# Network Analyzer Basics

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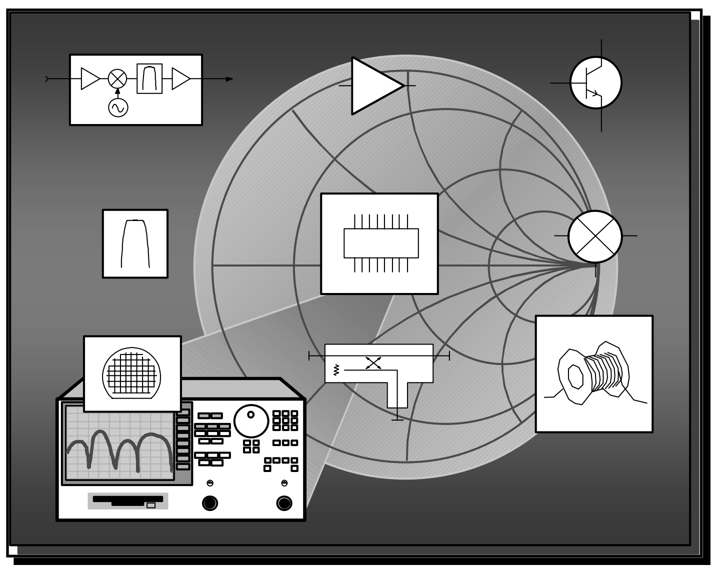


*1998 Back to Basics Seminar*

**Abstract**

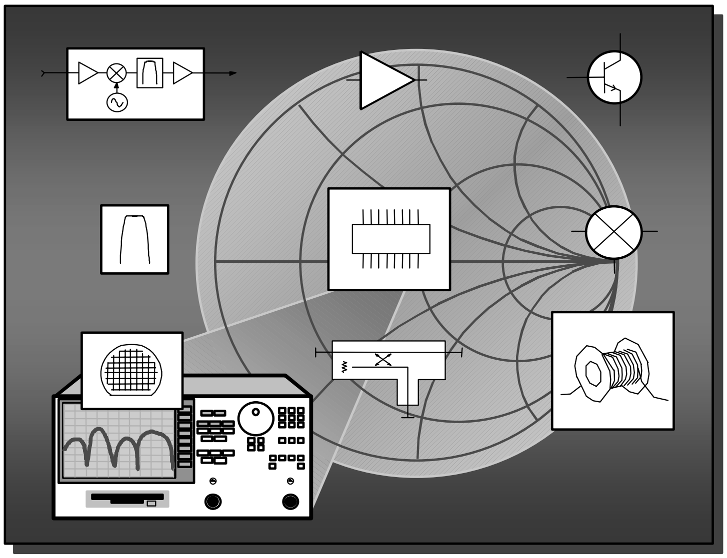
This presentation covers the principles of measuring high-frequency electrical networks with network analyzers. You will learn what kind of measurements are made with network analyzers, and how they allow you to characterize both linear and nonlinear behavior of your devices. The session starts with RF fundamentals such as transmission lines and the Smith Chart, leading to the concepts of reflection, transmission and S-parameters. The next section covers all the major components in a network analyzer, including the advantages and limitations of different hardware approaches. Error modeling, accuracy enhancement, and various calibration techniques will then be presented. Finally, some typical swept-frequency and swept-power measurements commonly performed on filters and amplifiers will be covered. An advanced-topics section is included as a pointer to more information.

**Author**

David Ballo is currently a Marketing Engineer for Hewlett-Packard's Microwave Instruments Division in Santa Rosa, California. David has worked for HP for over 17 years, where he has acquired extensive RF and microwave measurement experience. After getting a BSEE from the University of Washington in Seattle in 1980, he spent the first ten years in R&D doing analog and RF circuit design on a variety of Modular Measurement System (MMS) instruments. He followed that with a year in manufacturing. For the past six years, he has worked in the marketing department developing application notes, magazine articles, and seminar papers on topics including TWT amplifier test, group delay and AM to PM conversion of frequency-translating devices, adjacent-channel power measurements, design and calibration of RF fixtures for surface-mount devices, efficient test of multiport devices, and modern RF-design methodology.

