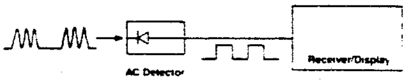
- When Modulation Affects the Measurement
  - Amplifiers with AGC
  - Amplifiers with Large Low Freq Gain
  - Narrow Band Devices (<10 MHz)
  - Power Measurements

Although AC detection is usually preferred over DC detection, DC detection does provide some benefits. The major benefit being no modulation of the RF signal to affect the DUT or the measurement results. Modulation can have adverse affects on the measurement of some devices.

AC ADVANTAGES

- No DC Drift
- Noise Immunity
- Reject Unwanted Signals
- Fast Response
- AC Detection Can Simulate DC Detection

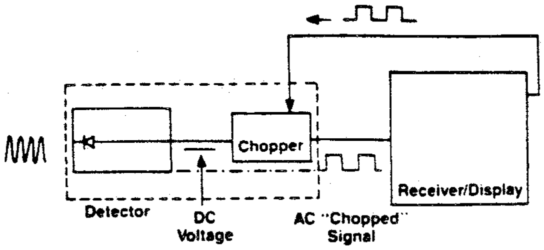
AC detection can provide many benefits over DC detection because the detector is not affected by signals at the input that are not modulated.



- Display Measurement in dB
- Convert Detected Voltage Signal to dB
- Process an AC Signal

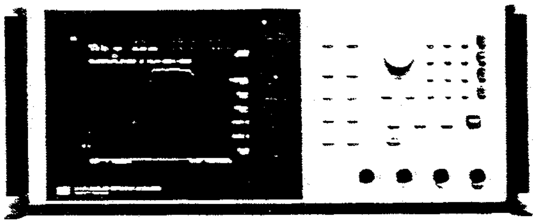
The main purpose of the receiver is to convert the voltage signal from the diode detector to a logarithmic or dB value for display. The time required to convert a DC voltage to a log value is largely dependent upon the power level incident on the detector - the lower the power level, the longer the processing time (thus the slower the sweep speed from the source). An AC voltage on the other hand is converted to a log value much faster and is not dependent upon the power level incident on the detector. HP scalar analyzers process an AC signal (27.8 kHz) to make processing time independent of incident power level.

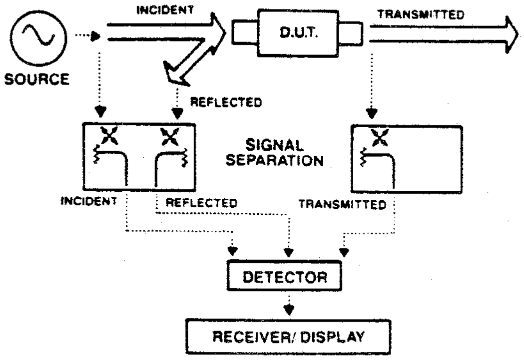
With HP's DC detectors, the DC voltage is "chopped" to create the AC signal which HP's scalar analyzers then process. Thus, with the use of HP's DC detectors, the advantages of DC detection are combined with the advantages of AC signal processing.



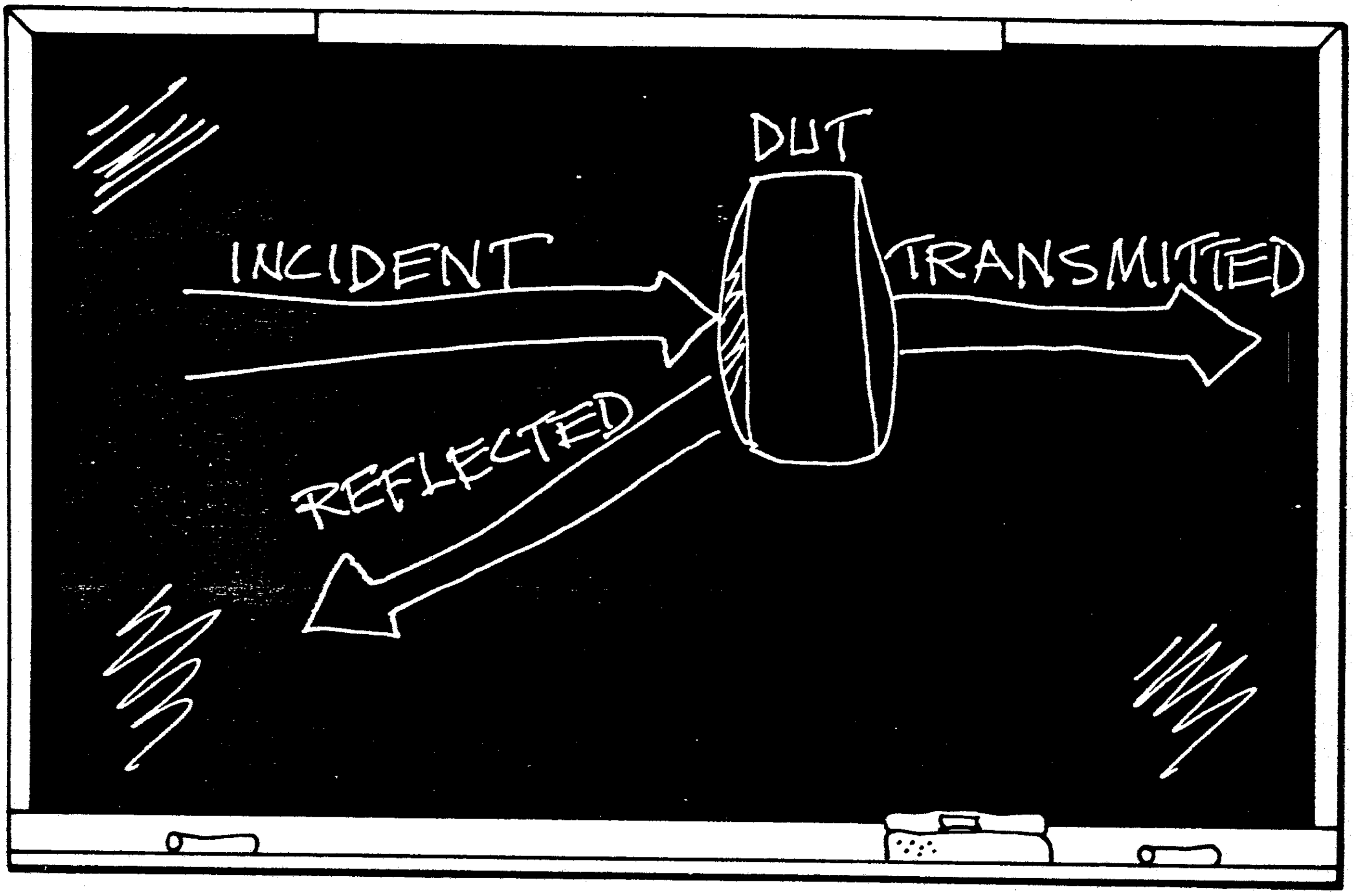
- Advantages of DC Detection
- Advantages of AC Signal Processing

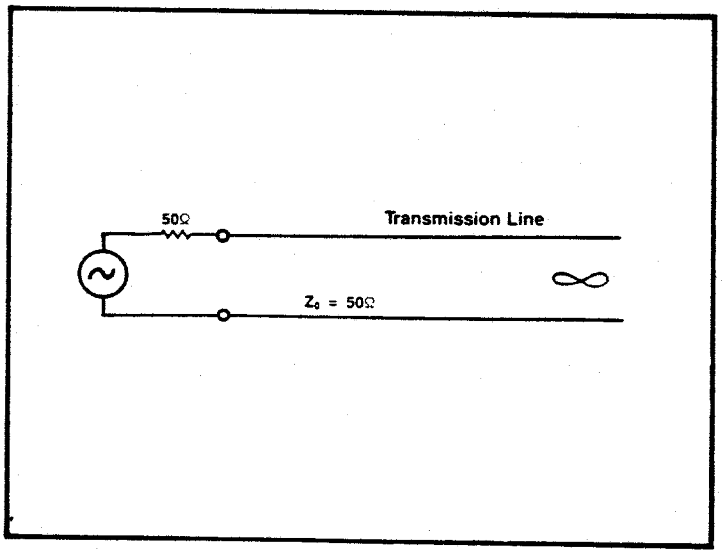
More commonly referred to as the network analyzer, the receiver/display not only processes the detected signals but also controls all of the display functions. The CRT displays measurement annotation, soft key labels, data traces and other information. Hard copies of the displayed information can be obtained by transferring the CRT information to a graphics plotter or printer directly.





Fundamentals

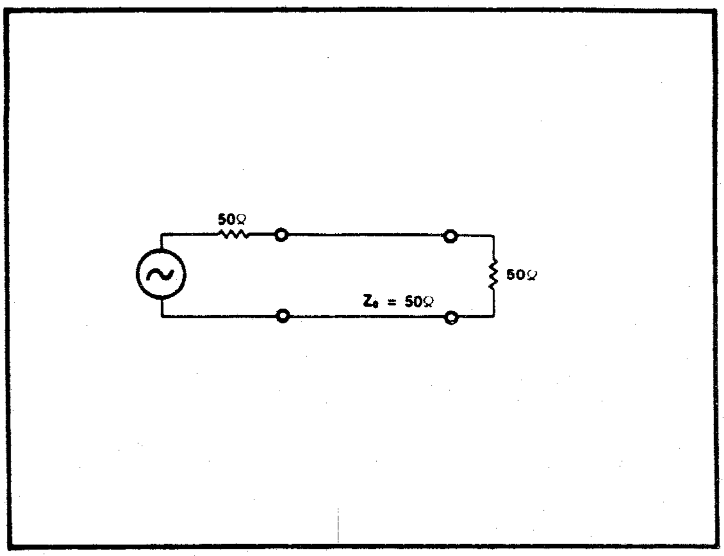




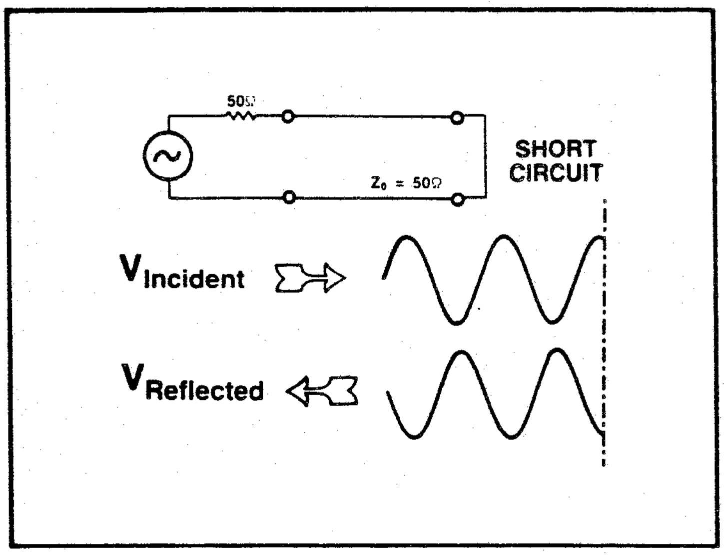
Before we discuss measuring reflected waves, let's take a look at what causes them. If we have a source of microwave energy with a source impedance of 50 ohms then we can deliver maximum power to the load if the load impedance is equal to the source impedance.

In this case, we have an infinitely long transmission line of 50 ohms (characteristic impedance).

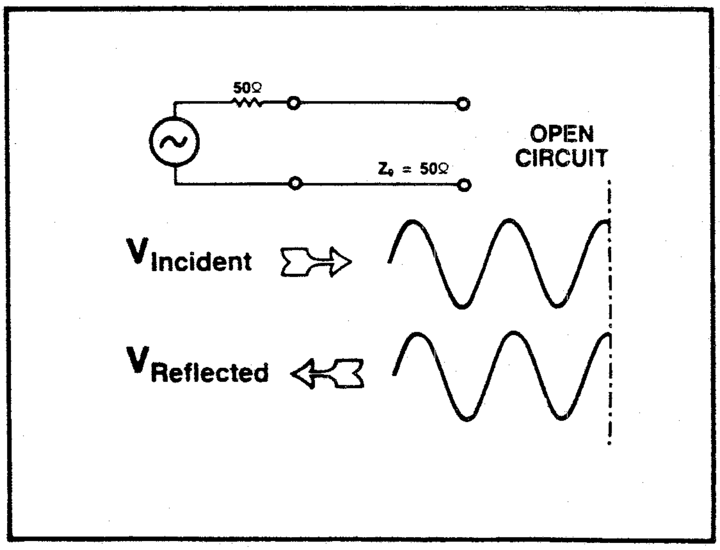
If we terminate the transmission line in 50 ohms, then the termination should absorb all of the power delivered from the source (i.e., the signal cannot tell the difference between a Zo load and a Zo transmission line of infinite length).

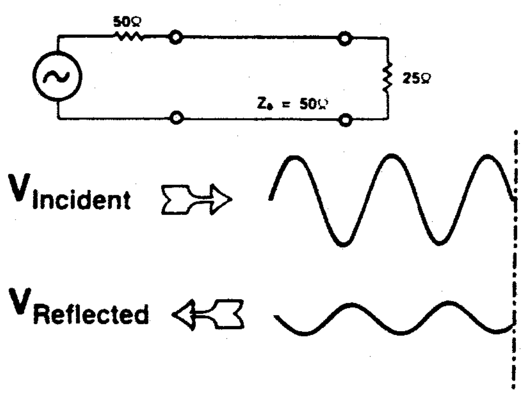


Let's terminate our line with a short circuit. Since a short circuit can dissipate no power, and since there is nowhere else for the energy to go, a "reflected" wave is launched back down the transmission line. Since the short can support no voltage, the reflected wave must be of equal magnitude to the incident wave and be 180 degrees out of phase with it (the sum of the incident voltage wave and the reflected voltage wave must equal zero at the short).



Similarly, when we terminate the transmission line with an open, there is nowhere for the energy to go (the load is an infinite impedance). A "reflected" wave is again launched back down the transmission line. Since an open can support voltage, the reflected voltage wave must be of equal magnitude and be in phase with the incident signal.





If we terminate our transmission line with a 25 ohm resistor (an impedance somewhere between an open and a short), we will find that our reflected voltage wave will have an amplitude of 1/3 of that of the incident wave and that the two waves will be 180 degrees out of phase with each other.

$$\Gamma = \frac{V_{REFL}}{V_{INC}} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

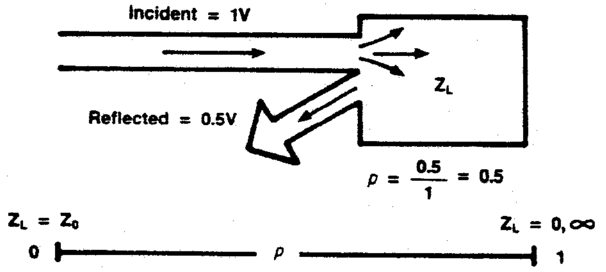
$$\rho = |\Gamma|$$

We can conclude from all of this that the reflected wave tells us something about the impedance of whatever we use to terminate the transmission line.

The exact mathematical relationship between the impedance of the termination and the reflected wave is shown on the slide. It's important because it shows clearly that since Zo is known, we can determine the load impedance by measuring Vi and Vr (the incident and reflected voltage waves).

SCALAR REFLECTION TERMS

$$\rho = \textit{REFLECTION COEFFICIENT} = \frac{\textit{E reflected}}{\textit{E incident}}$$



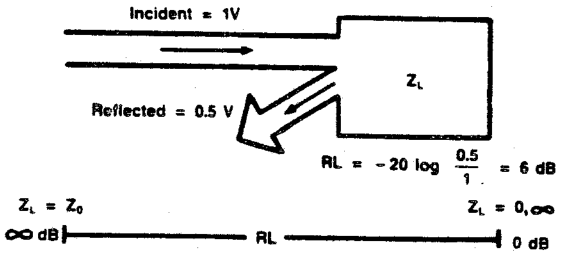
Reflection coefficient is equal to the ratio of the reflected voltage wave to the incident voltage wave. For a transmission line of characteristic impedance Zo terminated with a Zo load, all energy is transferred to the load and none is reflected: Er = 0 and ρ = 0. When the line is terminated with an open or short circuit, all of the energy is reflected and Er = Ei and ρ = 1. The range of possible values for ρ then is 0 to 1.

Since many displays are logarithmic, we need a term to express reflection coefficient in dB. Return loss can be thought of as the number of dB that the reflected signal is below the incident signal. The range of values for return loss are infinity for a Zo impedance to 0 for an open or short circuit.

For an explanation of dB's, refer to Appendix A.

SCALAR REFLECTION TERMS

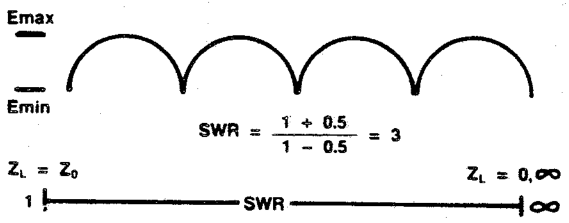
RETURN LOSS = −20 log ρ



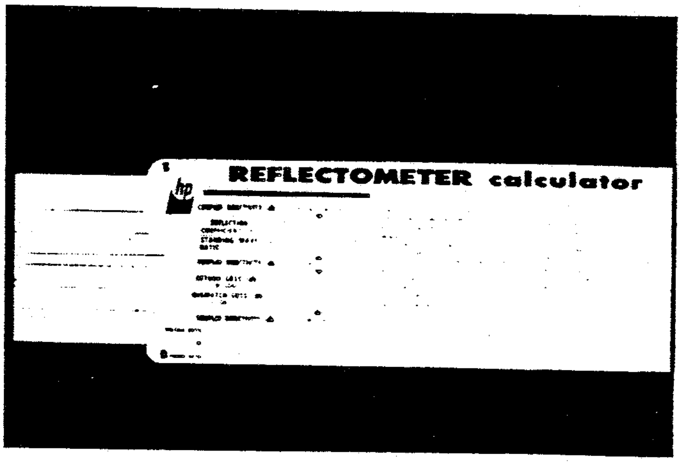
Any two waves traveling in opposite directions cause a "standing wave" to be formed on the transmission line. Standing wave ratio (SWR) is defined as the maximum voltage over the minimum voltage of the standing wave on our line. It can also be defined as $(1 + \rho)/(1 - \rho)$. The values of SWR are 1 to infinity.

SCALAR REFLECTION TERMS

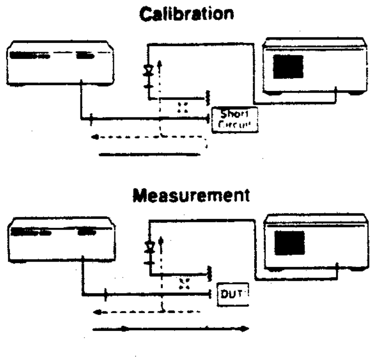
$$\text{SWR} = \frac{1 + \rho}{1 - \rho} = \frac{\text{Emax}}{\text{Emin}}$$



We can use a reflectometer calculator to convert between reflection coefficient, return loss, and SWR. For example: let ρ = 0.5 and use the calculator to determine the equivalent return loss and SWR. Move the slide portion of the calculator until 0.5 on the reflection coefficient scale (the upper scale on the slide) is directly below the blue arrow. Read the return loss value on the return loss scale directly below the blue arrow and read the SWR value on the SWR scale directly above the blue arrow.



BASIC REFLECTOMETER

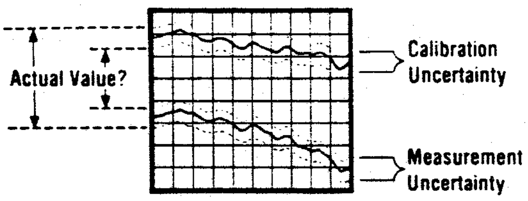


There are two parts to a reflection measurement. First, a reference line is established on the CRT with a known standard (in this case a short circuit). This trace is stored for future subtraction (normalization). Measuring the short establishes a zero dB return loss reference line.

The dB change when the DUT is inserted is the return loss of the DUT.

We assume that the source has a perfect Zo output impedance and that the signal separation device can separate the reflected signal without any leakage of the incident signal.

REFLECTION UNCERTAINTY



The actual value of our DUT is the difference between the measured RL value and the calibration (Meas - Cal). But since we do not have "perfect" measurement system components we have some uncertainty associated with our two measurements (Meas and Cal). This uncertainty will make it difficult for us to determine what the actual RL value of our DUT is. We need to qualify our system to determine if it is accurate enough for our requirements.

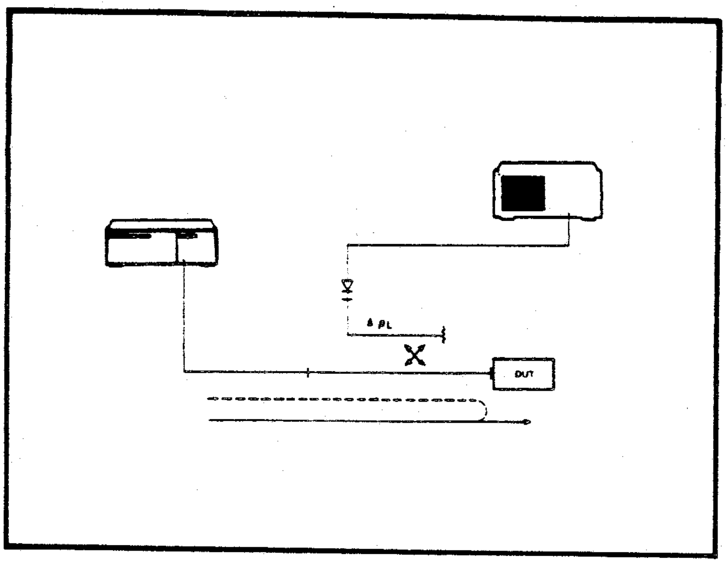
REFLECTION UNCERTAINTY EQUATION

$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

A = Directivity
B = Calibration Error, Frequency Response, Display and Instrument Errors
C = Effective Source Match
ρ_L = Reflection Coefficient of DUT

This equation is a simplification of a complex flowgraph analysis of reflectometer uncertainties. $\Delta\rho$ is the worst case uncertainty in the measurement where ρ_L is the measured reflection coefficient of the DUT. A, B, and C are all in linear terms. Each term in this equation will be analyzed separately.