**Slide #1**

# *Network Analyzer Basics*



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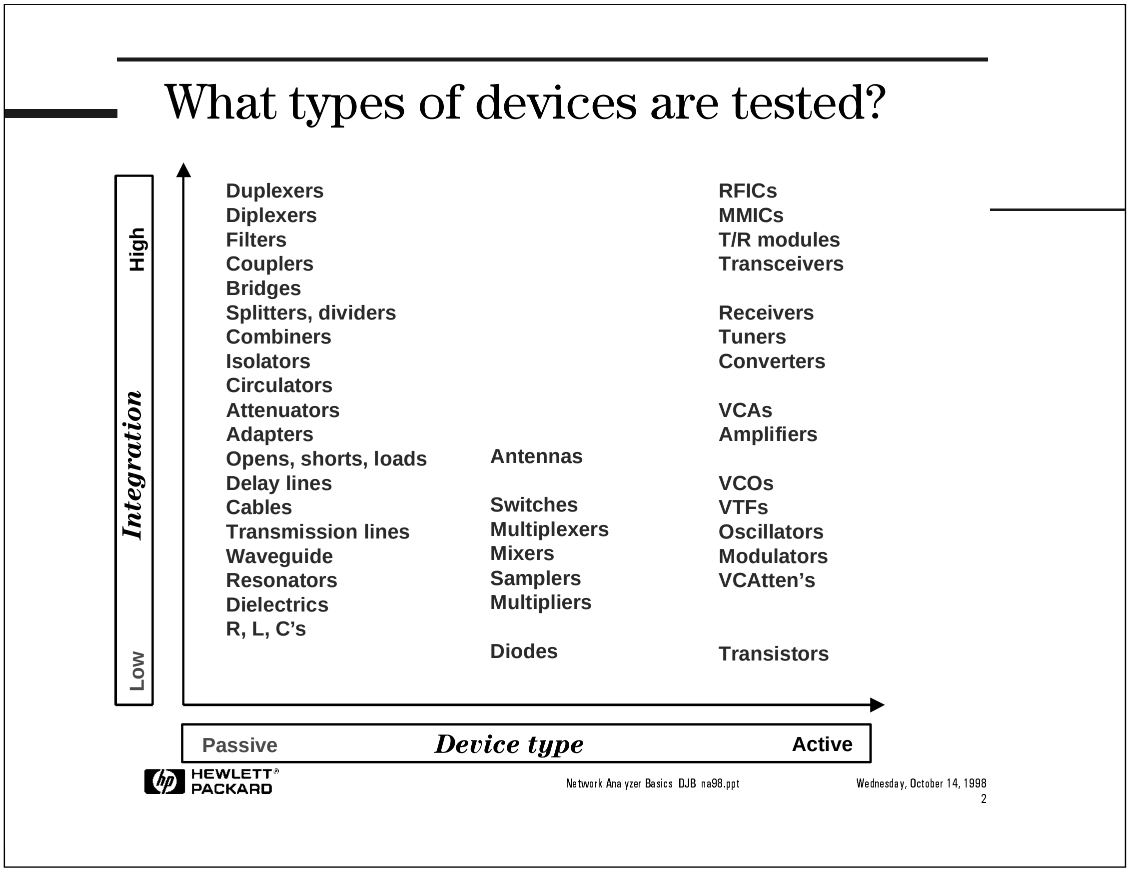
Network Analyzer Basics DJB na98.ppt

Wednesday, October 14, 1998

1

Welcome to Network Analyzer Basics.

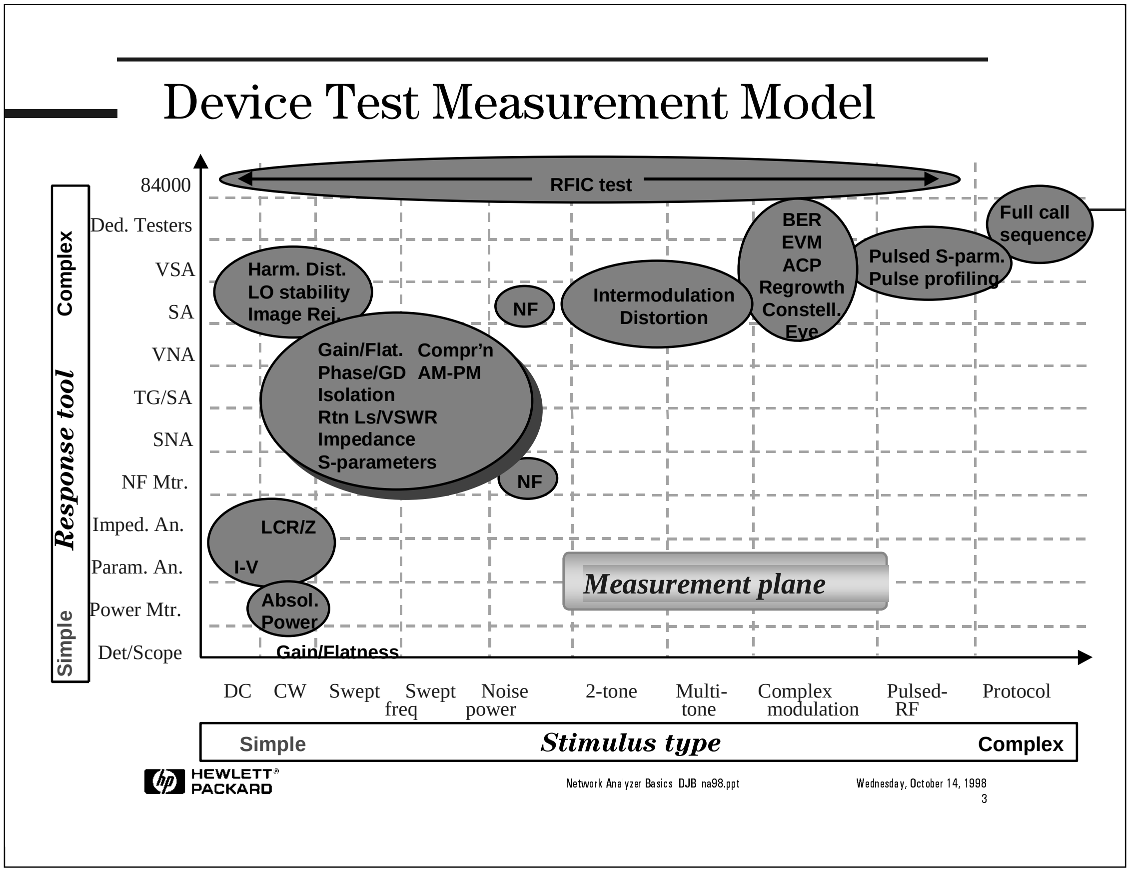
**Slide #2**




Here are some examples of the types of devices that you can test with network analyzers. They include both passive and active devices (and some that have attributes of both). Many of these devices need to be characterized for both linear and nonlinear behavior. It is not possible to completely characterize all of these devices with just one piece of test equipment.

The next slide shows a model covering the wide range of measurements necessary for complete linear and nonlinear characterization of devices. This model requires a variety of stimulus and response tools. It takes a large range of test equipment to accomplish all of the measurements shown on this chart. Some instruments are optimized for one test only (like bit-error rate), while others, like network analyzers, are much more general-purpose in nature. Network analyzers can measure both linear and nonlinear behavior of devices, although the measurement techniques are different (frequency versus power sweeps for example). This module focuses on swept-frequency and swept-power measurements made with network analyzers.

**Slide #3**




Here is a key to many of the abbreviations used above:

***Response***

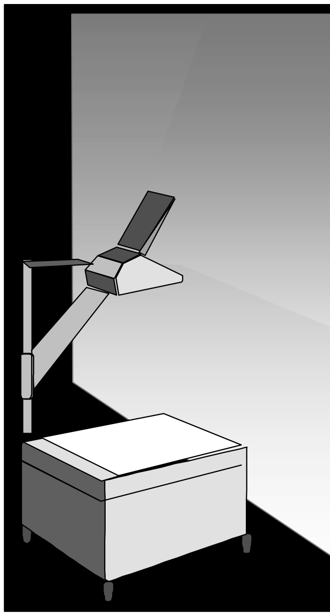
| | |
|---|---|
| 84000 | HP 84000 high-volume RFIC tester |
| Ded. Testers | Dedicated (usually one-box) testers |
| VSA | Vector signal analyzer |
| SA | Spectrum analyzer |
| VNA | Vector network analyzer |
| TG/SA | Tracking generator/spectrum analyzer |
| SNA | Scalar network analyzer |
| NF Mtr. | Noise-figure meter |
| Imped. An. | Impedance analyzer (LCR meter) |
| Power Mtr. | Power meter |
| Det./Scope | Diode detector/oscilloscope |

***Measurement***

| | |
|---|---|
| ACP | Adjacent channel power |
| AM-PM | AM to PM conversion |
| BER | Bit-error rate |
| Compr'n | Gain compression |
| Constell. | Constellation diagram |
| EVM | Error-vector magnitude |
| Eye | Eye diagram |
| GD | Group delay |
| Harm. Dist. | Harmonic distortion |
| NF | Noise figure |
| Regrowth | Spectral regrowth |
| Rtn Ls | Return loss |
| VSWR | Voltage standing wave ratio |

**Slide #4**

# Agenda



- **Why do we test components?**
- **What measurements do we make?**
  - Smith chart review
  - Transmission line basics
  - Reflection and transmission parameters
  - S-parameter definition
- **Network analyzer hardware**
  - Signal separation devices
  - Broadband versus narrowband detection
  - Dynamic range
  - T/R versus S-parameter test sets
- **Error models and calibration**
  - Types of measurement error
  - One- and two-port models
  - Error-correction choices
  - Basic uncertainty calculations
- **Typical measurements**
- **Appendix**

Network Analyzer Basics DJB na98.ppt

Wednesday, October 14, 1998

4

**Slide #5**

# Why do we need to test components?

**Components often used as building blocks**

- Need to verify specifications on:
  - filters to remove harmonics
  - amplifiers to boost LO power
  - mixers to convert reference signals

**When used to pass communications signals, need to ensure distortionless transmission**

- Linear networks
  - constant amplitude
  - linear phase / constant group delay
- Nonlinear networks
  - harmonics, intermodulation
  - compression
  - noise figure

**When absorbing power (e.g. an antenna), need to ensure good match**

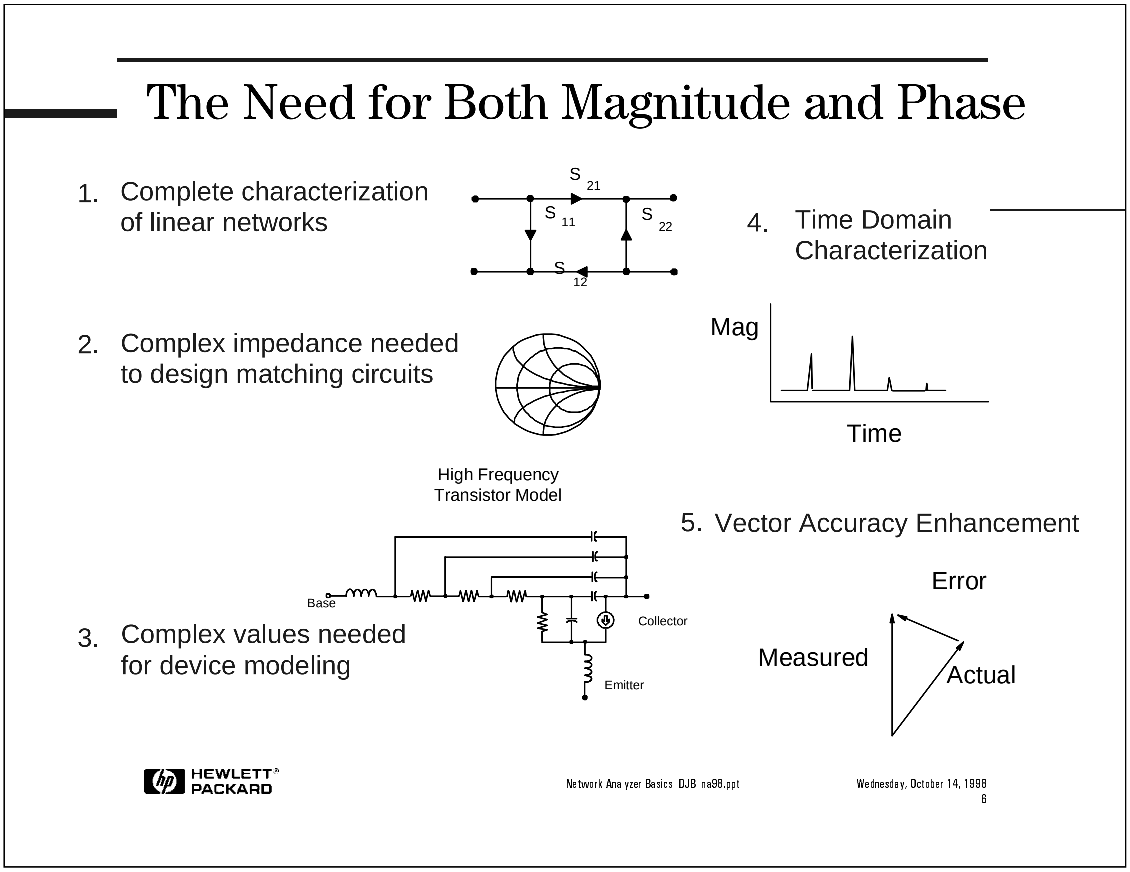
Network Analyzer Basics DJB na98.ppt Wednesday, October 14, 1998 5

Components are tested for a variety of reasons. For example, R&D engineers need to verify their designs, and production engineers need to verify proper performance. Without testing, a manufacturer cannot tell a prospective customer how a particular component will behave in a specific environment.