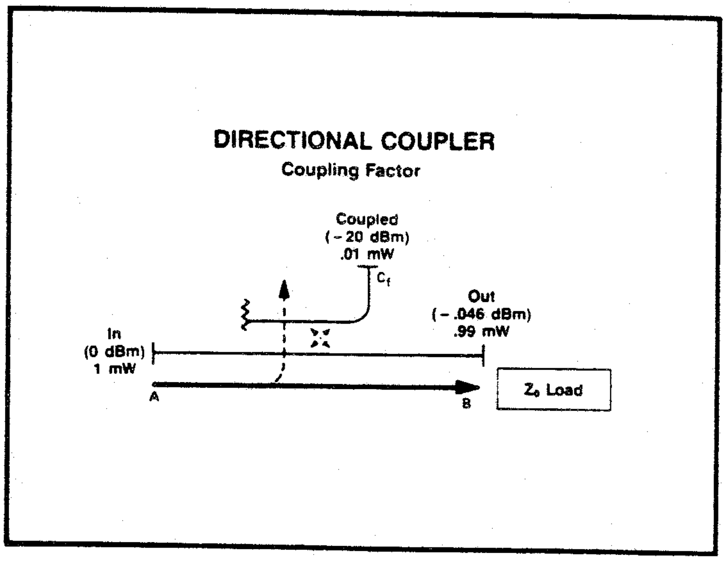
Let's take a closer look at our basic measurement system to see what causes error signals to exist and how those signals add to the uncertainty of our measurement. The component we will start with is our signal seperation device - the directional coupler.



Recall that a directional coupler couples a portion of the signal flowing through the main arm to the auxiliary arm. We've defined the coupling factor (dB) to be:

Coup Fact (dB) = $-10 \log [P_{cf}/P_{in}]$.

In defining coupling factor, we have assumed the coupler is terminated with a perfect load and thus no other signal is present in the auxiliary arm.

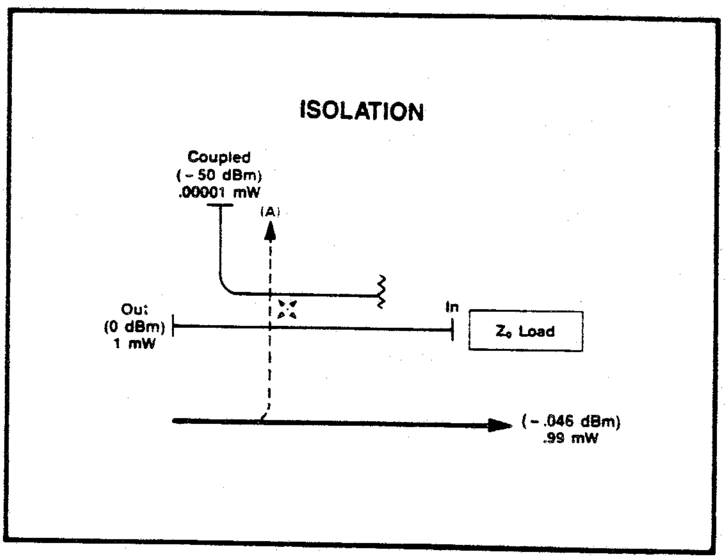


DIRECTIONAL COUPLER

Coupling Factor

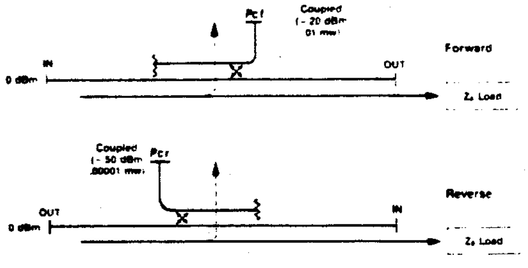
If we turn the coupler around and flow power in the reverse direction through the coupler, we ideally would measure no power in the auxiliary arm. However, some energy does leak across the coupler (sneaks in the back door). A measure of this leakage signal is defined as the isolation of the coupler:

Isolation (dB) = $-10 \log [P_{cr}/P_{in}]$.



ISOLATION

DIRECTIVITY



$$\text{Directivity} = 10 \log \frac{P\text{ coupled forward}}{P\text{ coupled reverse}}$$

The ablility to separate signals flowing in opposite directions within the coupler is directivity. We measure that ability by relating the power measured in the auxiliary arm from the coupler in the forward direction (P_{cf}) to the power measured in the auxiliary arm with the coupler in the reverse direction (P_{cr}). When measuring P_{cf} and P_{cr}, notice that we have the coupler terminated in a Zo load and that we have the same input power level.

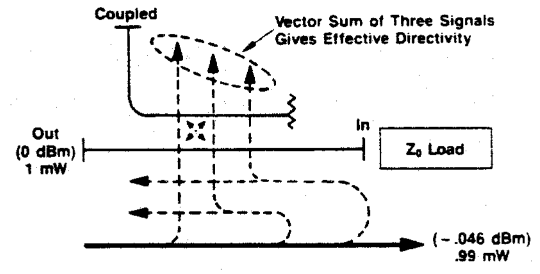
Directivity (dB)= 10 log [P_{cf}/P_{cr}].

Equivalent expressions for directivity are:

Dir (dB)= 10 log [Coup Fact/Iso]

Dir (dB)= Iso (dB) -Coup Fact (dB)

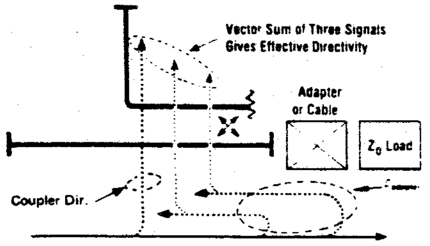
DIRECTIVITY



The sources of imperfect directivity are 1) leakage, 2) internal coupler load reflections, and 3) connector reflections. Coupler directivity is the sum of all three signals.

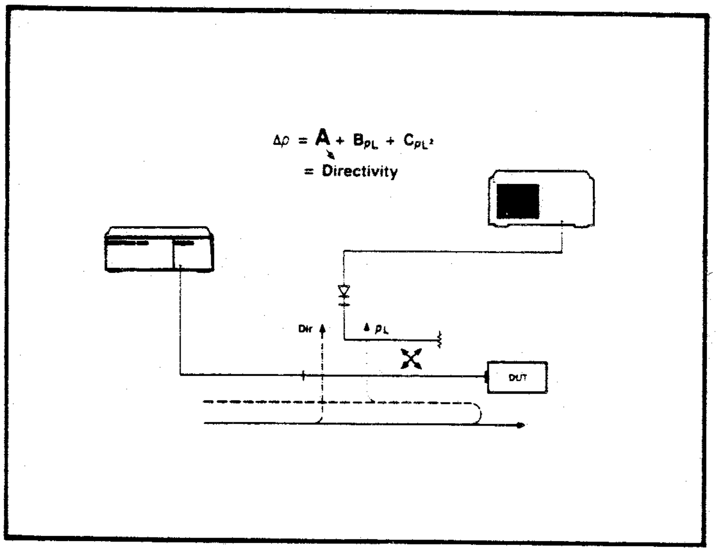
USING ADAPTERS OR CABLES ON DIRECTIONAL DEVICES

Measurement Directivity Degraded



The effects of adapters on effective directivity are often misunderstood. As the flow-graph shows, the adapter has the same relationship to directivity as the output connector on the coupler. If the adapter has a SWR of say 1.5:1 (the $2.00 variety), the effective directivity of the coupler drops to no better than 14 dB, even if the coupler has infinite directivity. In other words, with a perfect Zo load on the output of the adapter, the reflected signal appearing at the coupled arm would be 14 dB less than the reflection from a short circuit.

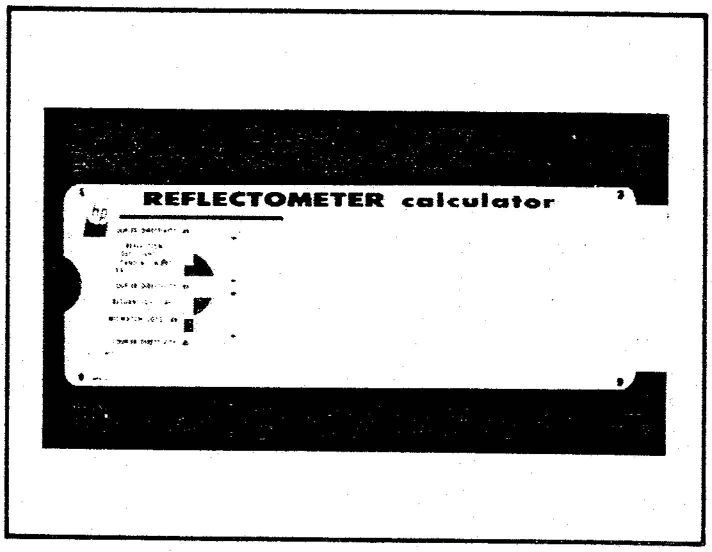
The effects of directivity on our measurement is shown here. The **A** term in our uncertainty equation is directivity. It is independent of the reflection coefficient of the DUT and adds (worst case) directly to the total uncertainty.
If the signal reflected from the DUT is large, for example for a short circuit, then the directivity will be small compared to the reflected signal and the effect of directivity will be insignificant.
If the signal reflected from the DUT is small (high return loss), then the directivity signal will be significant compared to the reflected signal and the uncertainty due to directivity is significant.


$\Delta\rho = \mathbf{A} + B_{\rho_L} + C_{\rho_L^2}$

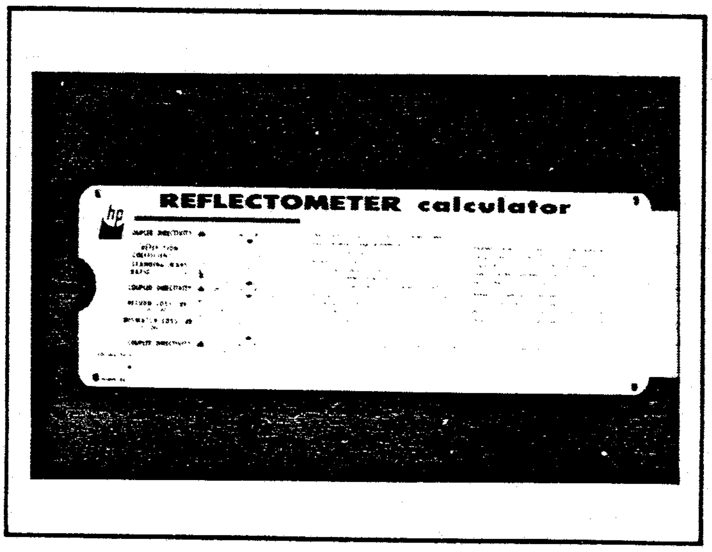
= Directivity

We can use the reflectometer calculator to convert directivity in dB to a linear term which is need in our uncertainty equation. For example, 40 dB directivity converts to A = .01.

On the reflectometer calculator, place the dB value for directivity under the blue arrow on the RL scale. Read the linear value below the blue arrow on the reflection coefficient scale.

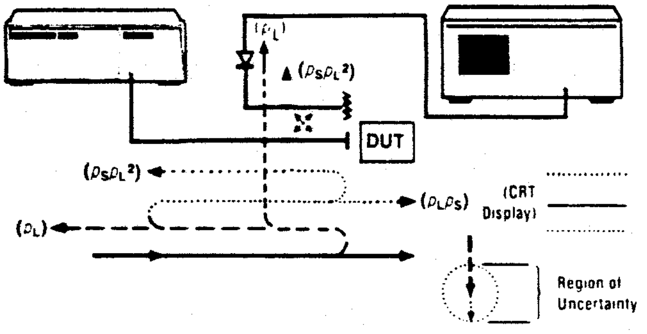


We can use the reflectometer calculator to see the effect of directivity on the uncertainty of a measurement. For example, set ρ to 0.05. Now read the error limits using the coupler directivity scale and the directivity of the coupler used in the measurement. A coupler with 40 dB directivity causes a ± 0.01 error in the measurement of $\rho = +0.05$. On the return loss scale this error is ± 2.5 dB. This is significant measurement error.
The coupler directivity error window in the refectometer calculator uses the sum of two error terms, A & C (directivity and source match terms) and assumes that A = C. A discussion of the C term will follow directly. When ρ_L is small, ρ_L^2 is even smaller, and therefore the effect of the C term is negligible, and directivity is the dominant error. If ρ_L is large, i.e. near 1.0, then the C term has a significant error contribution.



$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

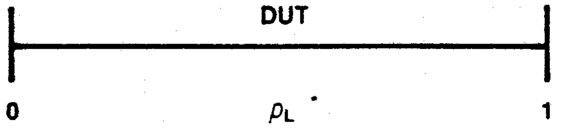
Effective Source Match of Measuring System



What happens to our measurement uncertainty when our source is not perfect (i.e. do not have perfect source impedance).
From following the flowgraph, we can see that effective source match belongs in the C term. The first reflection from the DUT is ρ_L ; this is what we want to measure. This signal continues back toward the source where it is re-reflected if the reflection coefficient of the source is not perfect, resulting in a signal normalized to $\rho_L * \rho_S$ flowing back toward the DUT where it is again reflected and sampled as $\rho_S * \rho_L^2$.

$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

Effective Source Match

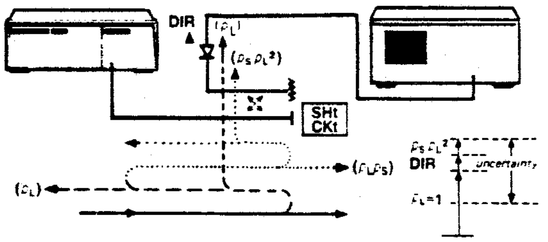


When the reflected signal is large (low return loss) source match is the major error term. For example, when measuring a 3 dB return loss device (ρ_L = 0.71) with a source match of 2:1 SWR (ρ_S =0.33) the uncertainty due to source match alone is 0.17 worst case. But if we measured a DUT with a return loss of 30 dB (ρ_L = 0.03) then the uncertainty of our measurement is 0.0003.

The last contributor to our measurement uncertainty that we will discuss is calibration error. We calibrate with a short circuit because we know that the reflection coefficient of the short has a value of 1 (Return Loss = 0 dB).

$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

Calibration Error Term



But instead of measuring just the reflection coefficient of the short, we also measure some error terms. Directivity is always present and will be measured. Source match is also present and will be measured along with our standard. The sum of directivity and source match (A + C) will cause uncertainty in the measurement of our standard. If we assume no other errors are present then our best case calibration error (B term) is equal to the sum of directivity and source match.

$$B = A + C$$

Lets take a look at a simple example:

RL of DUT = 6 dB (.5)
Directivity = 30 dB (.0316)
Source Match = 1.9:1 (0.31)
B= A+C = (.3416)

$$=A + B\rho_L + C\rho_L^2$$

$$= .0316 + .3416(.5) + .31(.5)^2$$
$$= \pm 0.28$$

SIMPLE REFLECTOMETER ACCURACY

$$\Delta\rho = A + B\rho_L + C\rho_L^2$$

Example:

Directivity	30 dB	(.0316)
Source SWR	1.9:1	(0.31)
A = .0316	Load R.L. = 6 dB (.5)	

B = .3416
C = 0.31
ρ_L = 0.5

$$\Delta\rho = .0316 + .3416(.5) + .31(.5)^2 = \pm 0.28$$

From our previous example, it should be obvious that we would want much less uncertainty in our measurement if we could get it. We can improve our reflection measurements (i.e., reduce uncertainty) by either removing our calibration error or by improving source match or both.

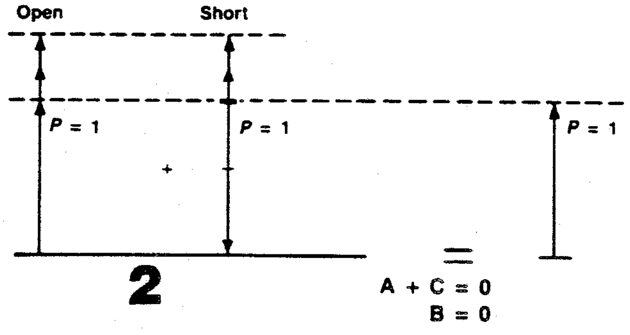
IMPROVING REFLECTION MEASUREMENTS

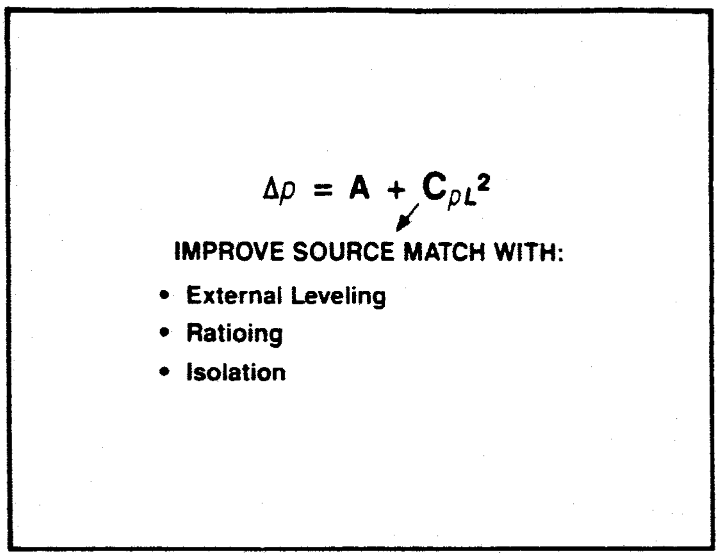
- Remove Calibration Error
- Improve Source Match

The calibration error due to the sum of the directivity and source match errors can be removed by averaging the short and open circuit responses. Though the reflection from an open circuit is 180 degrees out of phase with that from a short circuit, the errors due to the sum of directivity and source match do not change phase when the load is changed from an open to a short.

The open/short average then averages out calibration error thus making B=0.

OPEN/SHORT AVERAGING REMOVES CALIBRATION ERROR



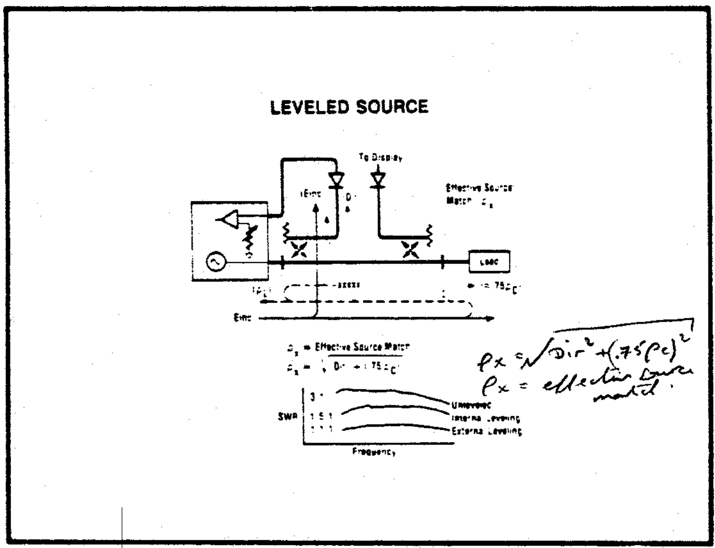


$$\Delta\rho = A + C\rho_L^2$$

IMPROVE SOURCE MATCH WITH:

- External Leveling
- Ratioing
- Isolation

The other method of reducing uncertainty is by improving source match.



LEVELED SOURCE

ρ_x = Effective Source Match

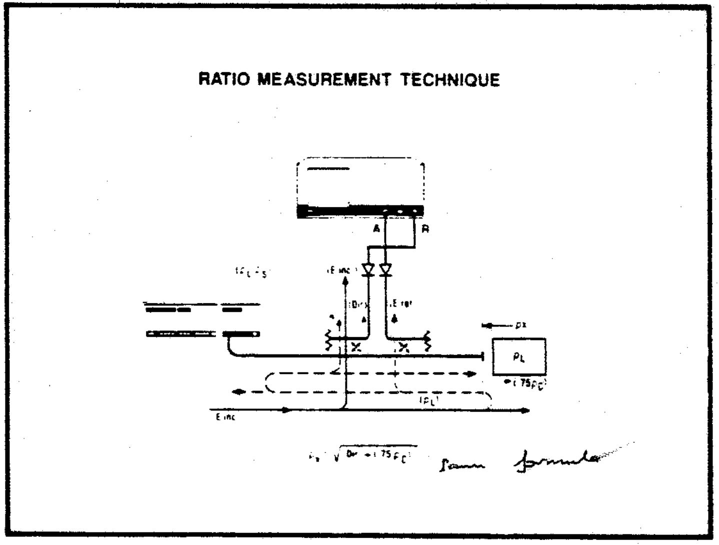
$\rho_x = \sqrt{Dir^2 + (.75\rho_C)^2}$

$\rho_x = \sqrt{Dir^2 + (.75\rho_c)^2}$

ρ_x = effective source match.