If a component is used in a communications system, distortion is of paramount concern. You are probably comfortable with the concept of non-linear (harmonic) distortion. When we discuss transmission, we will see that devices operating linearly can also introduce distortion.

We will start our discussion of measurements by reviewing transmission lines. Generating RF and microwave energy is expensive; we cannot afford to throw it away due to unnecessary losses. We will see what is required to efficiently transfer RF energy from one circuit, network, or system to another.

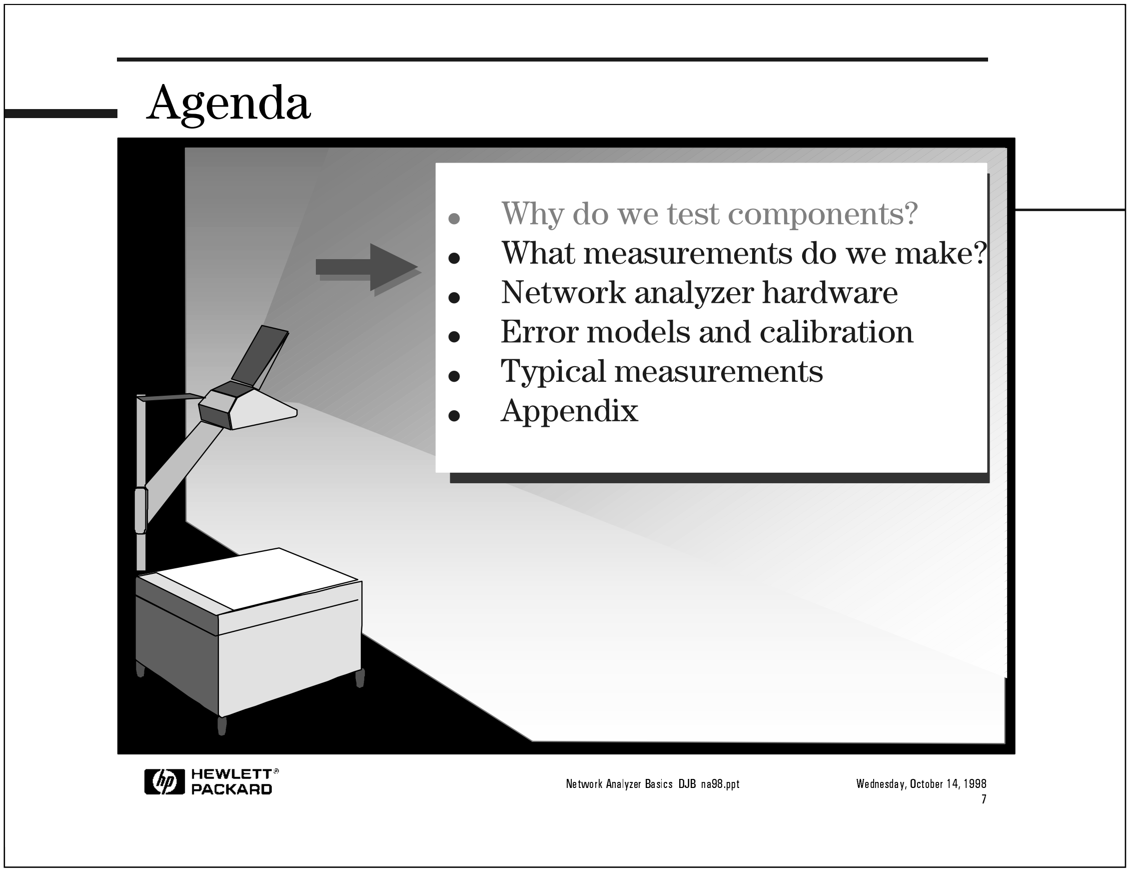
**Slide #6**




In many situations, magnitude-only measurements are sufficient. For example, the simple gain of an amplifier or stop-band rejection of a filter may be all the data that we need. However, there certainly are situations in which phase is a critical element of the data.

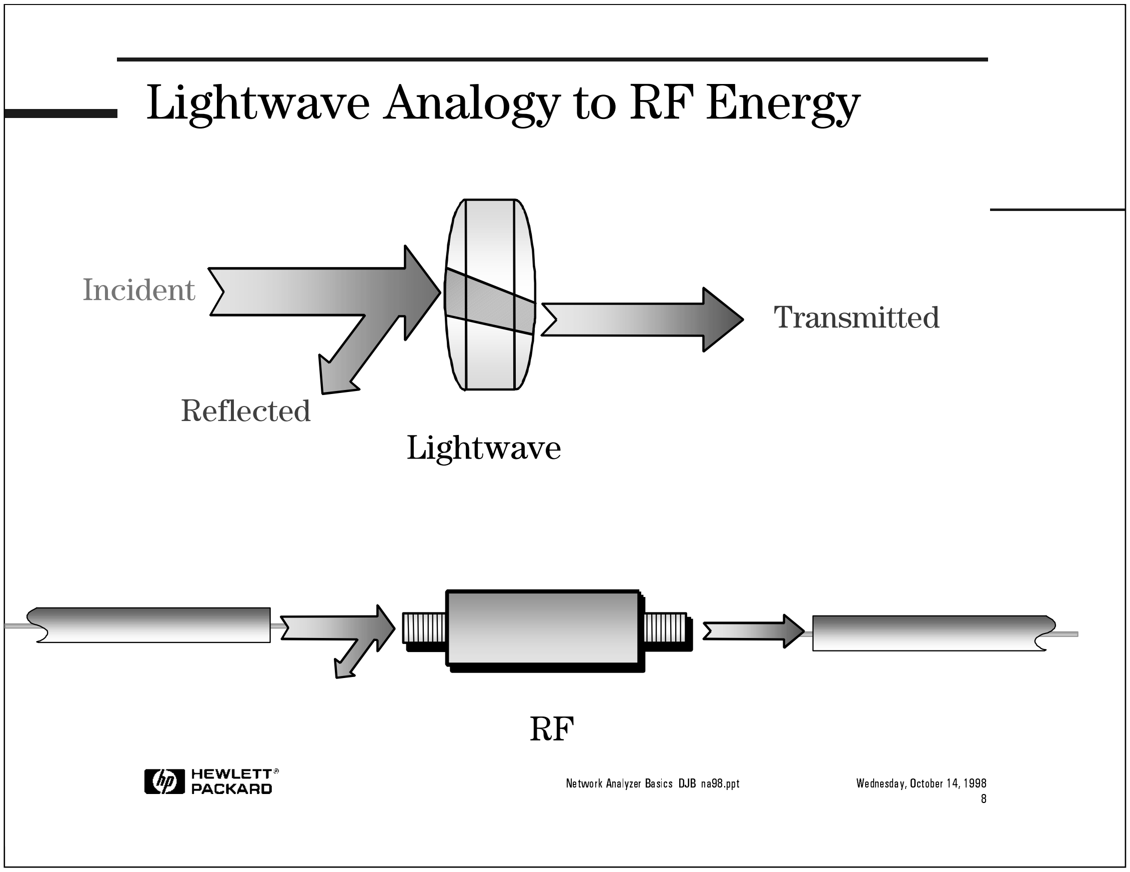
Complete characterization of devices and networks involves measurement of phase as well as magnitude. Complex impedance must be known for proper matching. Models for circuit simulation require complex data. Time-domain characterization requires magnitude and phase information to perform the inverse Fourier transform. Finally, for best measurement accuracy, we need vector error correction, as we will see later.

**Slide #7**




In this section we will review reflection and transmission measurements. We will see that transmission lines are needed to convey RF and microwave energy from one point to another with minimal loss, that transmission lines have a characteristic impedance, and that a termination at the end of a transmission line must match the characteristic impedance of the line to prevent loss of energy due to reflections. We will see how the Smith chart simplifies the process of converting reflection data to the complex impedance of the termination. For transmission measurements, we will discuss not only simple gain and loss but distortion introduced by linear devices. We will introduce S-parameters and explain why they are used instead of h-, y-, or z-parameters at RF and microwave frequencies.

**Slide #8**

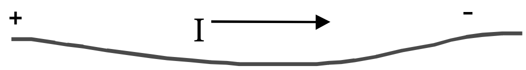



The most fundamental concept of high-frequency network analysis involves incident, reflected and transmitted waves traveling along transmission lines. It is helpful to think of traveling waves along a transmission line in terms of a lightwave analogy. We can imagine incident light striking some optical component like a clear lens. Some of the light is reflected off the surface of the lens, but most of the light continues on through the lens. If the lens had mirrored surfaces, then most of the light would be reflected and little or none would be transmitted. This concept is valid for RF signals as well.

Network analysis is concerned with the accurate measurement of the *ratio* of the reflected signal to the incident signal and/or the transmitted signal to the incident signal. The difference is that in our discussion the signal frequency is in the RF range rather than the optical range and the devices under test are electrical components or networks rather than optical.

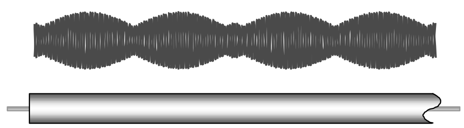
**Slide #9**

# Transmission Line Basics




***Low frequencies***

- wavelengths >> wire length
- current (I) travels down wires easily for efficient power transmission
- measured voltage and current not dependent on position along wire



***High frequencies***

- wavelength ≈ or << length of transmission medium
- need transmission lines for efficient power transmission
- matching to characteristic impedance (Z0) is very important for low reflection and maximum power transfer
- measured envelope voltage dependent on position along line



The need for efficient transfer of RF power is one of the main reasons behind the use of transmission lines. At low frequencies where the wavelength of the signals are much larger than the length of the circuit conductors, a simple wire is very useful for carrying power. Current travels down the wire easily, and voltage and current are the same no matter where we measure along the wire.

At high frequencies however, the wavelength of signals of interest are comparable to or much smaller than the length of conductors. In this case, power transmission can best be thought of in terms of traveling waves.

Of critical importance is that a lossless transmission line takes on a characteristic impedance; that is, an infinitely long line appears to be a resistive load! The geometry of the line determines the impedance. For low-power situations (cable TV, for example) coaxial transmission lines are designed to have a characteristic impedance of 75 ohms for low loss. For RF and microwave applications, where high power might be encountered, coaxial transmission lines are designed to have a characteristic impedance of 50 ohms, a compromise between maximum power handling (30 ohms) and minimum loss.