The perfect source would deliver a constant power to a load regardless of the reflections from the load. Leveling the sweeper output improves source match by varying the power out of the source to compensate for the power reflected by the source, thus the power delivered to the load is constant (it appears the source has absorbed the power flowing into it). Any signal re-reflected from the sweeper is sensed by the leveling loop which corrects the output from the sweeper accordingly.

Although leveling improves source match, there are still inherent uncertainties. First, the output connector of the coupler has some reflection. In addition, directivity error also enters in. ρ_x is defined as effective source match . If a non HP coupler is used, then effective source match must be calculated from the equation shown. The equation for effective source match very closely approximates the flowgraph analysis of a leveled source using a directional coupler.

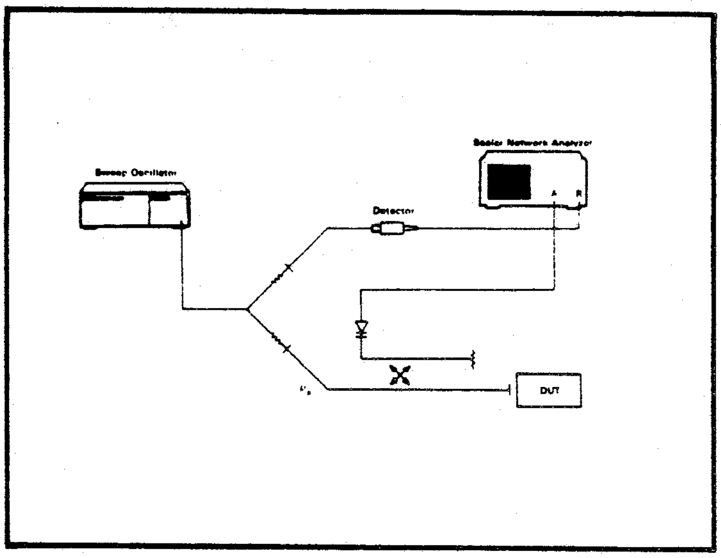


RATIO MEASUREMENT TECHNIQUE

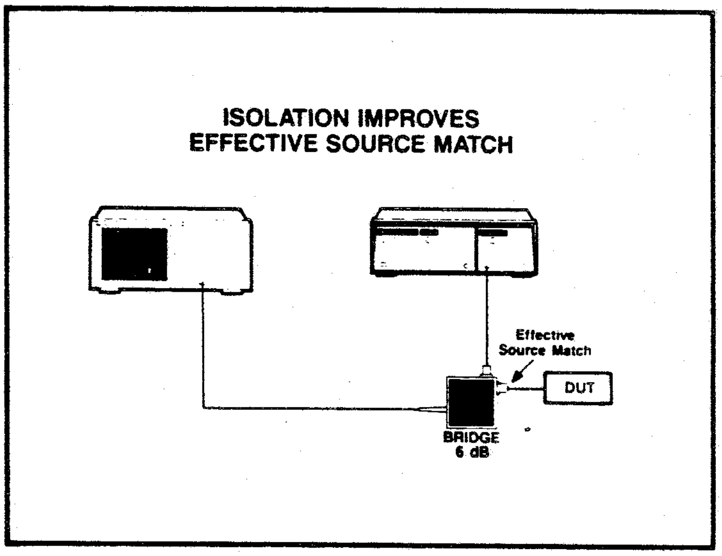
$\rho_x = \sqrt{Dir^2 + (.75\rho_C)^2}$ same formula

Effective source match can also be improved by ratioing the incident and reflected signals. With this technique, the absolute incident power is not controlled as in leveling but the variations are ratioed out. Any re-reflections are seen by both detectors and when you take the ratio A/R on the analyzer, the effect of ρ_S is cancelled. Again since the coupler is not perfect, the "effective source match" must be calculated. Effective source match is determined by the same equation as for leveling.

Of course, a power splitter may be used for both of these examples rather than a directional coupler. A splitter may be a better choice due to its smaller size and broadband response but the equation for effective source match does not apply. Since we essentially "buy" source match, the effective source match is the specification on the data sheet of the splitter.



The insertion loss of a directional bridge isolates the DUT from the source and improves the effective source match by attenuating the reflected and re-reflected signals each time they pass through the bridge. Assuming 12 dB isolation (6 dB each way) and bridge test port match of 1.25 SWR, the effective source match is improved from 1.9 to 1.46 SWR. A 6 dB attenuator would serve the same purpose.



ISOLATION IMPROVES
EFFECTIVE SOURCE MATCH

Let's see how much we reduce the uncertainty in the example measurement using open/short averaging and ratioing with a coupler.

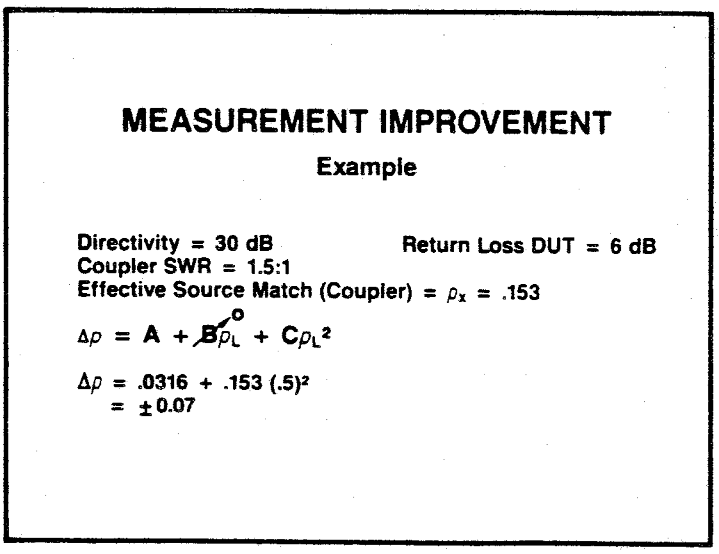
Effective source match = ρ_x =0.153

$$= A + B\rho_L + C\rho_L^2$$

$$= .0316 + .153(.5)^2 \qquad B=0$$

$$= \pm .07$$

whereas before we had ± 0.28.



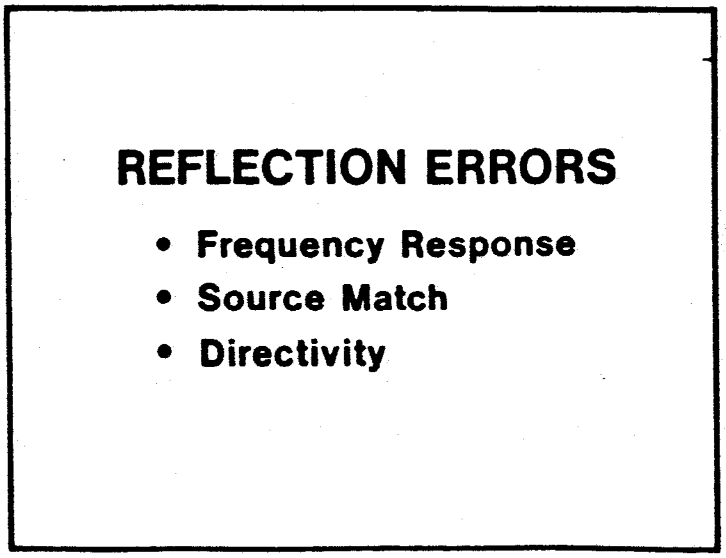
MEASUREMENT IMPROVEMENT

Example

Directivity = 30 dB Return Loss DUT = 6 dB
Coupler SWR = 1.5:1
Effective Source Match (Coupler) = ρ_x = .153

$$\Delta\rho = A + \overset{0}{\cancel{B}}\rho_L + C\rho_L^2$$

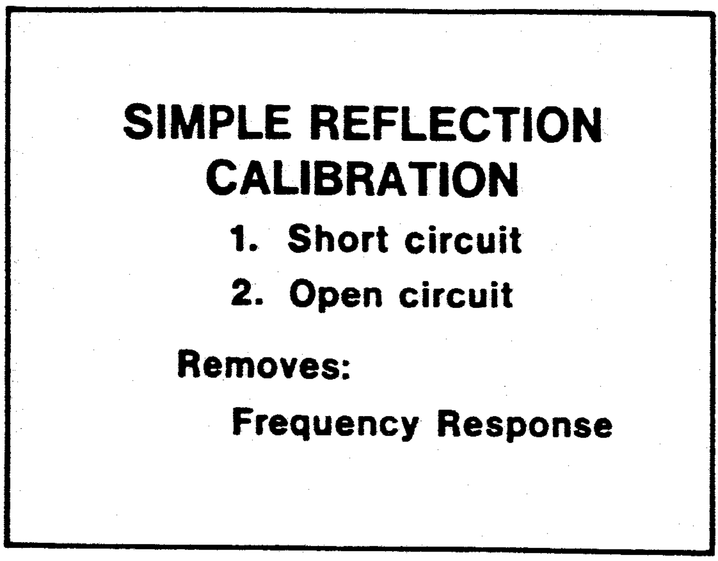
$$\Delta\rho = .0316 + .153\,(.5)^2$$
$$= \pm 0.07$$



REFLECTION ERRORS

- Frequency Response
- Source Match
- Directivity

Since Reflection is a more difficult measurement, especially in calibration, let's try it first. These three errors are common in reflection measurements due to system imperfections. Couplers, connectors, cables, etc. all are contributors. High reflection measurements are badly distorted by Source Match error, low reflection measurements by Directivity effects.



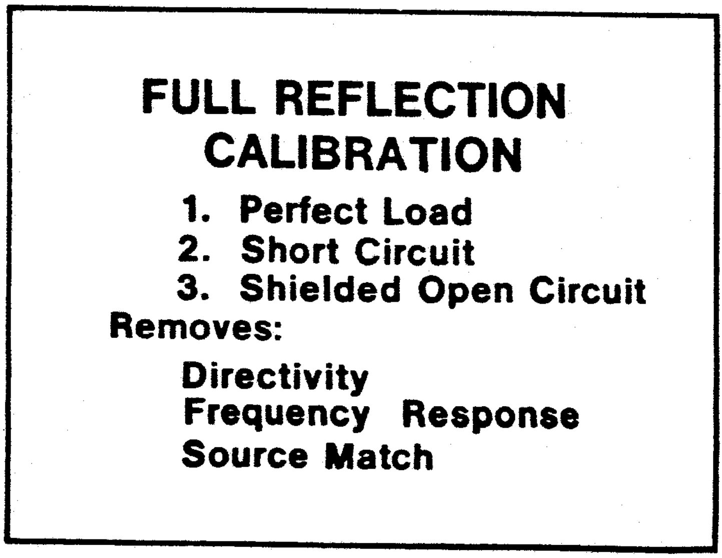
SIMPLE REFLECTION CALIBRATION

1. Short circuit
2. Open circuit

Removes:

Frequency Response

A simple calibration might involve measuring just a short circuit to remove frequency response as is done on most manual systems. Automatic systems can improve this by measuring a short and an open circuit -- the open/short average for a more accurate calibration.



FULL REFLECTION CALIBRATION

1. Perfect Load
2. Short Circuit
3. Shielded Open Circuit

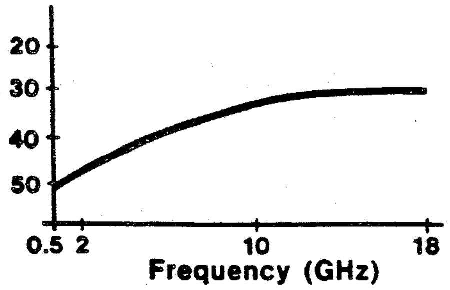
Removes:

Directivity
Frequency Response
Source Match

A more extensive vector calibration involves three calibration standards: a perfect load, a short, and an open circuit. The perfect load has no reflection, hence measuring it provides the directivity error (a residual signal). After knowing directivity, measuring the short and open provide the Source Match and Frequency Response. The vector capability allows removing these errors completely whereas a scalar system could only approximate their affects.

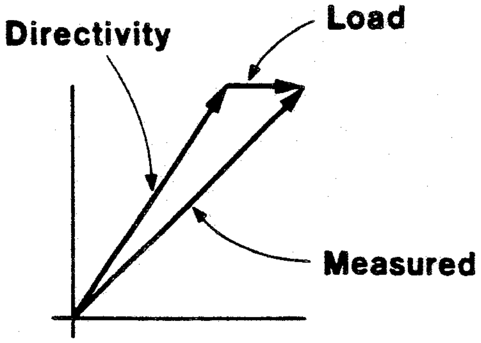
Here is a typical response of an APC-7 fixed termination. Since the load sets the no reflection signal level, we are assuming that the measured signal is the system directivity.

TYPICAL FIXED LOAD RETURN LOSS (dB)

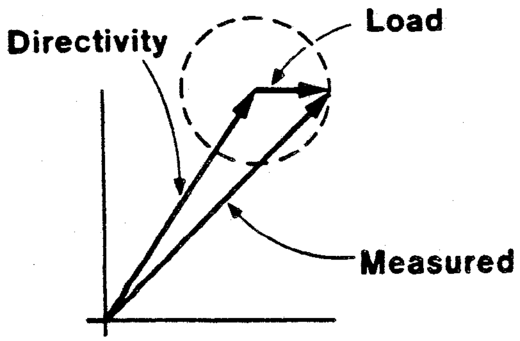


With a load we actually measure both the load response and the system directivity. If the load reflection is small or non-existent, then we measure just the directivity. Since directivities of 26 dB or higher are common, a 30 dB Return Loss load is no longer a small reflection.

WHAT'S MEASURED?

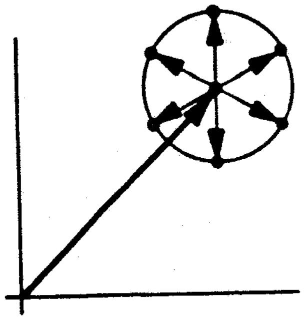


SLIDING LOAD RESPONSE



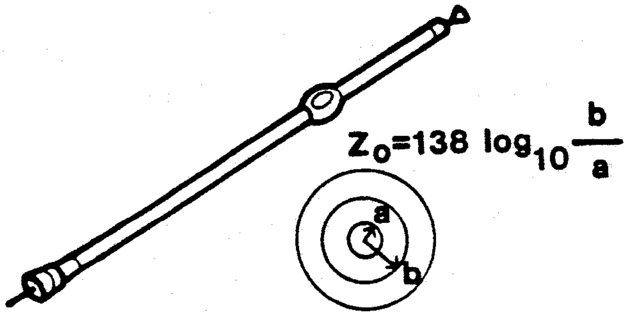
A sliding load is a termination placed inside an air line. This termination can be moved in position inside this air line and thus vary its distance from the point at which we measure. Since the load reflection magnitude is constant, this movement translates to a phase shift and a circle can be generated.

DETERMINE CENTER OF CIRCLE



We slide the load 6 times in an ANA to generate a complete circle at each test frequency. The 6 positions allow for over-determination of the error circle. This compensates for overlapping of some points. By mathematically determining the center of the circle, we then know the directivity.

SLIDING LOAD DESIGN



$$Z_0 = 138 \log_{10} \frac{b}{a}$$

Since the sliding load response need not be perfect to generate a circle, better than 40 dB directivity can be measured. How? The beadless air line typically has better than 50 dB Return Loss in APC-7 because of the geometry of the conductors.

MEASURE EFFECTIVE DIRECTIVITY

Before Cal:

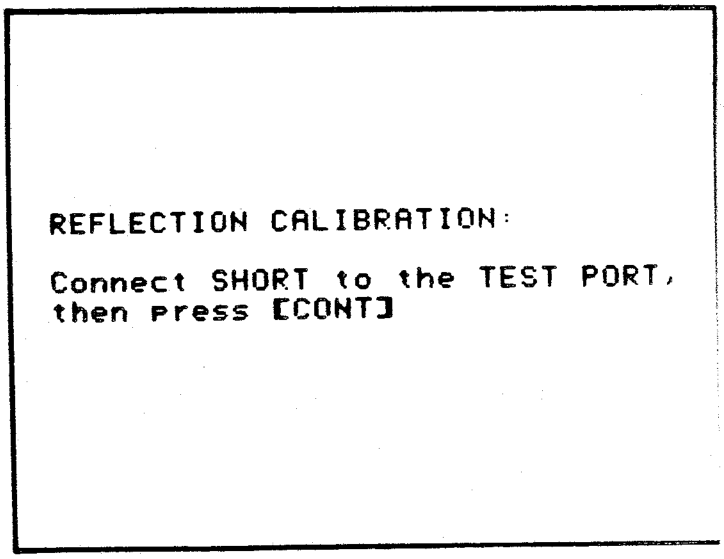
Test Set	**<26 dB**
Connectors	**<30 dB**

After Cal:

Effective Dir. >40 dB

Directivity is determined by several imperfections. The total system directivity is usually in the 20 to 30 dB range. After calibration our effective system directivity will be determined by the residual error of the load used.

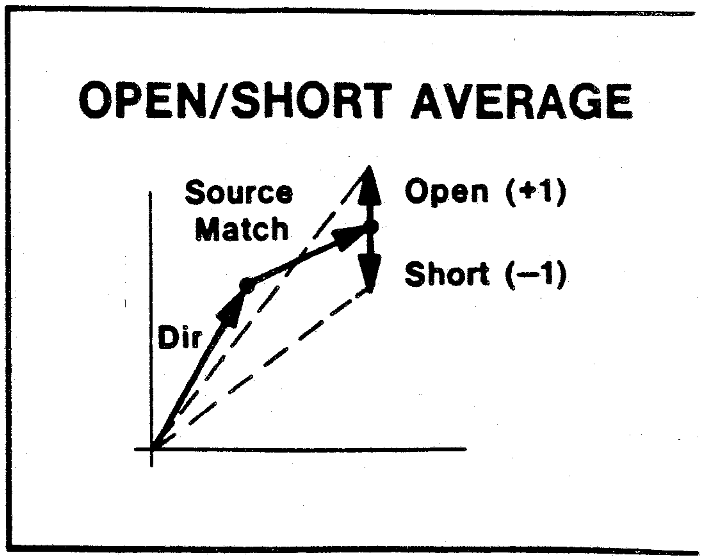
Now for the other errors. The short circuit has a Return Loss of 0 dB and 180 degrees. We measure it first, then the shielded open circuit.



REFLECTION CALIBRATION:

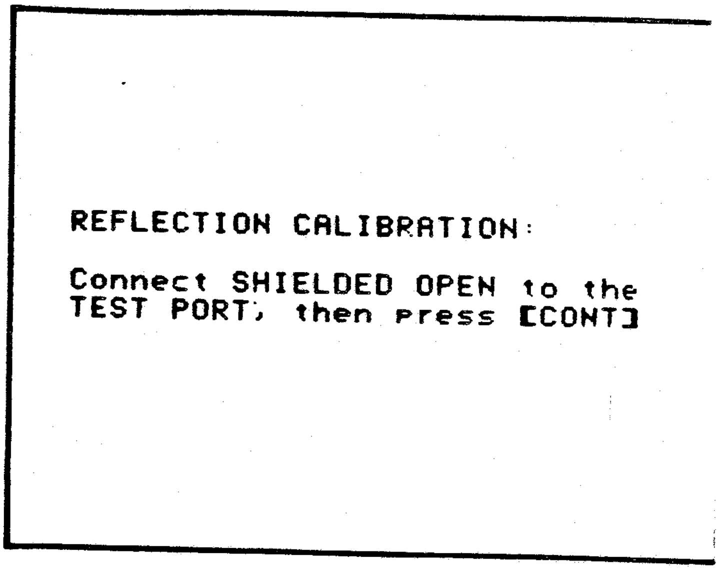
Connect SHORT to the TEST PORT, then press [CONT]

The open and short are used to measure Source Match and Frequency Response. The open and short both are a 100% reflection, but differ in phase shift by 180 degrees. Measuring both yields a maximum and minimum variation with the Source Match and Directivity errors. Removing the Directivity leaves Source Match.



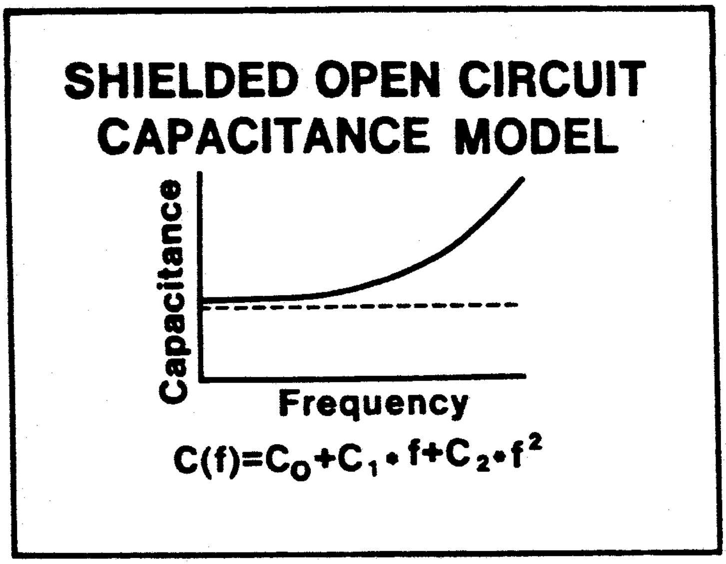
OPEN/SHORT AVERAGE

Now we measure the shielded open circuit. The shield is a ground plane extension allowing us to more accurately model the open circuit capacitance.



REFLECTION CALIBRATION:

Connect SHIELDED OPEN to the TEST PORT, then press [CONT]



SHIELDED OPEN CIRCUIT CAPACITANCE MODEL

$C(f)=C_0+C_1 \cdot f+C_2 \cdot f^2$

The open circuit capacitance is best modeled as the sum of three terms. The frequency squared term results from the fact that an open looks like waveguide beyond cutoff. These three coefficients are empirically determined for the connectors and calibration standards used with the ANA.

WHY FULL ERROR CORRECTION?

Devices with:

- **Low Reflection**
- **High Reflection**

Why worry about all of these errors? A Directivity of 20 to 26 dB limits a 26 dB device measurement to a 100% error. A typical Source Match of 1.5 SWR (14 dB Return Loss) will cause a ± 1.8 dB error in measuring a short.

THE 12-TERM ERROR MODEL

REFLECTION ERRORS:

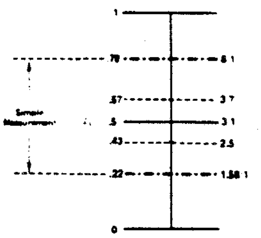
- Directivity
- Source Match
- Frequency Response

TRANSMISSION ERRORS:

- Frequency Response
- Crosstalk
- Load Match

Higher accuracy also comes from a better characterization of the measurement system. A more extensive calibration model like the 12-Term Vector Error Model improves transmission measurements by calibrating for more mismatch and isolation errors as well as improving reflection measurements of two port devices. The 12 terms include 6 each for forward and reverse calibrations.

MEASUREMENT IMPROVEMENT



The improvement in reflection coefficient uncertainty translates to an improvement in SWR

IMPROVING REFLECTION MEASUREMENTS

1. Leveling, Ratioing, Isolation

$\Delta\rho = A + B\rho_L + \mathbf{C}\rho_L^2$

2. Open/Short Averaging

$\Delta\rho = A + \mathbf{B}\rho_L + C\rho_L^2$

We have discussed two ways of improving the accuracy of reflection measurements - both are simple and inexpensive. The source match improvement techniques reduce the C term of the uncertainty equation and open/short averaging (which is included in the HP 8756/8757 firmware) removes the B term from the same equation.

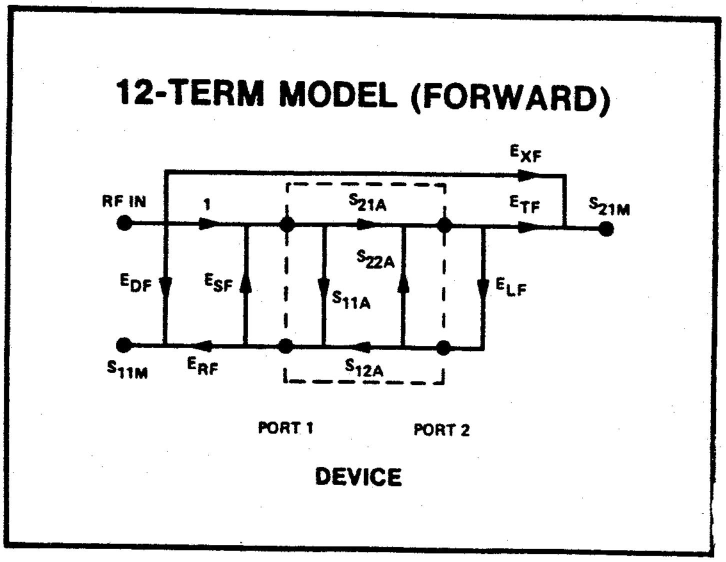
The high directivity (40 dB) bridges in HP's product line reduce the A term for accurate measurements of low reflection DUT's.

EXAMPLE

$$S_{21A} = \frac{\left[1 + \left(\frac{S_{22M} - E_{DR}}{E_{RR}}\right)\left(E_{SR} - E_{LF}\right)\right]\left(\frac{S_{21M} - E_{XF}}{E_{TF}}\right)}{\left[1 + \left(\frac{S_{11M} - E_{DF}}{E_{RF}}\right)E_{SF}\right]\left[1 + \left(\frac{S_{22M} - E_{DR}}{E_{RR}}\right)E_{SR}\right] - \left[\left(\frac{S_{21M} - E_{XF}}{E_{TF}}\right)\left(\frac{S_{12M} - E_{XR}}{E_{TR}}\right)E_{LF}E_{LR}\right]}$$

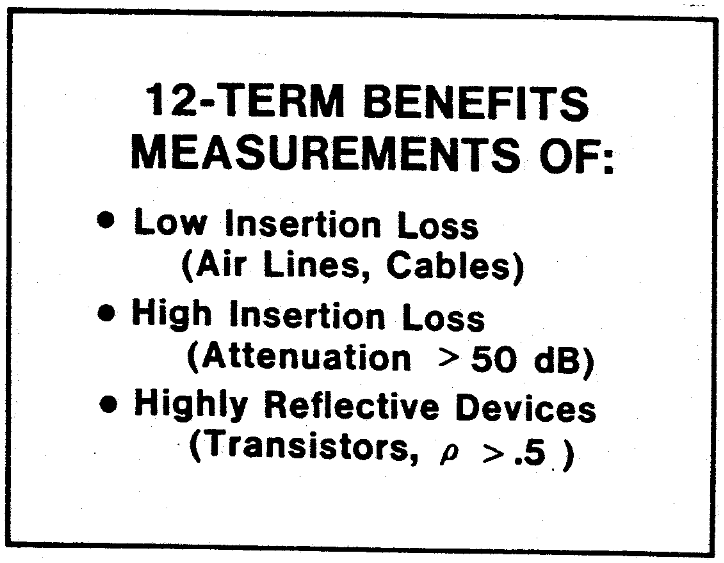
This is the error-correction equation for one of the S-Parameters. We can see that the other 3 measured S-Parameters are also included in it.

This is the 12 term flow graph. Note that the reflection part of the model is the same as before for a single port device.



12-TERM MODEL (FORWARD)

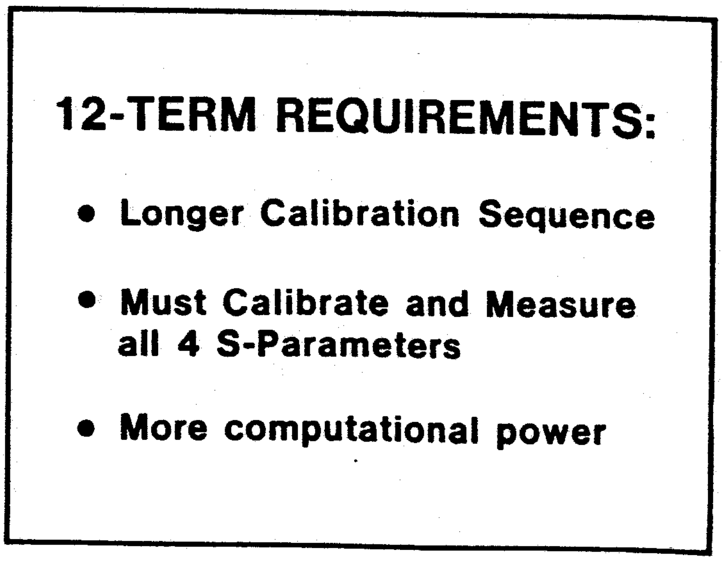
Why use the 12-term model? It enhances measurements of two port devices, especially those with low or high insertion losses. Also devices that are highly reflective are vulnerable to large mismatch errors.



12-TERM BENEFITS MEASUREMENTS OF:

- Low Insertion Loss
  (Air Lines, Cables)
- High Insertion Loss
  (Attenuation > 50 dB)
- Highly Reflective Devices
  (Transistors, $\rho > .5$)

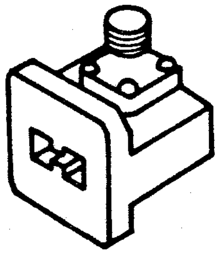
Unfortunately the 12-term model requires a more extensive calibration and measurement cycle. To measure a single parameter requires that we calibrate and measure all 4 S-Parameters! This longer time and computational power will call for a more powerful computer.



12-TERM REQUIREMENTS:

- Longer Calibration Sequence
- Must Calibrate and Measure all 4 S-Parameters
- More computational power

WHAT IF NON-STANDARD CROSS-SECTION? RIDGED?

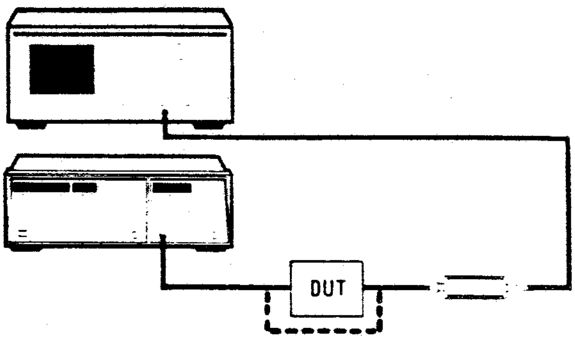


1. Need Calibration Standards
2. Know Calibration Parameters

For non-standard or unusual waveguide, the ANA can also measure them as long as the proper calibration standards and performance data are available.

"THROUGH" SUBSTITUTION

Normalizes Transmission Measurement



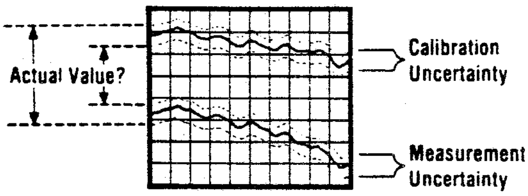
"Through" Substitution

Before we can determine transmission loss or gain we must establish a reference (i.e. we need to know what the incident power is). By measuring a "through", we establish a 0 dB reference trace on the analyzer display.

By subtracting our "thru" reference from the measurement obtained with a DUT (**normalization**) we can determine what the insertion loss or gain of the DUT is. Normalization also removes the frequency response of the test setup.

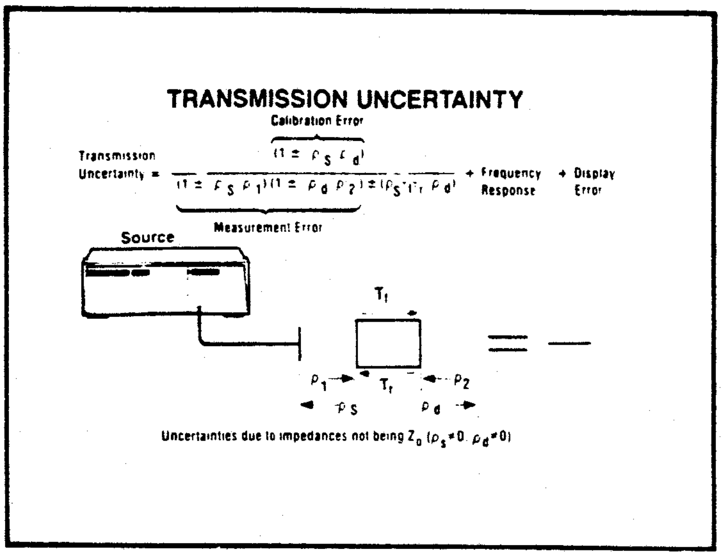
TRANSMISSION MEASUREMENT UNCERTAINTIES

WITH SOURCE ≠ Z_0
&/or DETECTOR ≠ Z_0



As with reflection measurements, uncertainties exist with transmission measurements. Total measurement uncertainty is affected by calibration uncertainty as well as the measurement uncertainty.

The transmission uncertainty equation quantifies the worst case error window caused by the system. Source and detector mismatch cause an uncertainty around both the calibration and the measurement traces. Frequency response errors are elimated through normalization. Since it is a tedious task to evaluate this linear equation, let's explore an easier and more understandable quantification.



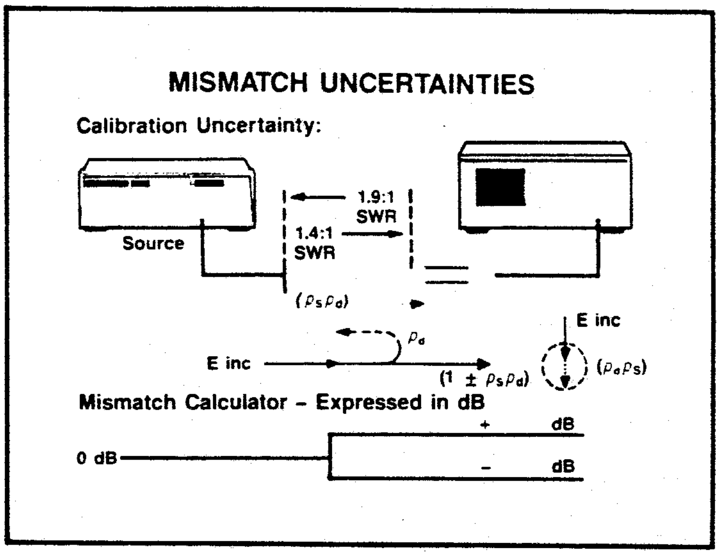
TRANSMISSION UNCERTAINTY

$$\text{Transmission Uncertainty} = \frac{\overbrace{(1 \pm \rho_S \rho_d)}^{\text{Calibration Error}}}{\underbrace{(1 \pm \rho_S \rho_1)(1 \pm \rho_d \rho_2)}_{\text{Measurement Error}} \pm (\rho_S T_f T_r \rho_d)} + \text{Frequency Response} + \text{Display Error}$$

Uncertainties due to impedances not being Z_0 ($\rho_s \neq 0$, $\rho_d \neq 0$)

Let's quantify the transmission uncertainties. Please get out your reflectometer calculator.

Let's first investigate the uncertainties associated with the calibration stage of the measurement. As shown, when the detector is connected to the source, the incident signal first encounters the detector impedance where part of the incident is reflected (normalized to ρ_d). This reflected signal is then re-reflected by the source mismatch resulting in an uncertainty vector of $\rho_s * \rho_d$ at some unknown phase relationship to the incident signal. Worst case, the extremes of the signal seen by the detector would be $1 \pm \rho_s * \rho_d$.

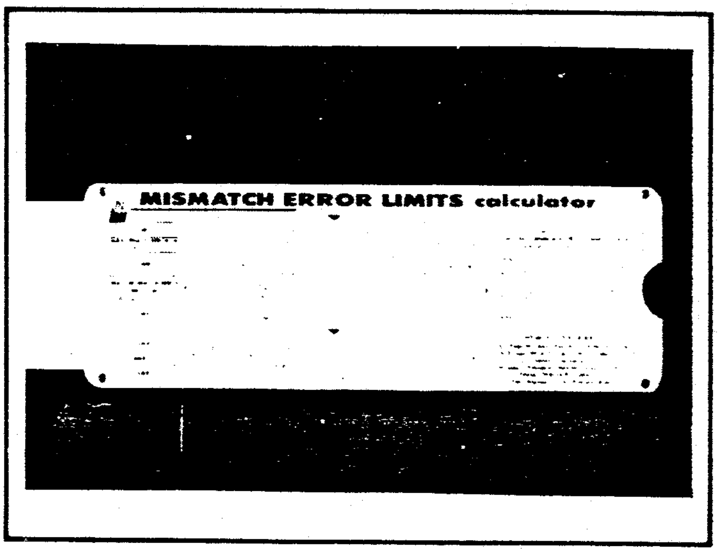


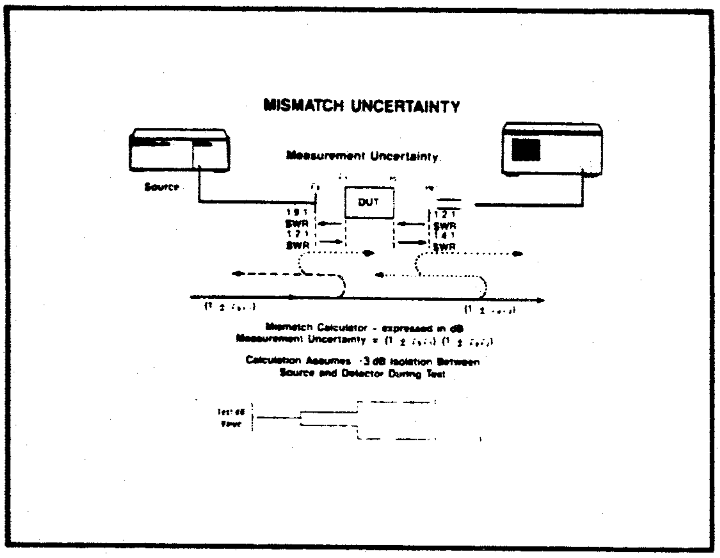
MISMATCH UNCERTAINTIES

Calibration Uncertainty:

Mismatch Calculator - Expressed in dB

The Mismatch Error Limits side of the refectometer calculator converts the two SWR's into the uncertainty limits in dB. Place the black arrow over the SWR of either the source (1.9 in this example) or the detector (1.4). Under the SWR of the other device read the (+) value of Max Mismatch Error (+.44 dB). Directly below this read the (-) value (-.46 dB).





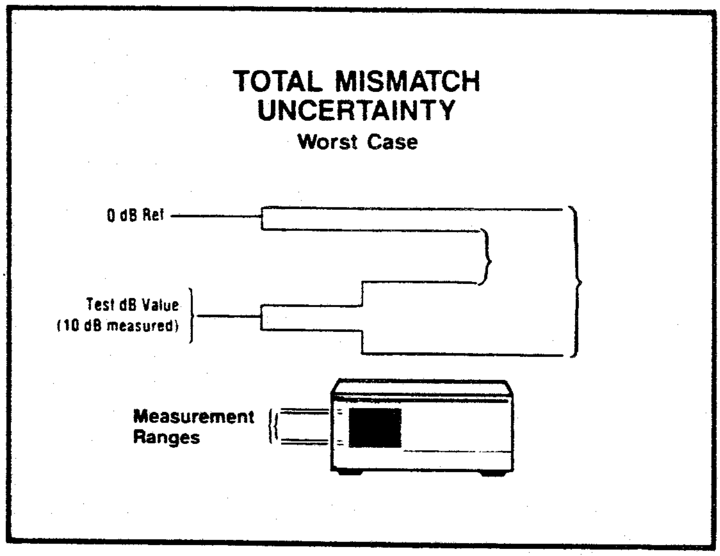
MISMATCH UNCERTAINTY

Mismatch Calculator - expressed in dB
Measurement Uncertainty = $(1 \pm \rho_s \rho_1)(1 \pm \rho_2 \rho_d)$

Calculation Assumes >3 dB Isolation Between Source and Detector During Test

Similarly, there are uncertainties in the measurement stage due to the source/DUT input mismatch $(1 \pm \rho_S * \rho_1)$ and the DUT output/detector mismatch $(1 \pm \rho_2 * \rho_d)$. Each of these uncertainties can be found with the Mismatch Error Limits Calculator just as before. They are then added to get the uncertainty window in the measurement stage. Assume that the DUT SWR is 1.2:1.

In this example we assume that the DUT has an input to output isolation of >3 dB so that multiple reflections have a negligible effect on the measurement uncertainty.



TOTAL MISMATCH UNCERTAINTY

Worst Case

As the diagram shows, when the calibration window and the measurement window are combined, the total worst case uncertainty for a particular measuring system and DUT can be determined. In this example we see that the worst case uncertainty in measuring a 10 dB attenuator is almost ± 1 dB.

IMPROVE TRANSMISSION MEASUREMENTS

- Improve Source Match
- Improve Detector Match

Obviously, as seen from the previous example, we need to improve our transmission measurement accuracy.

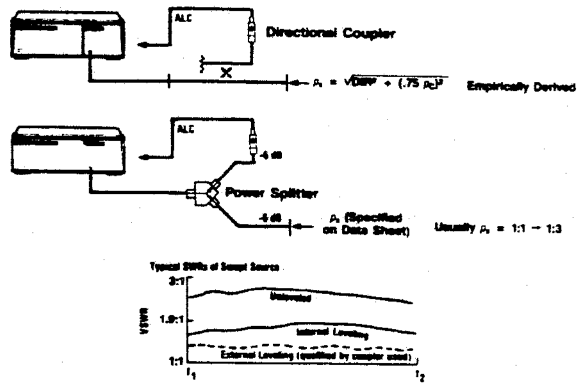
As with reflection measurements, we can reduce transmission uncertainty by improving source match.

IMPROVE SOURCE MATCH WITH:

- External Leveling
- Ratioing
- Isolation

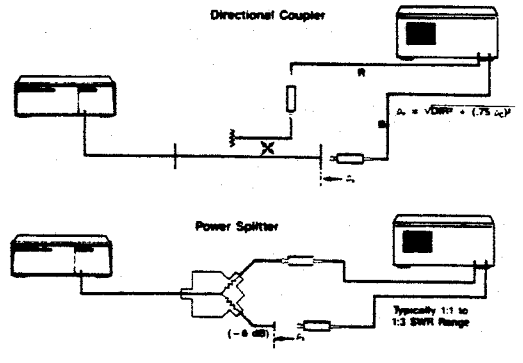
Leveling the source improves source match for transmission measurements. Source match improvement (when using a directional coupler) is similar to that gained in reflection measurements.