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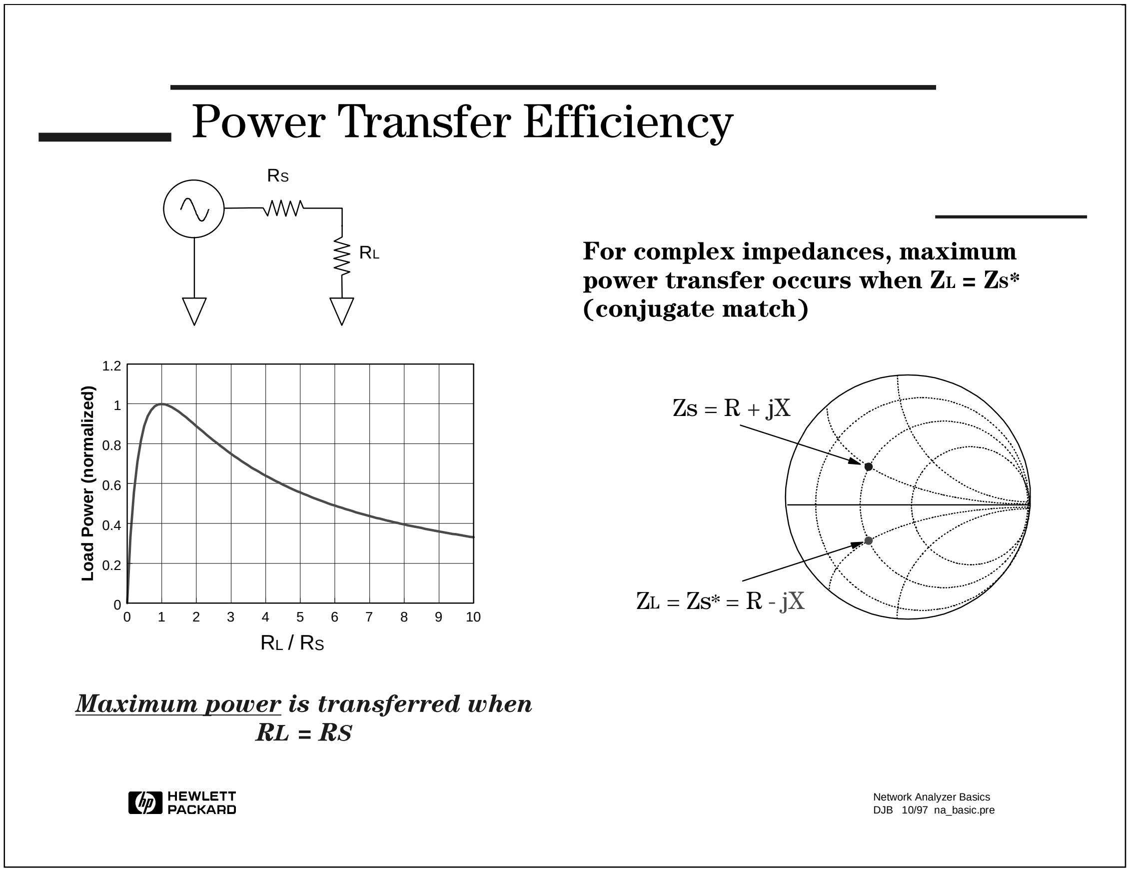


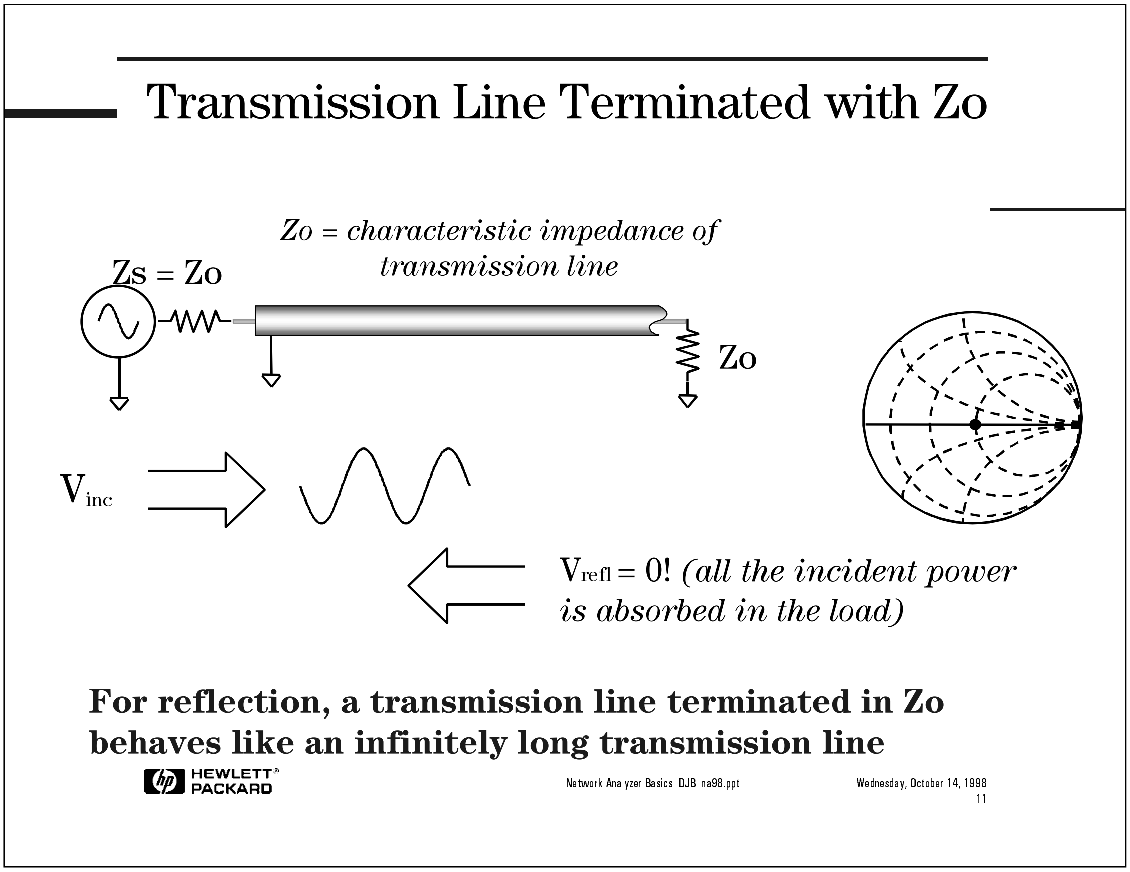

Before we begin our discussion about transmission lines, let us look at the condition for maximum power transfer into a load, given a source impedance of Rs. The graph above shows that the matched condition ($R_L = R_S$) results in the maximum power dissipated in the load resistor. This condition is true whether the stimulus is a DC voltage source or an RF sinusoid.