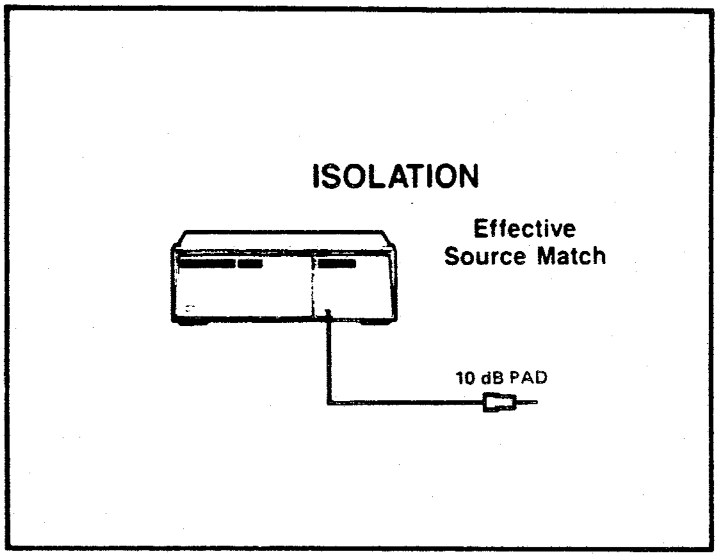
IMPROVING SOURCE REFLECTION COEFFICIENT USING LEVELING TECHNIQUES



The effects of power variations are removed with ratioing. Source match improvement similar to that gained previously in reflection measurements is obtained.

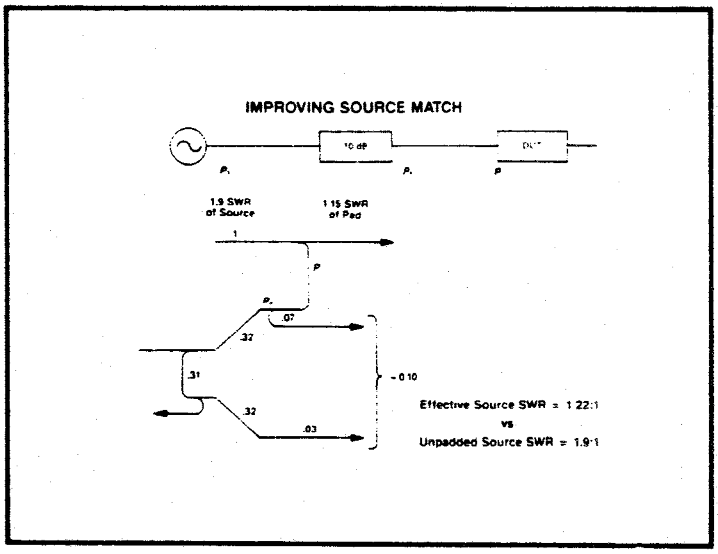
IMPROVING EQUIVALENT SOURCE REFLECTION COEFFICIENT USING RATIO TECHNIQUES





ISOLATION

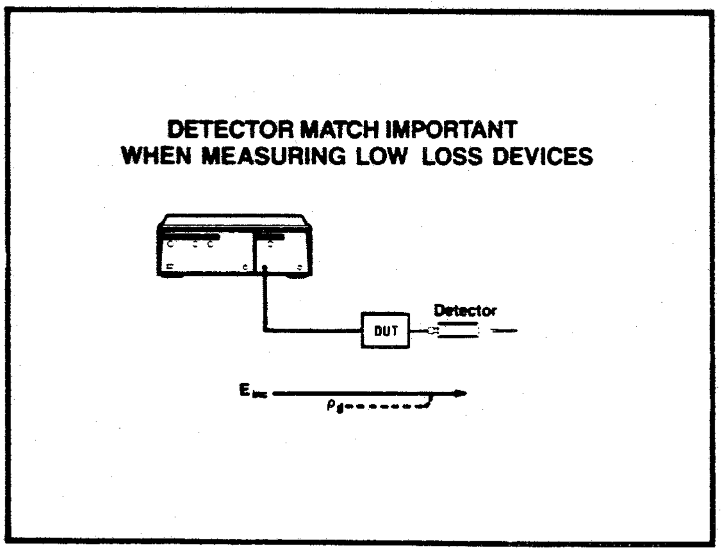
Isolation improves source match by attenuating the reflected signal each time the reflected signal flows through the attenuator.



IMPROVING SOURCE MATCH

Effective Source SWR = 1.22:1
vs
Unpadded Source SWR = 1.9:1

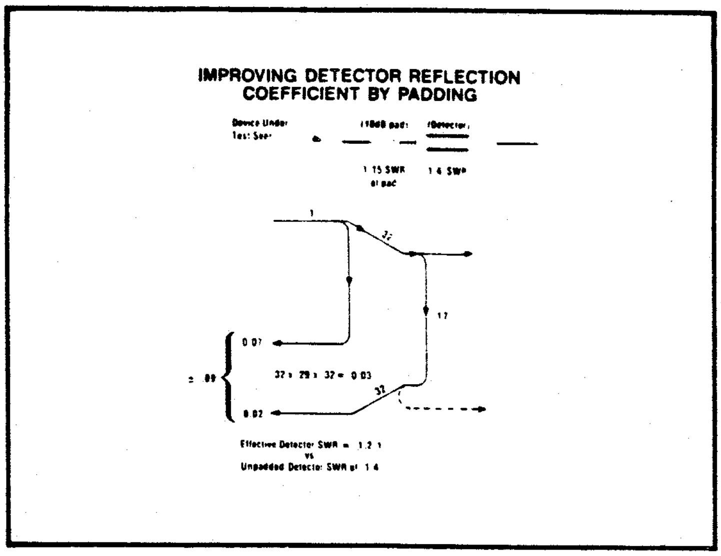
We can see by this flow diagram that the 10 dB attenuator improves effective source match considerably. The major drawback to this method is the loss of measurement dynamic range.



DETECTOR MATCH IMPORTANT WHEN MEASURING LOW LOSS DEVICES

Detector mismatch also contributes to transmission measurement error. If the DUT has low insertion loss (for example a transmission line), then the signal reflected from the detector and re-reflected from the source will cause a significant error.

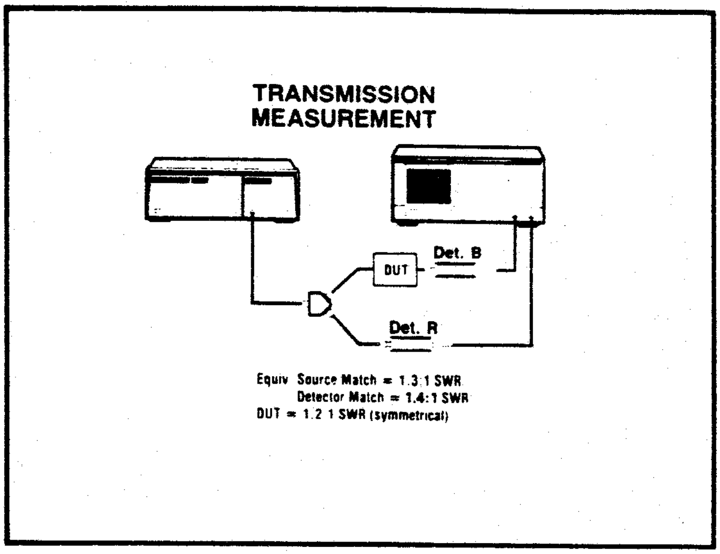
A way of reducing detector mismatch is to use an attenuator for isolation. In this example, a 10 dB pad is used at the input of the detector to improve the match to that of the pad. The major drawback of this technique is that dynamic range is decreased.



IMPROVING DETECTOR REFLECTION COEFFICIENT BY PADDING

Effective Detector SWR = 1.2:1
vs
Unpadded Detector SWR of 1.4

Here is an example on an improved transmission measurment system. Ratioing is used to improve the source match to 1.3:1 SWR.



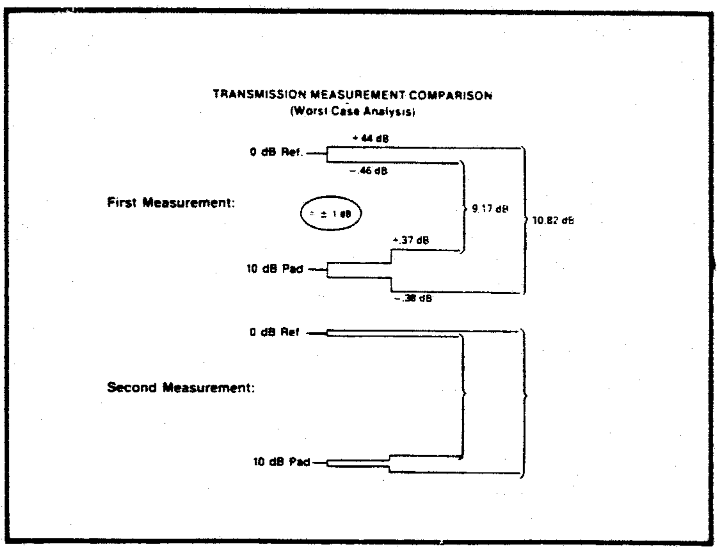
TRANSMISSION MEASUREMENT

Equiv Source Match = 1.3:1 SWR
Detector Match = 1.4:1 SWR
DUT = 1.2:1 SWR (symmetrical)

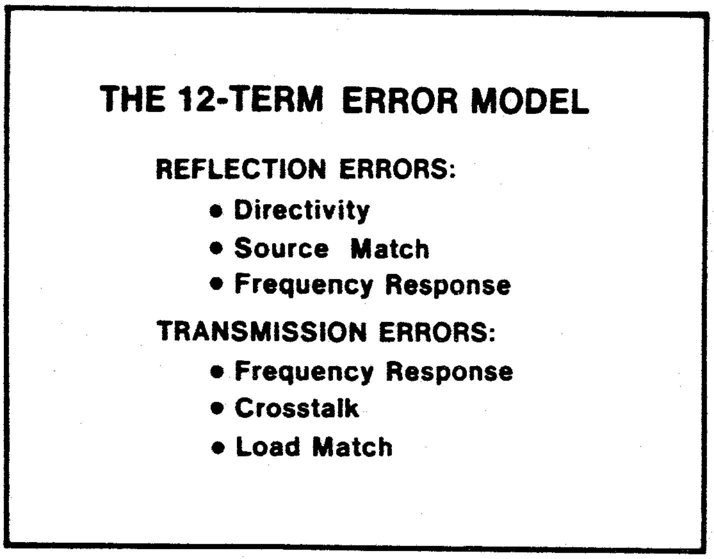
How much improvement do we get with ratioing? The top brackets show the uncertainty we obtained with the original simple measurement: approximately ± 1 dB.

Using the Mismatch Error Limits calculator to calulate the improved uncertainty, we see it is approximately ± .4 dB. Again this is worst case uncertainty.

Note that the uncertainty is independent of the measured value of transmission as long as it is >3 dB (for example, if we measued a 20 dB pad the uncertainty would be the same).



TRANSMISSION MEASUREMENT COMPARISON
(Worst Case Analysis)



THE 12-TERM ERROR MODEL

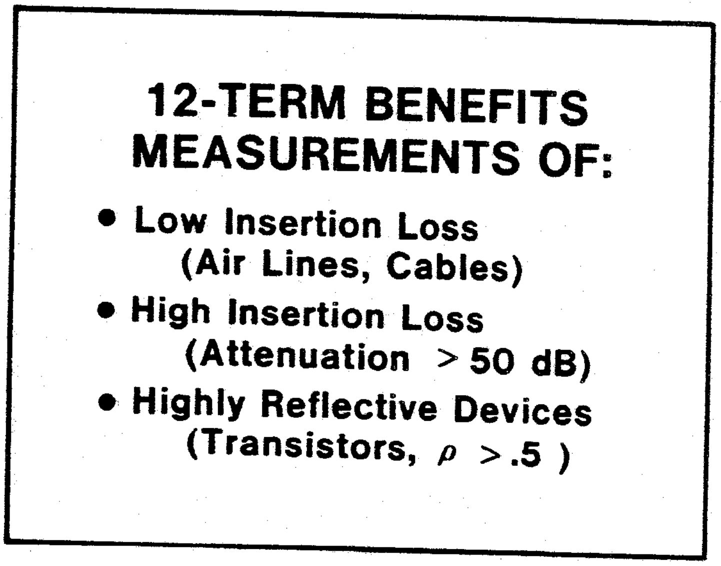
REFLECTION ERRORS:

- Directivity
- Source Match
- Frequency Response

TRANSMISSION ERRORS:

- Frequency Response
- Crosstalk
- Load Match

Higher accuracy also comes from a better characterization of the measurement system. A more extensive calibration model like the 12-Term Vector Error Model improves transmission measurements by calibrating for more mismatch and isolation errors as well as improving reflection measurements of two port devices. The 12 terms include 6 each for forward and reverse calibrations.



12-TERM BENEFITS MEASUREMENTS OF:

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  (Air Lines, Cables)
- High Insertion Loss
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- Highly Reflective Devices
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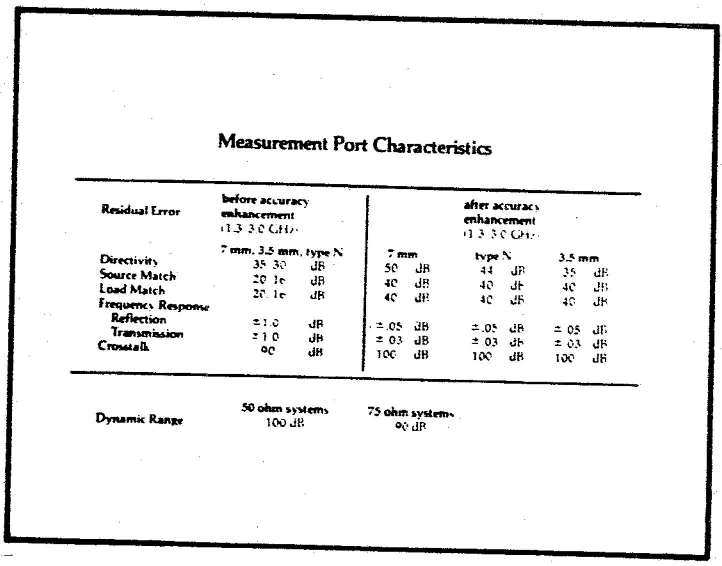


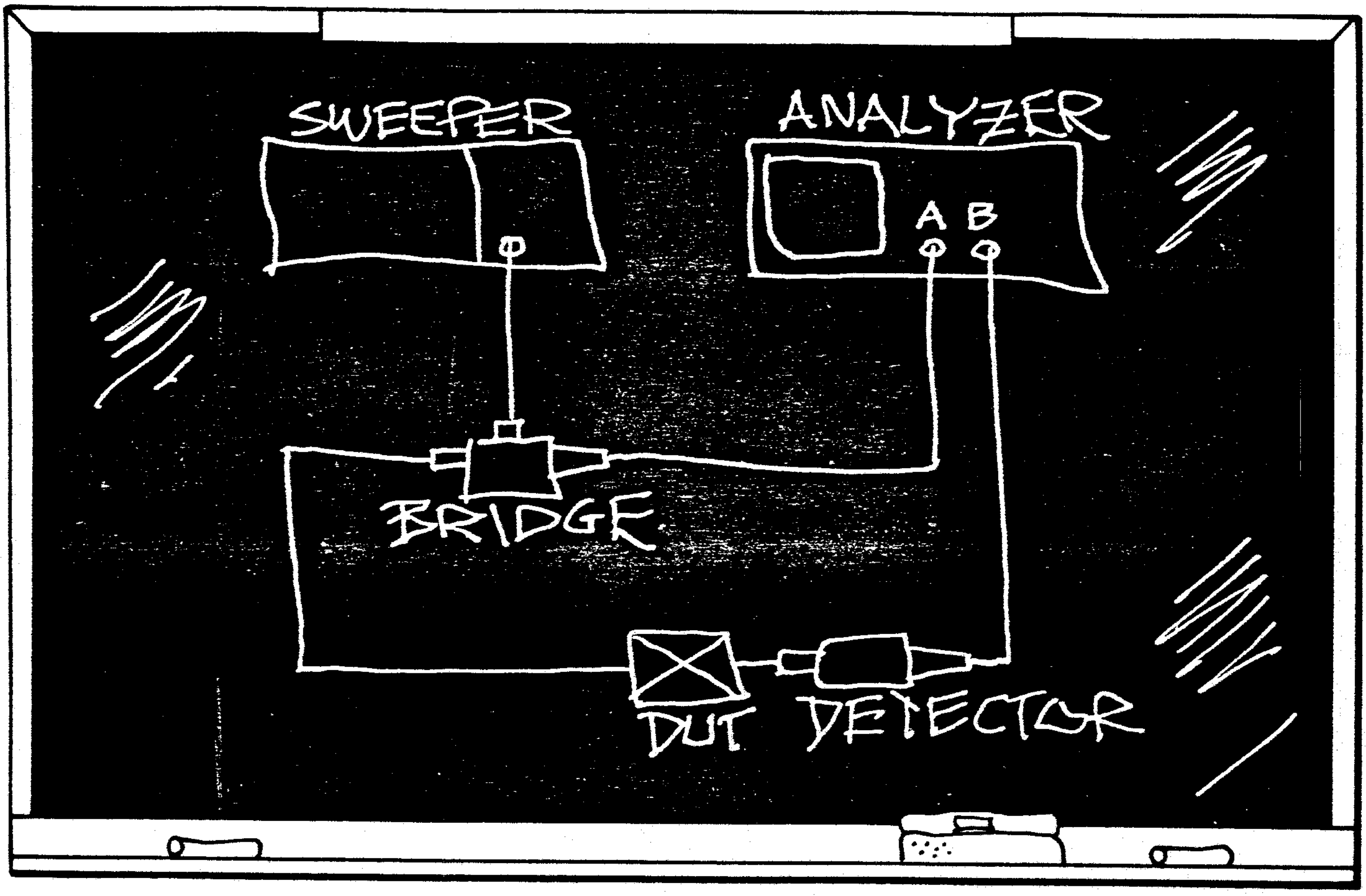
Measurement Port Characteristics

Residual Error	before accuracy enhancement (1.3-3.0 GHz)	after accuracy enhancement (1.3-3.0 GHz)		
	7 mm, 3.5 mm, type N	7 mm	type N	3.5 mm
Directivity	35/30 dB	50 dB	44 dB	35 dB
Source Match	20/16 dB	40 dB	40 dB	40 dB
Load Match	20/16 dB	40 dB	40 dB	40 dB
Frequency Response				
Reflection	±1.0 dB	±.05 dB	±.05 dB	±.05 dB
Transmission	±1.0 dB	±.03 dB	±.03 dB	±.03 dB
Crosstalk	90 dB	100 dB	100 dB	100 dB

Dynamic Range	50 ohm systems	75 ohm systems
	100 dB	90 dB

Also included in the HP 8753A data sheet are specifications for the main sources of error in a measurement. The error terms are specified both before and after accuracy enhancement, since uncorrected measurements are sometimes convenient.

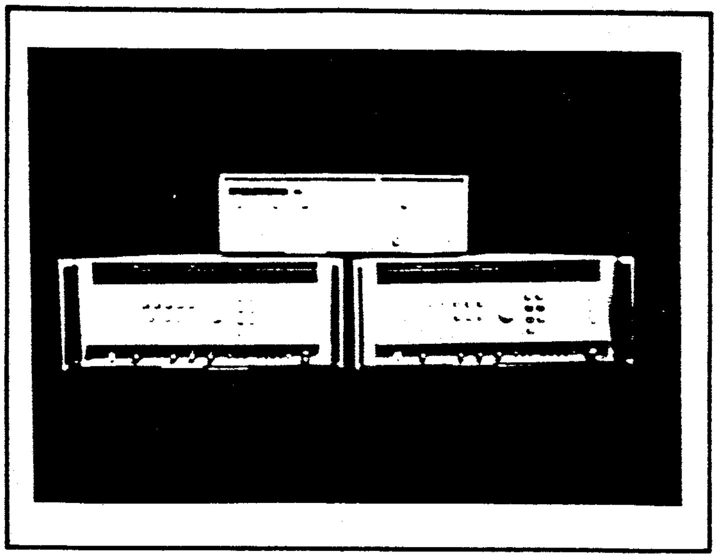
Measurement Systems



HP offers a complete product line for scalar network analysis from sources to network analyzers.

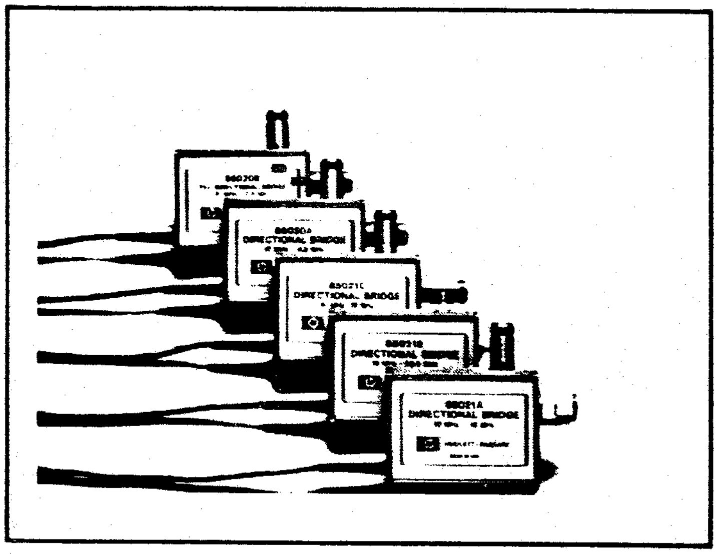
HEWLETT-PACKARD PRODUCT LINE OFFERING

- **Sources**
- **Signal Separation Devices**
- **Detectors**
- **Network Analyzers**

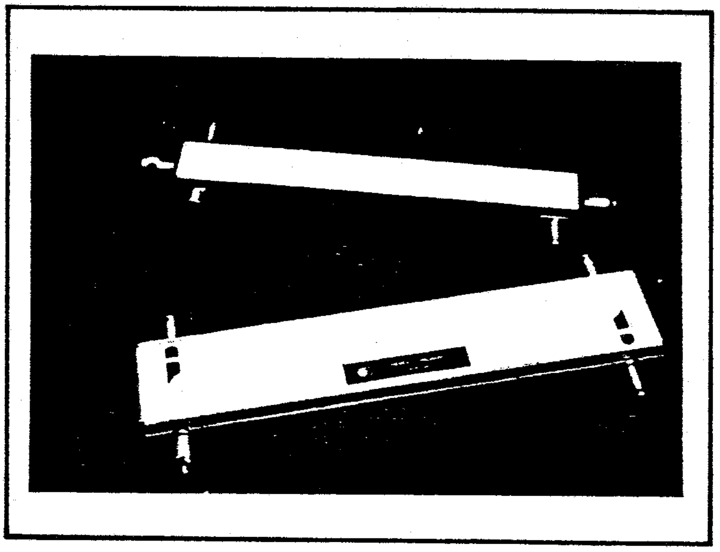


The HP 8350B and family of plug-ins offer a low cost, wide selection of swept sources.