For maximum transfer of energy into a transmission line from a source or from a transmission line to a load (the next stage of an amplifier, an antenna, etc.), the impedance of the source and load should match the characteristic impedance of the transmission line. In general, then, Zo is the target for input and output impedances of devices and networks.

When the source impedance is not purely resistive, the maximum power transfer occurs when the load impedance is equal to the complex conjugate of the source impedance. This condition is met by reversing the sign of the imaginary part of the impedance. For example, if $R_S = 0.6 + j0.3$, then the complex conjugate $R_S^* = 0.6 - j0.3$.

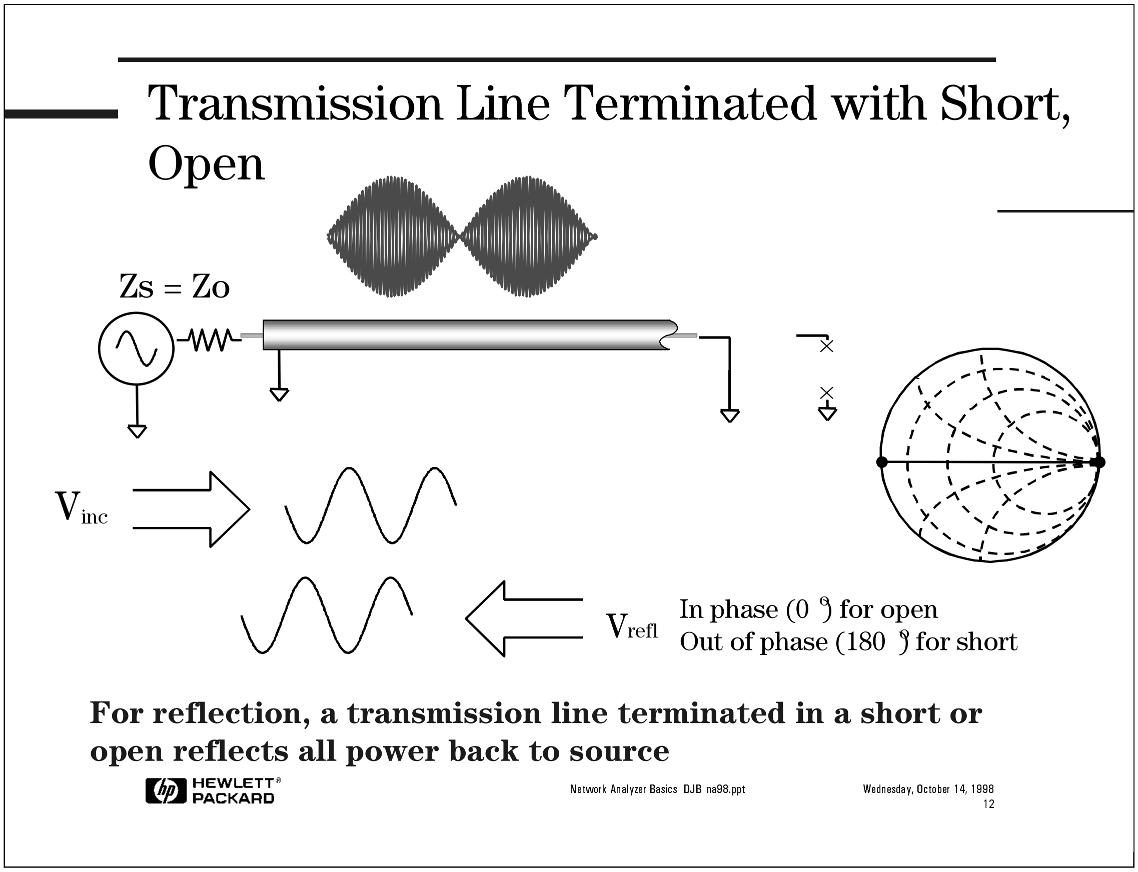
Sometimes the source impedance is adjusted to be the complex conjugate of the load impedance. For example, when matching to an antenna, the load impedance is determined by the characteristics of the antenna. A designer has to optimize the output match of the RF amplifier over the frequency range of the antenna so that maximum RF power is transmitted through the antenna.

**Slide #11**




Let's review what happens when transmission lines are terminated in different impedances, starting with a Zo load. Since a transmission line terminated in its characteristic impedance results in maximum transfer of power to the load, there is no reflected signal. This result is the same as if the transmission line was infinitely long. If we were to look at the envelope of the RF signal versus distance along the transmission line, it would be constant (no standing-wave pattern). This is because there is energy flowing in one direction only.