The HP 8340/8341 Synthesized Sweepers offer both analog sweep and stepped CW sweep, with synthesizer accuracy CW frequencies.

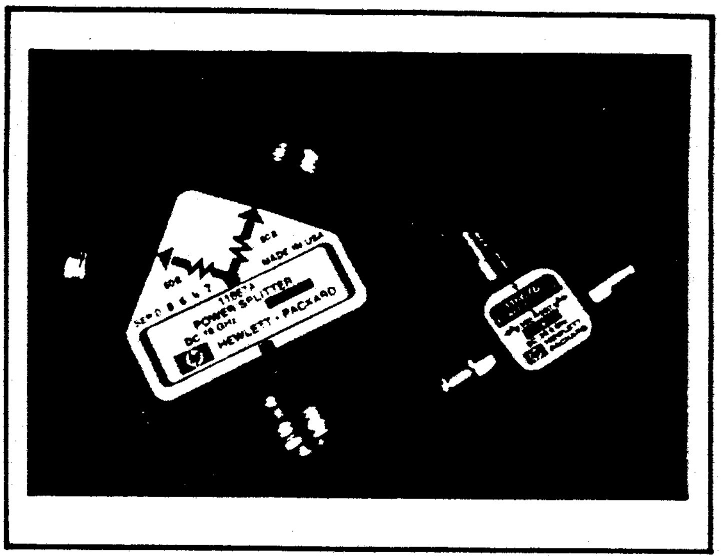


A full line of 40 dB directivity reflectometer bridges from 10 MHz to 40 GHz are available to separate reflection signals.

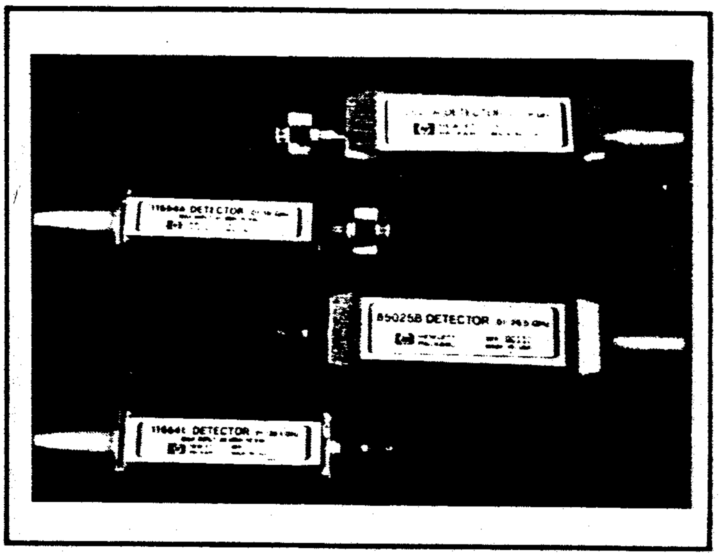


Broadband as well as octave band couplers are available.

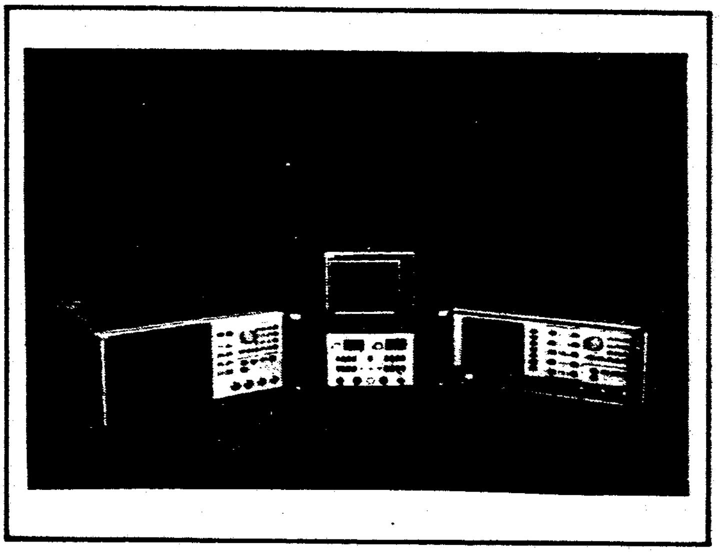
Power splitters with various connector types are also available.

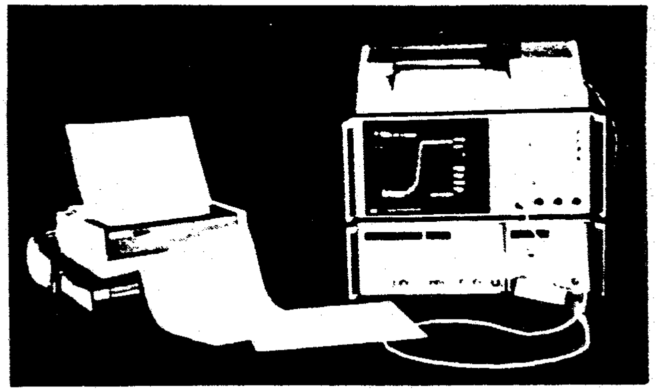


We have AC detectors from 10 MHz to 40 GHz and AC/DC detectors from 10 MHz to 26.5 GHz.

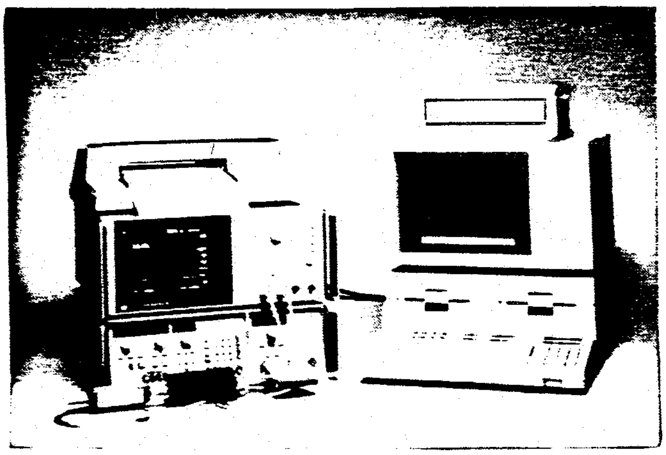


Three HP network analyzers are available with very friendly operation and superb performance. In the next session of the seminar, we'll show how all these instruments are used in specific applications.



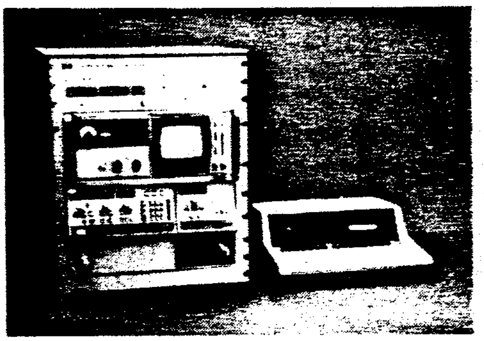


A complete measurement system might look like this, with a plotter and a printer.

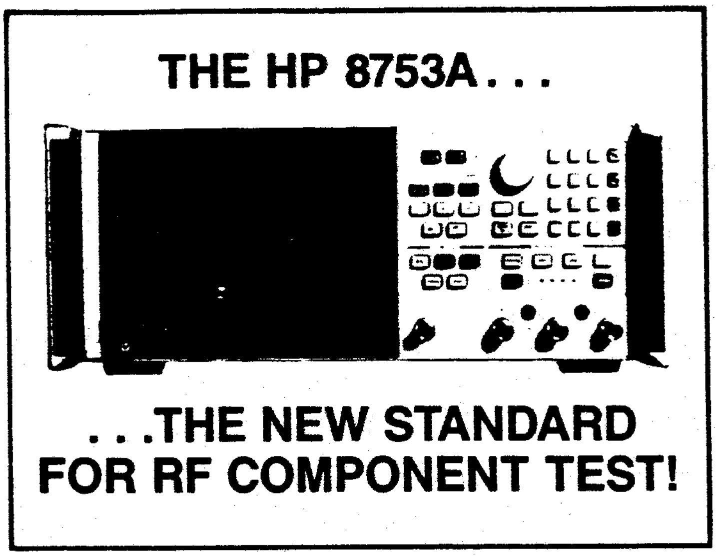


By adding a computer, you can automate your system to improve productivity. The last section of this seminar will discuss automatic measurements.

8410C MICROWAVE NETWORK ANALYZER



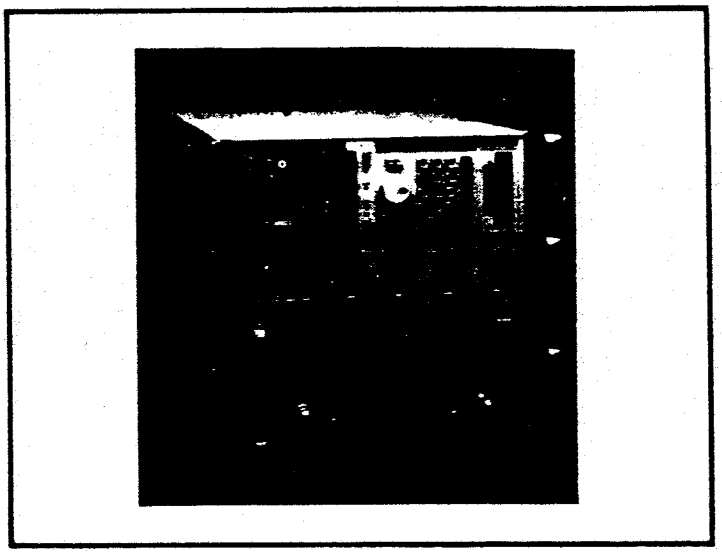
A sample vector network analyzer is the HP 8410C. Automatic versions are the HP 8408B (shown here) and the HP 8409C.



THE HP 8753A . . .

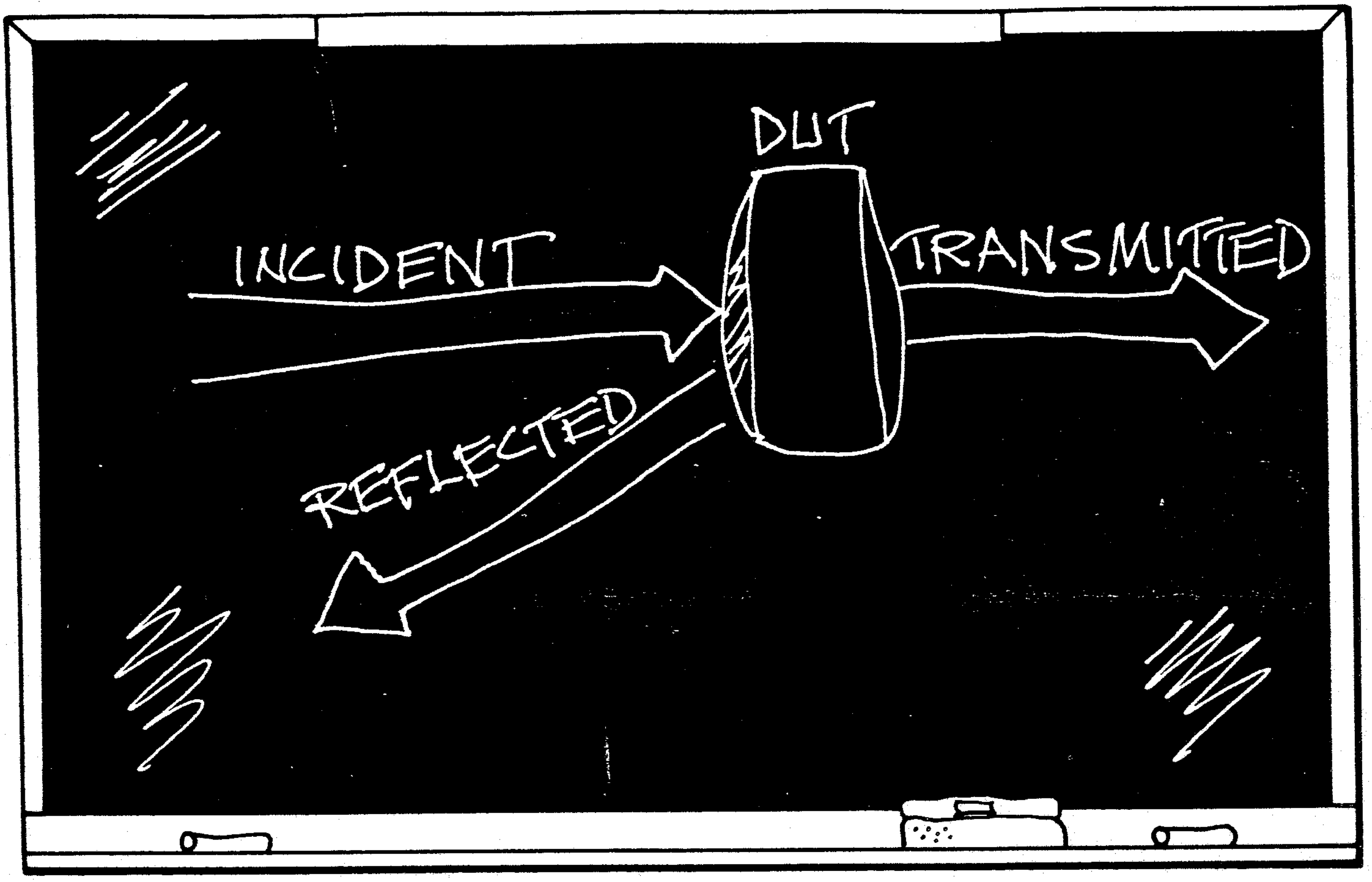
. . .THE NEW STANDARD FOR RF COMPONENT TEST!

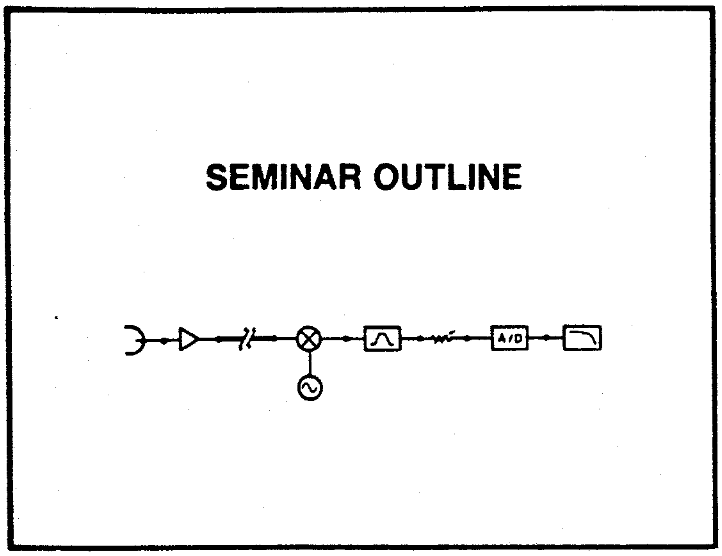
We feel the HP 8753A will become the "New Standard" of the RF Component Test industry, providing a cost-effective path for a wide variety of test applications. [4009]



...the HP 8510.

Applications

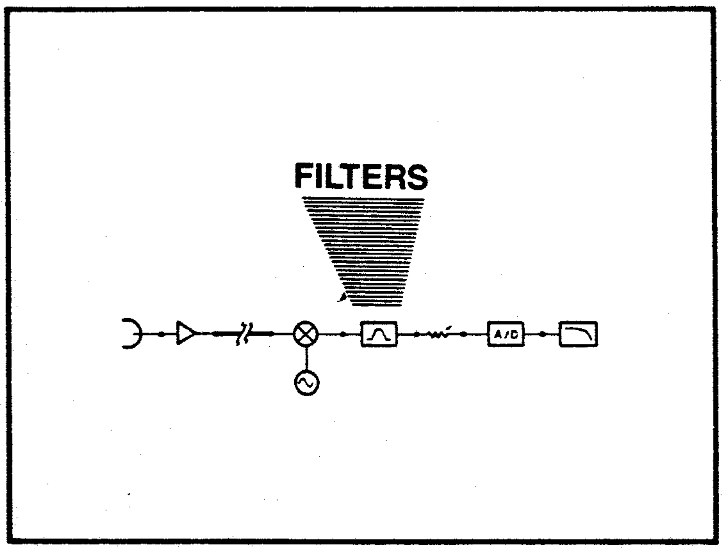




SEMINAR OUTLINE

This simplified block diagram of a basic superhetrodyne receiver will lead us through several common applications in the RF test world. This outline is not intended to portray any specific system, but does contain components found in many different RF systems.

ANTENNAS
AMPLIFIERS
CABLES
TRANSISTORS
FILTERS
ATTENUATORS
WAFER PROBING
FIBER OPTICS
PRODUCTION
 AUTOMATION



FILTERS

The first device we will investigate is the filter.

FILTER TEST CONSIDERATIONS

- Low Insertion Loss (.1 dB)
- Wide Dynamic Range (100 dB)
- Narrowband (Q's to 1000+)
- Electrically Long (to Kilometers)
- Distortion (Group Delay)
- Real-Time Tuning

Testing filters involves several important considerations. Many filters exhibit low passband insertion loss (on the order of tenths of a dB) and require high resolution and accuracy. Ultimate rejection of 100 dB is possible with multi-section filters. Many crystal and cavity filters are narrowband, with Q's of over 10000 possible. Some filters, such as surface acoustic wave (SAW) filters, are electrically very long, with several kilometers typical. Demanding distortion measurements are required on communications filters, such as those used in digital radio applications. Interactive tuning is also a key requirement in a production environment.

Next we'll look at a surface acoustic wave (SAW) filter. This filter is a brick-wall bandpass filter used to remove the effects of terrestrial interference on TVRO earth-station receivers. The filter is used in the 134 MHz IF of satellite receiver systems and is designed to optimize rejection at 10 MHz away from the center frequency (the 3 dB bandwith is about 15 MHz, making this a very abrupt filter). Bi-directional SAW filter designs tend to have relatively large passband insertion losses: the average loss of this filter is 23 dB. An ultimate rejection of over 80 dB is possible with this filter.

SAW FILTER

F_c = 134 MHz

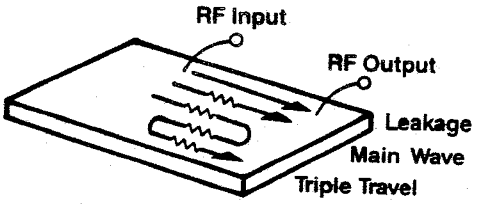
BW = 15 MHz

Insertion Loss
≈ 23 dB

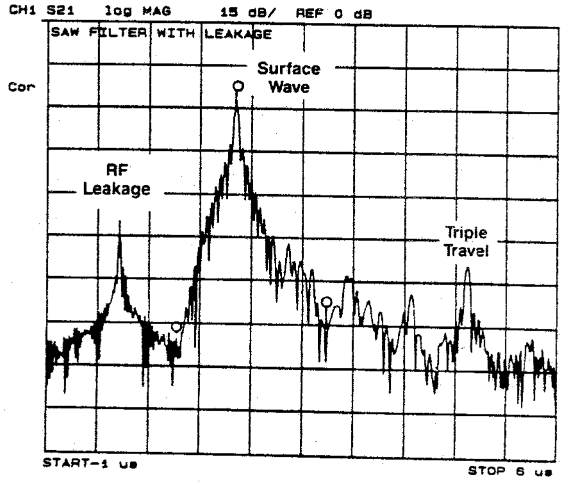
Ultimate Rejection
≥ 80 dB

Before discussing time domain in the HP 8753A, let's review SAW device physics a little. SAW devices operate primarily through the acoustic propagation of signals that add constructively at certain frequencies. Because the main wave propagates at near the speed of sound, SAW devices behave as if they are electrically long. The actual responses are separated in time and consist of the direct RF leakage or "punch-through", followed by the main acoustic response, and later by the triple travel of the main response after it bounces between ports. These undesired responses can limit the filter's performance.

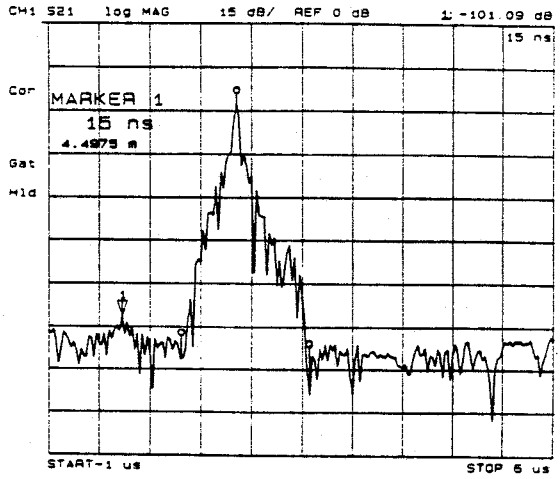
SAW FILTER PHYSICS



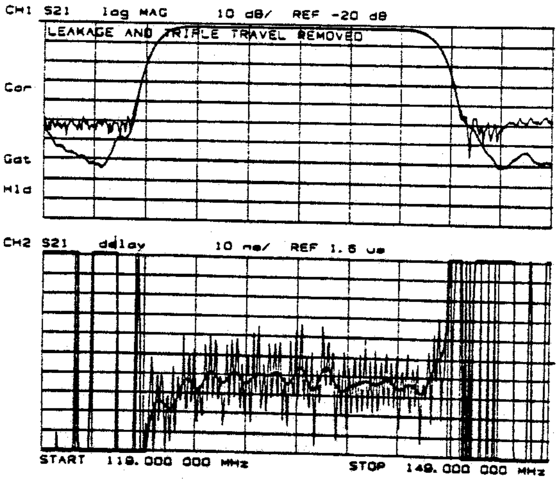
With the time domain option the HP 8753A can view the transmission response versus time, allowing us to clearly see individual responses separated in time. We will talk about time domain in more detail in a later section. For now, let us say that the analyzer gathers high accuracy, error-corrected data in the frequency domain, then transforms it into the time domain using an inverse Fourier transform. The resulting data retains all the accuracy inherent in the frequency domain. A digital filter, called a GATE, can be designed using the lollipop markers. When turned on, the GATE can selectively remove...



... the effects of signals separated in time. Here, we are making this filter appear to have very little leakage and triple travel. Likewise , the effects of a single error could be selectively removed or its response analyzed. When we return to the frequency domain the result is more striking.

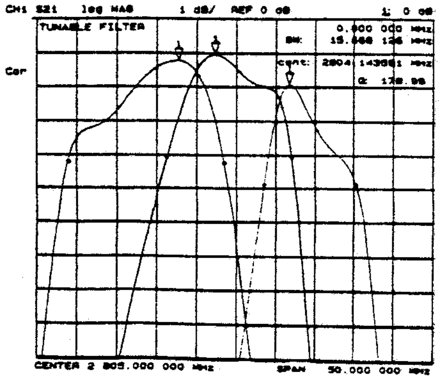


Notice the theoretical improvement in rejection and distortion. By viewing the effects of the leakage term it can be shown that the rejection was primarily limited by RF leakage from input to output. In this filter, the designer achieved 80dB rejection by adding an RF shield between ports. In production, this technique can be used for general purpose troubleshooting, such as isolating faulty grounds.



With TRACKING selected the marker searches continuously from sweep to sweep. This allows you to perform interactive adjustments while viewing 3 or 60 dB bandwidths. Bandwidths can be measured with respect to a fixed frequency, to the center frequency, or to the maximum or minimum amplitude with MAX or MIN SEARCH enabled.

TRACKING BANDWIDTHS

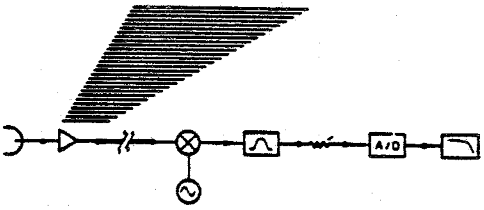


We have seen how important marker resolution and dynamic range are when measuring filters. The built-in synthesized source lets you accurately measure narrowband and electrically long filters. Optional time domain provides insight into SAW device design and manufacturing. The tracking markers simplify the tuning process. The HP 8753A makes precision group delay measurements on the most demanding filters. Finally, the frequency list mode saves you time in the production environment while letting you test at only the frequencies you really need.

8753A FILTER TEST SUMMARY

- .001 dB/.01°/1 psec Marker Resoluti
- 100 dB Dynamic Range
- 1 Hz Synthesized Source
- Time Domain (SAWs)
- Tracking Markers
- List Frequency Sweep

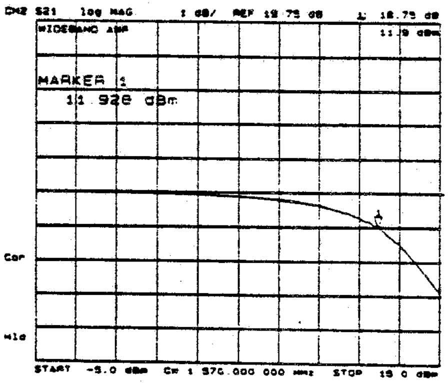
The next device we are going to discuss is the amplifier.

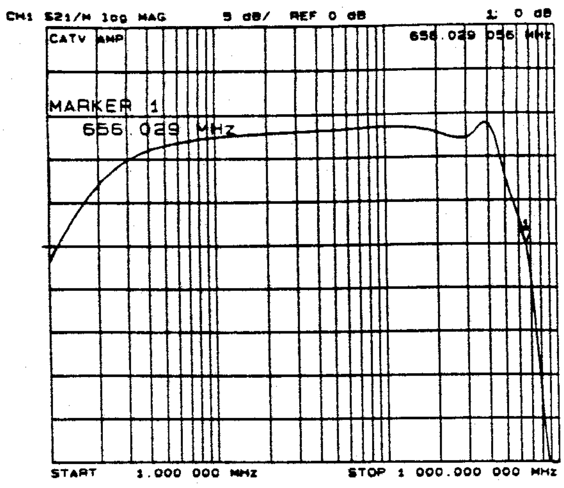


AMPLIFIERS

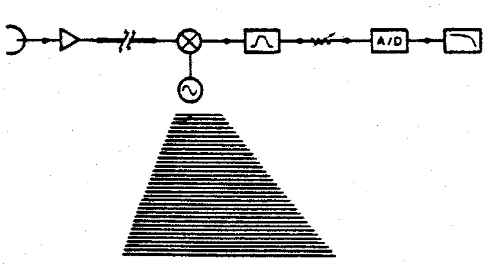
Many amplifiers exhibit non-linear input impedances. In these cases swept gain compression, though convenient, may not be appropriate. Another method is to sweep the input power at a CW frequency. Here we sweep from -5 to +15 dBm using the power sweep capability of the HP 8753A. The test set attenuation of 13 dB results in a input power sweep to our device of -18 to +2 dBm. The output power sweeps from +2 to compression and is attenuated by the pad by 20 dB and the test set another 6 dB.

CW POWER SWEEP COMPRESSION





Gain can also be measured with a logarithmic frequency sweep. In this mode each log point is measured rather then derived from a linear sweep. This mode allows you to generate BODE plots and look at gain margin. Our CATV amplifier shows significant rolloff above 450 MHz.



TRANSISTORS

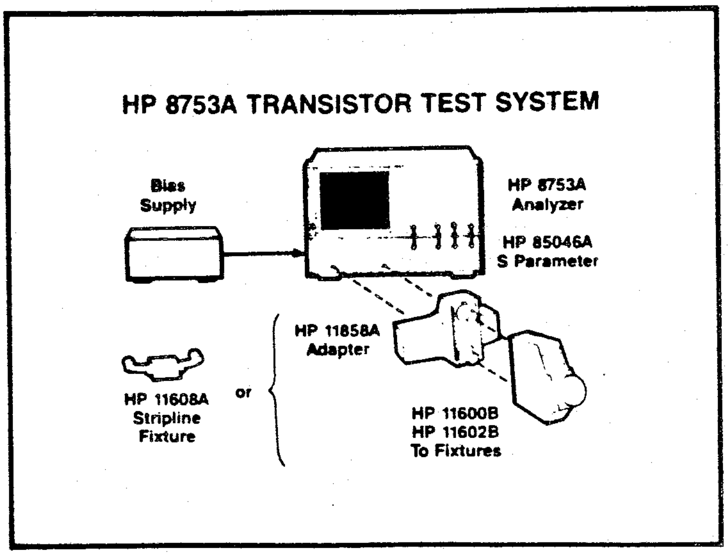
Now we move on to packaged transistor testing.

TRANSISTOR TEST CONSIDERATIONS

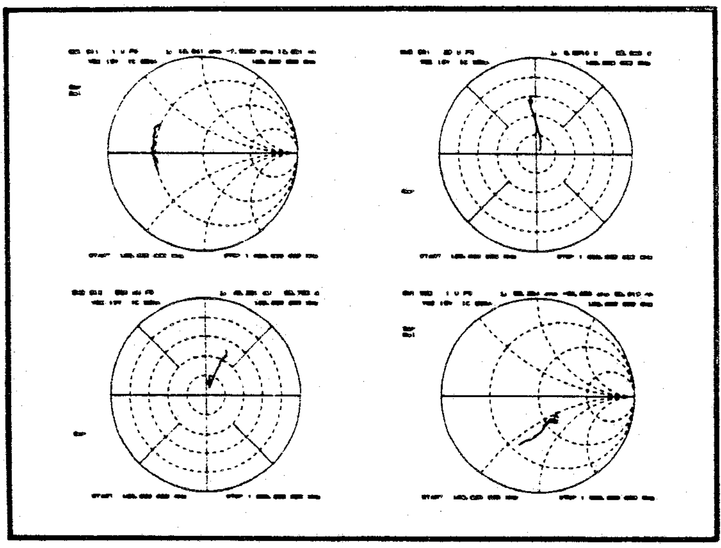
- S Parameter Test Set
- Accuracy Enhancement
- Impedance and Admittance
- Biasing
- Fixturing

Packaged transistors can be difficult to test accurately. A typical transistor test requires a 2 port test set and full accuracy enhancement to remove the effects of the instrumentation and fixtures from the measurements. Additional information such as impedance and admittance is needed for matching applications. The analyzer must also be able to provide DC bias to the transistor under test. Fixturing is another consideration, with commercial availability of fixtures for common TO and stripline packages a real plus.

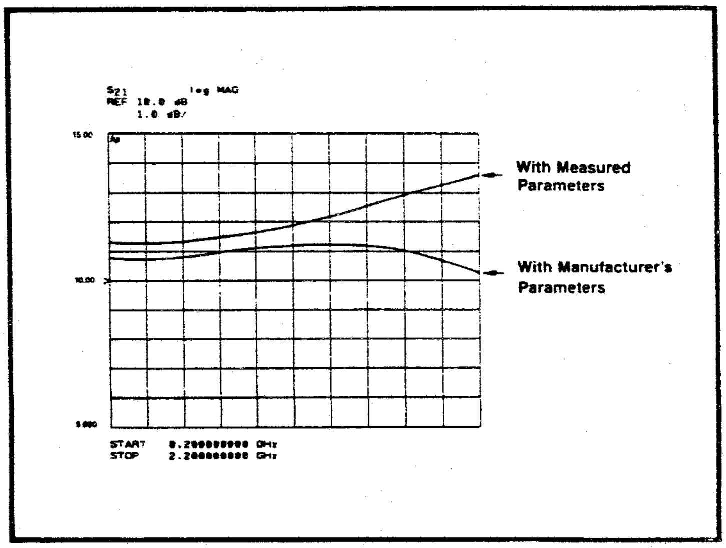
The HP 8753A network analyzer can be configured to make accurate and repeatable measurements on a wide variety of TO and stripline packages. A typical TO package test system includes an S parameter test set, a fixture adapter (adapts the hoizontal axis of the test set to the vertical axis of the fixtures), one of two HP transistor fixtures (the HP 11600B for TO-18/TO-72 packages and the 11602B for TO-5/TO-12 packages), and a bias supply. For stripline packages HP makes the 11608A transistor fixture. The standard 11608A comes with a through-line microstrip and a bolt-in ground structure which can be user-machined to accomodate up to .45 inch diameter packages. A pre-machined option is available for .205 inch diameter packages such as the HPAC-200 or "K" Disc packages (.205 inch diameter X .020 inch thick).



With built-in plotter dump, the HP 8753A quickly provides a hardcopy of our transistor's S-parameters. The HP 8753A can act as a system controller on HP-IB, and plot directly to most HPGL plotters without the need for a computer. This is an example of the ability of the HP 8753A to plot on different quadrants of the page.



Circuit design using transistor parameters is a common application. The ability to tie into a circuit simulator or optimizer is greatly desired by RF engineers. There are several such programs available for HP computers as well as others including SPICE, SUPERCOMPACT, and TOUCHSTONE. High accuracy measured S-parameters can be used in place of manufacturers typical specifications to improve modeling and design ability. Actual measurements allow you to gain insight to the devices operation with unspecified operating conditions, such as different bias levels. Here we see a FET feedback amplifier simulated with manufacturer's parameters for the FET, then substituting measured parameters. The measured parameters yield a predicted performance within the 11 dB gain the circuit was designed for.



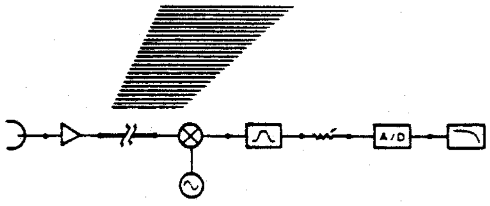
In summary, the S-parameter test set was required to obtain full error correction in both directions. Having accuracy enhancement built in is very convenient for transistor test applications. The impedance and admittance conversions are helpful in designing matching networks. For biasing, both the T/R test set and the S-parameter test set have bias TEEs built in and can handle up to 500 mA. Finally, HP supplies several fixtures for TO and stripline device characterization.

8753A TRANSISTOR TEST SUMMARY

- **85046A S Parameter Test Set**
- **Built-in Accuracy Enhancement**
- **Impedance and Admittance**
- **Built-in Bias TEEs**
- **Variety of Fixtures**

Next we move on to coaxial cable testing.

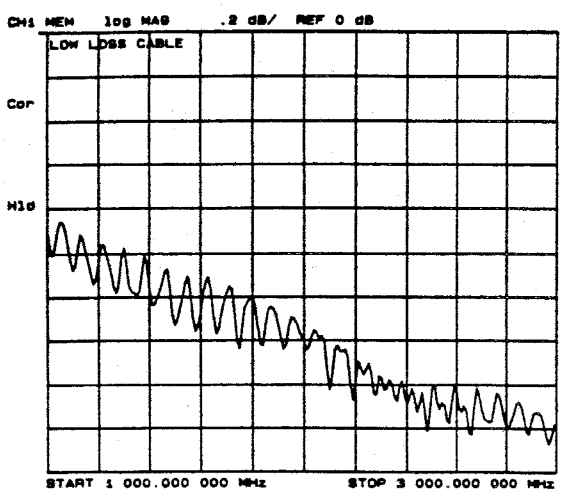
CABLES



RF cables are conceptually very simple, but can be challenging to test. Insertion loss and return loss measurements can be difficult on very low-loss cables. Electrical length is sometimes required with high resolution, or cables must be matched to demanding tolerances. Cable fault location may be desired. Cable manufacturers, among others, also need to test dielectric materials that are used in coaxial cable production.

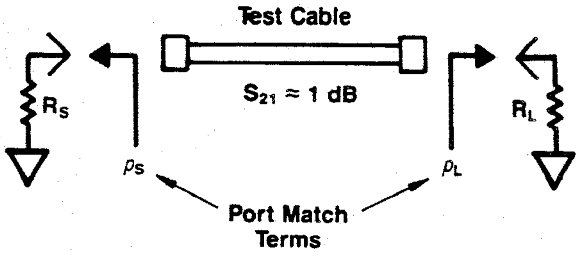
CABLE TEST CONSIDERATIONS

- **Low Insertion Loss (<.2 dB/ft. @ 1 GHz)**
- **Length/Phase Matching**
- **Fault Location**
- **Dielectrics**



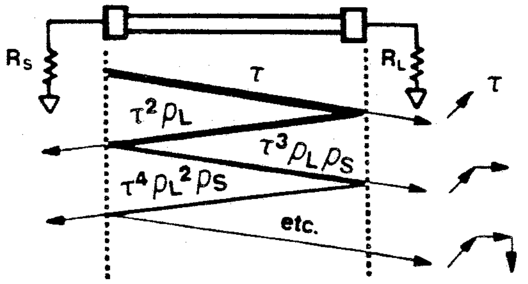
As you can see, above 1 GHz there is about 2 dB of peak-to-peak ripple on our insertion loss measurement. This effect is very common on low-loss cables and can present an error in our measurement. Let's take a closer look at this to see how we can reduce its effect.

TRANSMISSION MEASUREMENTS ON LOW-LOSS CABLES



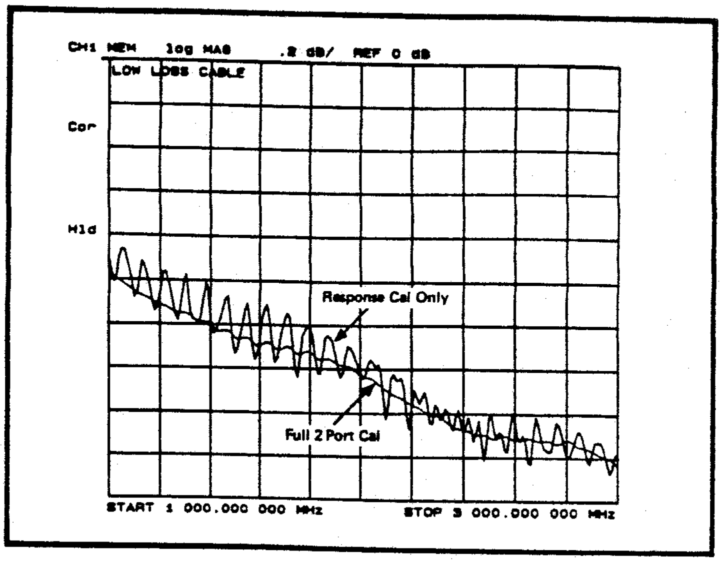
The real culprit in a low-loss transmission measurement is port match error. This is due to the non-ideal port match of the test equipment. The low loss of the test device can aggrivate this condition since it provides very little isolation between input and output. For example...

IMPORTANCE OF LOAD MATCH ON LOW-LOSS CABLES

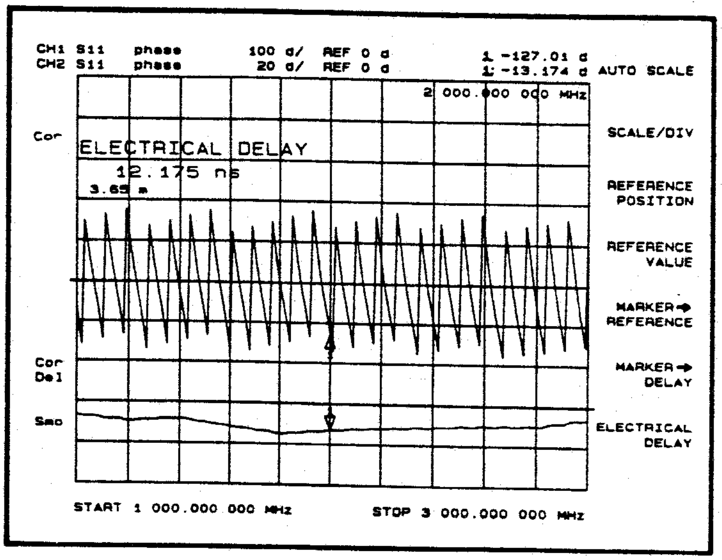


Here we can see how test equipment port match can create an error term that is scaled by the cable insertion loss as it bounces between the cable input and output. If the input source match and output load match can be measured, their effects can be reduced in our measurement. The HP 8753A does this as part of a 2 port calibration, making accurate measurements possible on low-loss devices like cables.

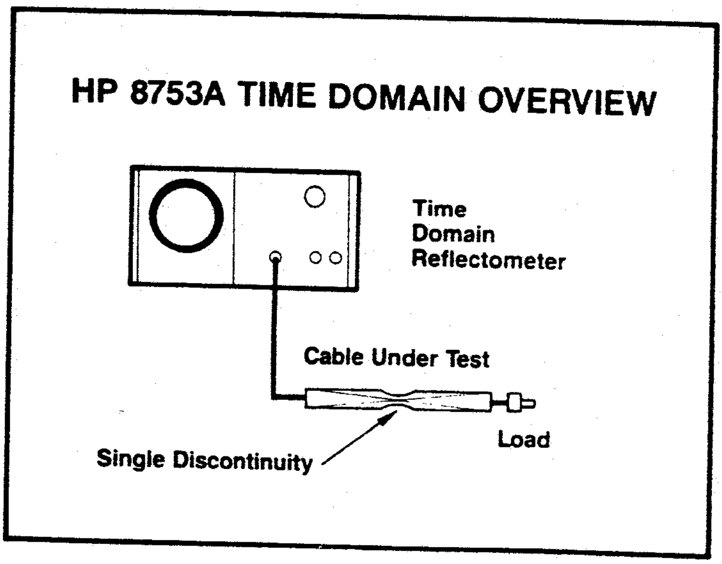
Notice the improvement in our measurement with the full 2 port calibration turned on. The load match effects have been mathematically removed using vector error correction. This requires a vector network analyzer because both the magnitude and phase of the load match term need to be measured. Scalar network analyzers can not correct for load match. Full built in accuracy enhancement has only recently become available in network analyzers.



Cable electrical length is a very common measurement. Phase matching of cables is required for phased array antennas, in computer system cables, and in communications systems. The HP 8753A can make this measurement with up to 1 picoseconds of resolution. The cable return loss phase is measured with a short on the end over the frequency range of interest. Using electrical delay offset we can manually subtract length mathematically from the measurement, or with the AUTO-DELAY softkey, until a flat phase response is obtained. The resultant delay is twice the average electrical length of the cable in air. Incredible resolution is one advantage of this mathematical technique.

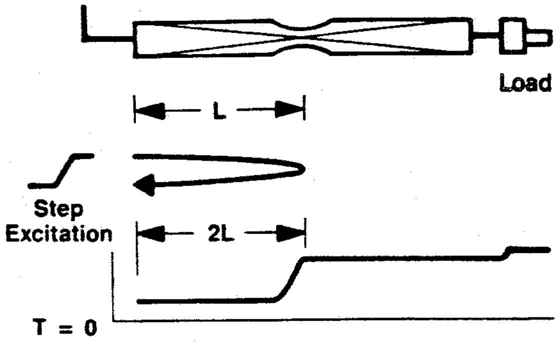


Before we get into cable fault location, this is a good time to review the basics of traditional time domain reflectometry. This will allow us to contrast how the HP 8753A option 010 time domain option operates, and how it benefits us. The traditional time domain reflectometer (TDR) is used to locate impedance discontinuities versus time along a length of cable.



HP 8753A TIME DOMAIN OVERVIEW

TDR FUNDAMENTALS



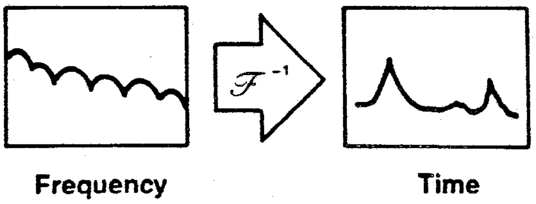
Take the case of this cable. The TDR generates a step or impulse excitation which propagates down the suspect cable. A discontinuity at length L would cause a reflection whose magnitude is proportional to the level of mismatch and whose phase is dependent on the reactive portion of the mismatch and the length L. The return signal is then displayed versus time (or distance in air) and is useful for locating faults due to loose connectors, deformation, or damage.

LIMITATIONS TO TRADITIONAL TDR

- Accuracy
- Reflection Only
- Not Frequency Selective
- DC Path

A conventional TDR has accuracy limited by errors such as uncorrected directivity and port match. Also, a TDR is limited to reflection while the HP 8753A can measure transmission versus time as well. The traditional TDR is not frequency selective. The broadband detectors respond to all signals within their bandwidth, making it difficult to evaluate faults in the presence of interference or over just a selected bandwidth. TDR also requires a DC path to transmit the stimulus signal. The HP 8753A, on the other hand, does not, and can measure non-DC-coupled systems such as filters and attenuators.

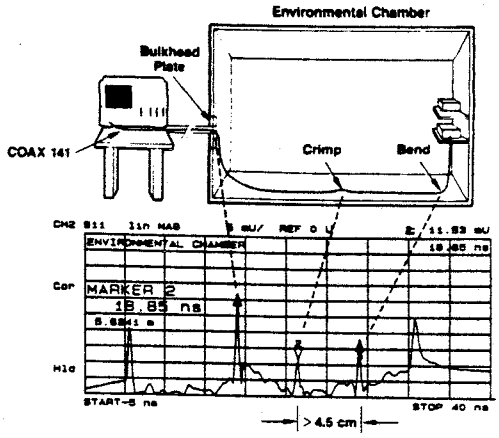
8753A OPTION 010 TIME DOMAIN



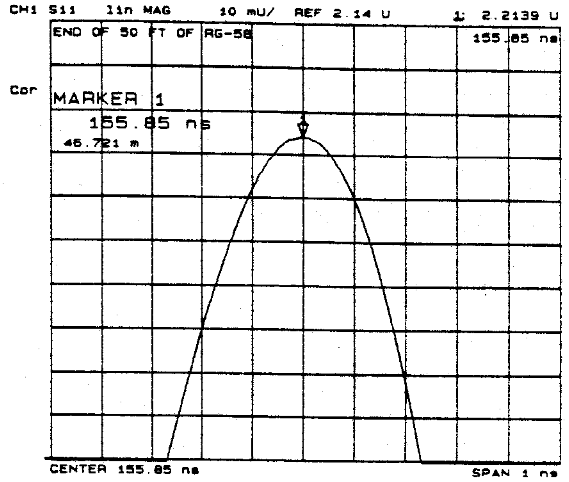
The HP 8753A option 010 allows us to compute the time domain response of devices like cables. Swept frequency domain measurements are transformed into the equivalent time domain response using an inverse Fourier transform. Time domain measurements retain all the error-corrected accuracy of the original swept response as well as up to 1601 points of resolution. In the past these measurements required the addition of a computer and could take many minutes to compute. The HP 8753A with option 010 can transform 401 points in less than a second!

Use as a diagnostic tool for an environmetal chamber is a good example. It could be very expensive to lower the chamber temperature and break the hermetic seal to fix a cable problem in the middle of a test.

8753A FAULT LOCATION



Range resolution is the analyzer's ability to resolve the location of a single fault. Range resolution is limited by the analyzers' source stability, linearity, and time-base accuracy. On the HP 8753A in a reflection measurement this is about 1 picosecond or 150 micrometers in air. Here we see the location of the open at the end of a 50 foot section of RG-58 coaxial cable. The velocity factor of the cable dielectric can also be entered. This factor is used for all length computations.

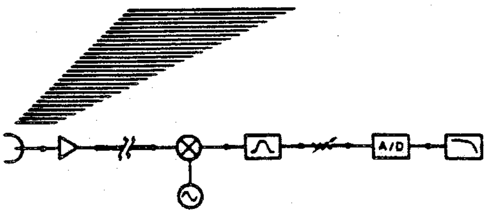


In summary, the time domain capability of the HP 8753A is more than just TDR. All time domain measurements retain the same error-corrected accuracy that we saw in the frequency domain, as well as up to 100 dB of dynamic range. The BANDPASS stimulus mode allows us to measure in tuned and high-pass systems. We have also seen that there are two important types of resolution to consider. The 1601 point mode means you can measure broadband cables over long distances (160 metres for a full sweep and 1601 points). All this capability in compact, integrated, and lightweight package is a plus for measuring in remote or unusual locations, especially since no computer is required to perform basic fault location measurements. Finally, time domain is a powerful add option to the basic network analyzer itself.

8753A FAULT LOCATION SUMMARY

- **TDR Plus:**
  - **— Error-Corrected Accuracy**
  - **— 100 dB Dynamic Range**
  - **— Frequency Selective**
  - **— Excellent Range Resolution**
- **Up to 1601 Points**
- **Compact & Lightweight Package (No Computer Required)**
- **Option 010 to Basic Instrument**

ANTENNAS



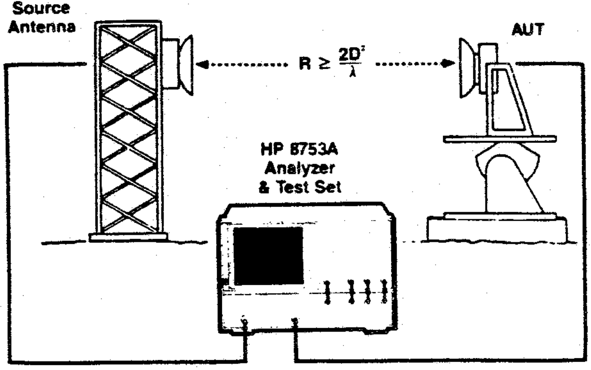
Typical requirements for the testing of antennae include a receiver with wide dynamic range. This is especially true for testing over long distances where the 1/R loss can be considerable. The antenna test engineer must also consider the errors due to unwanted responses. Direct display of SWR and the removal of the effects of long cable runs are also common requirements. Finally the tests involved may include both swept and CW measurements.

ANTENNA TEST CONSIDERATIONS

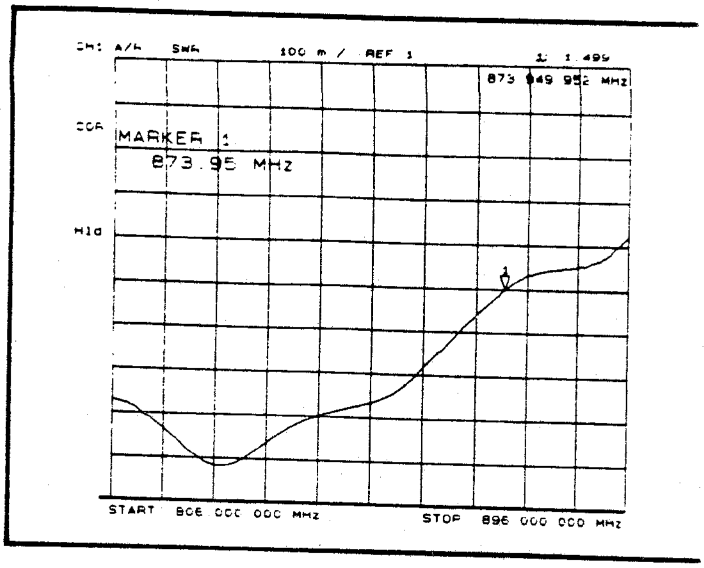
- **Wide Dynamic Range (20 m @ 3 GHz ≈ 68 dB)**
- **Unwanted Responses**
- **SWR**
- **Swept and CW**

Far-field tests are typically done in an outdoor elevated range at RF frequencies, using a standard gain receiving antenna, although some work is being done in large anechoic chambers. The distance between antennas is determined by the fraunhofer or "far-field" zone which is determined by the range equation: R=2*(D*D)/lambda, for a 22.5 degree maximum phase error. At lower frequencies the antenna apertures required are quite large, making far-field distances to tens of metres common. The long distances result in large free-space losses, requiring a wide dynamic range receiver. The vertical height of the antenna towers is typically tuned to minimize the effects of unwanted range or room reflections, which are difficult to remove and bothersome to antenna engineers.

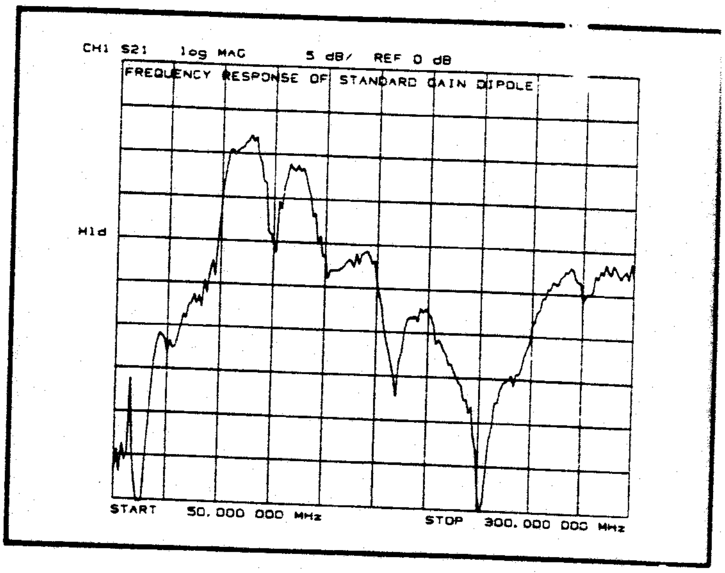
FAR FIELD ANTENNA RANGE



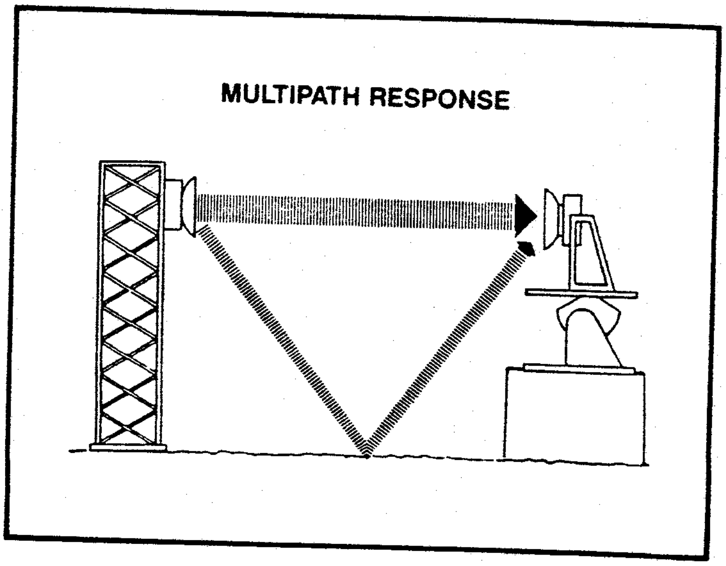
The 8753A can directly display swept SWR measurements. The scale can be easily entered or use the popular AUTOSCALE key. This antenna's SWR is out of specification at the high end.

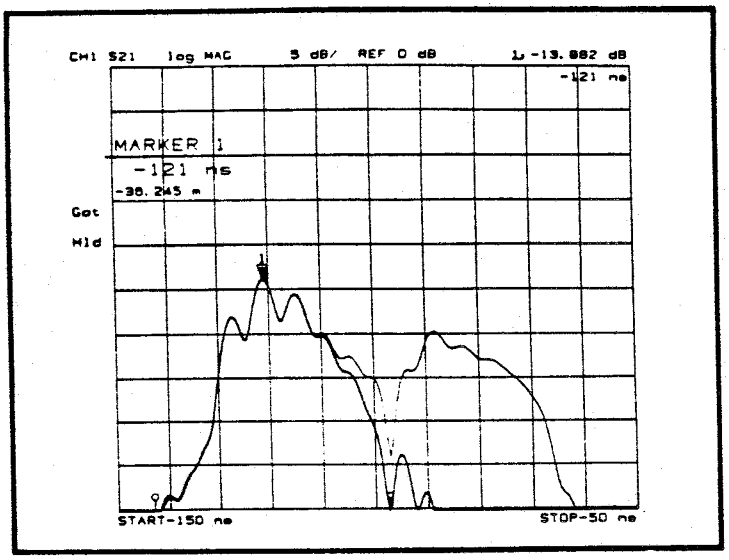


The transmission response of a standard gain (balanced) dipole antenna is shown here between 50 and 150 MHz. This antenna is difficult to measure since it is omnidirectional. Note the ripple pattern superimposed on the actual response of the antenna. What do you think causes this ripple effect in our measurement?

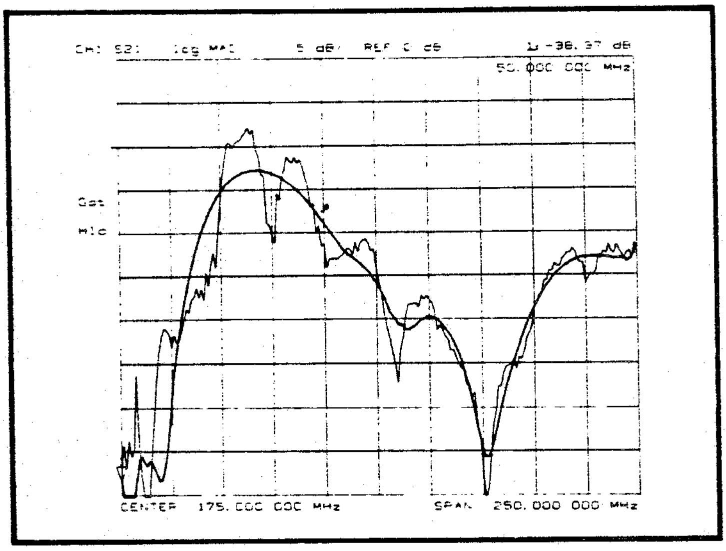


As we can see, the ripple is due to the multiple paths of the excitation signal. These reflections are due to real world things such as multipath from surrounding hills, cars, cables, railings, nails, practically anything! It would be advantageous if our receiver could distinguish between the desired and undesired signals.



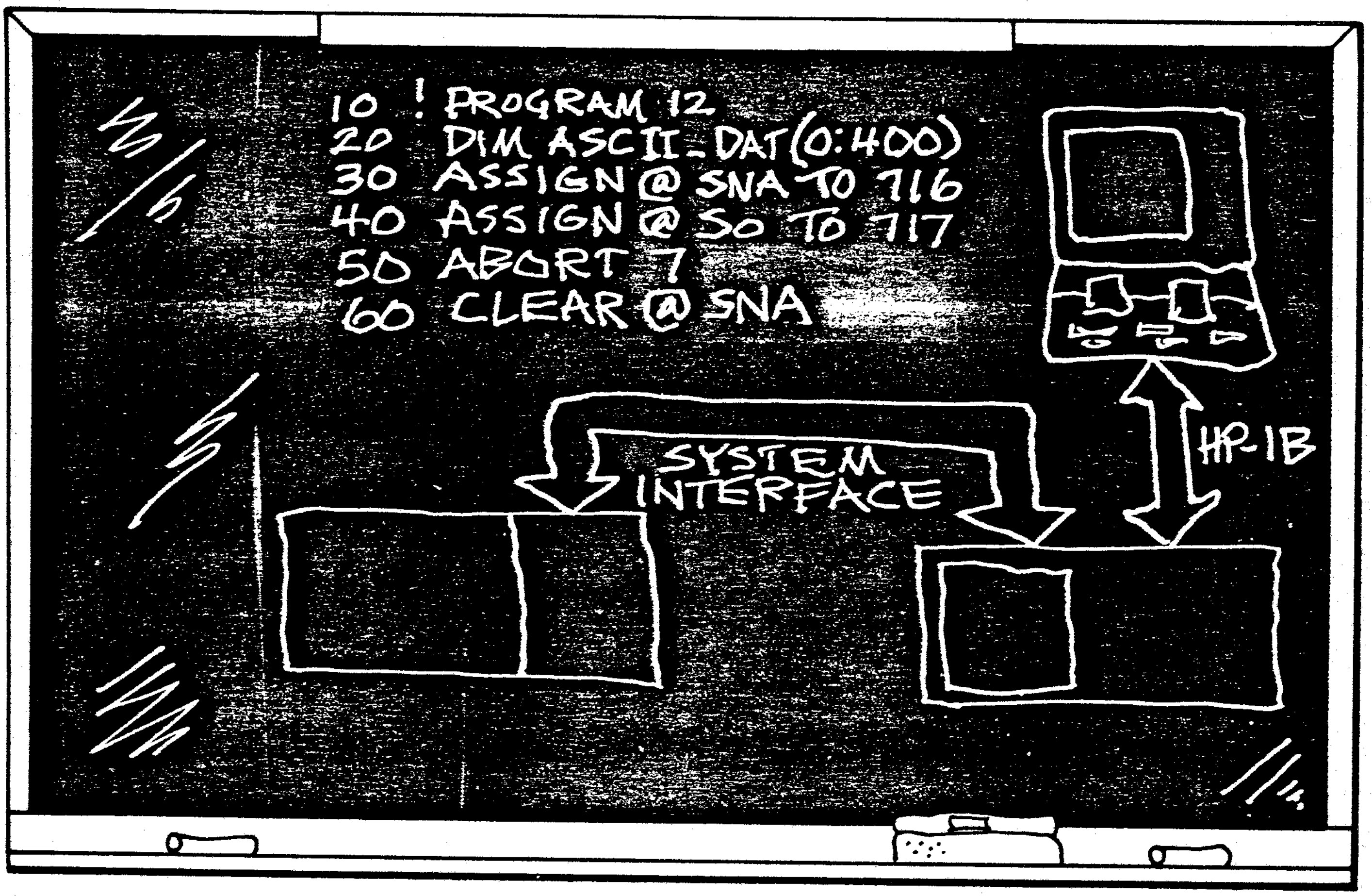


With time domain the HP 8753A has the capability to digitally filter the data to remove the effects of unwanted responses seperated in time. Here we see the filter, called a gate, centered around the main response. Turning the Gate on drops the multipath responses into the noise. Thus when we view the gated response back in the frequency domain...



...we see that the ripple has been removed to provide us a more accurate picture of the antenna's transmission response.

Automatic Systems



```
10 ! PROGRAM 12
20  DIM ASCII_DAT(0:400)
30 ASSIGN @ SNA TO 716
40 ASSIGN @ So TO 717
50 ABORT 7
60 CLEAR @ SNA
```

Prewritten Software also come with thorough documentation in the form of an operating manual.

Documentation of Contracted Software can be provided by your HP Systems Engineer.

Documentation of User Generated Software is left to the user to develop.

DOCUMENTATION

Prewritten Software Pacs	Software Operating Manual
Contracted Software	By Contract from Your HP Systems Engineer
User-Generated Software	User-Developed

Prewritten Software have the advantage of that the Operating Manual is frequently updated.

Support of Contracted Software can be provided by your HP Systems Engineer.

Support of User Generated Software is left to the user to develop.

SUPPORT

Prewritten Software Pacs	Manual Change Sheets
Contracted Software	By Contract from Your HP Systems Engineer
User-Generated Software	User-Developed

The cost of Prewritten Software is almost always the most economical choice if the software meets your measurement needs.

Contracted Software is usually more expensive than the Prewritten software, but the software is written specifically for your application. Contracted Software is usually more economical than User Generated Software because it is written by experienced programmers.

User Generated Software is usually the most expensive. Let's take a closer look at why this is true.

COST

Prewritten Software Pacs	Low
Contracted Software	Medium
User-Generated Software	High