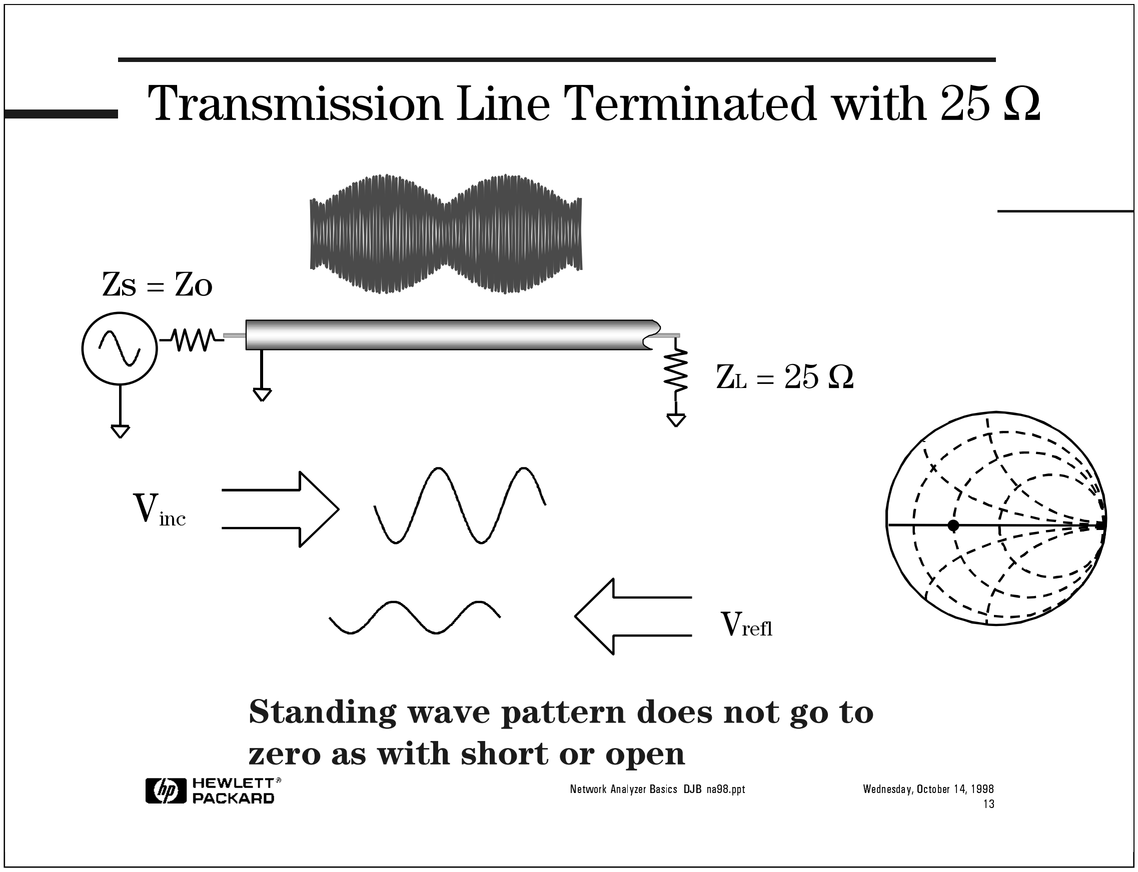
**Slide #12**




Next, let's terminate our line in a short circuit. Since purely reactive elements cannot dissipate any power, and there is nowhere else for the energy to go, a reflected wave is launched back down the line toward the source. For Ohm's law to be satisfied (no voltage across the short), this reflected wave must be equal in voltage magnitude to the incident wave, and be 180° out of phase with it. This satisfies the condition that the total voltage must equal zero at the plane of the short circuit. Our reflected and incident voltage (and current) waves will be identical in magnitude but traveling in the opposite direction.

Now let us leave our line open. This time, Ohm's law tells us that the open can support no current. Therefore, our reflected current wave must be 180° out of phase with respect to the incident wave (the voltage wave will be in phase with the incident wave). This guarantees that current at the open will be zero. Again, our reflected and incident current (and voltage) waves will be identical in magnitude, but traveling in the opposite direction. For both the short and open cases, a standing-wave pattern will be set up on the transmission line. The valleys will be at zero and the peaks at twice the incident voltage level. The peaks and valleys of the short and open will be shifted in position along the line with respect to each other, in order to satisfy Ohm's law as described above

**Slide #13**




Finally, let's terminate our line with a 25 Ω resistor (an impedance between the full reflection of an open or short circuit and the perfect termination of a 50 Ω load). Some (but not all) of our incident energy will be absorbed in the load, and some will be reflected back towards the source. We will find that our reflected voltage wave will have an amplitude 1/3 that of the incident wave, and that the two waves will be 180° out of phase at the load. The phase relationship between the incident and reflected waves will change as a function of distance along the transmission line from the load. The valleys of the standing-wave pattern will no longer go zero, and the peak will be less than that of the short/open case.

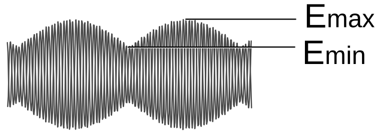
The significance of standing waves should not go unnoticed. Ohm's law tells us the complex relationship between the incident and reflected signals at the load. Assuming a 50-ohm source, the voltage across a 25-ohm load resistor will be two thirds of the voltage across a 50-ohm load. Hence, the voltage of the reflected signal is one third the voltage of the incident signal and is 180 degrees out of phase with it. However, as we move away from the load toward the source, we find that the phase between the incident and reflected signals changes! The vector sum of the two signals therefore also changes along the line, producing the standing wave pattern. The apparent impedance also changes along the line because the relative amplitude and phase of the incident and reflected waves *at any given point* uniquely determine the measured impedance. For example, if we made a measurement one quarter wavelength away from the 25-ohm load, the results would indicate a 100-ohm load. The standing wave pattern repeats every half wavelength, so the apparent impedance does as well.

**Slide #14**

## Reflection Parameters

**Reflection Coefficient** $\Gamma = \dfrac{V_{reflected}}{V_{incident}} = \rho \angle \Phi = \dfrac{Z_L - Z_O}{Z_L + Z_O}$

**Return loss** = -20 log($\rho$), $\rho = |\Gamma|$

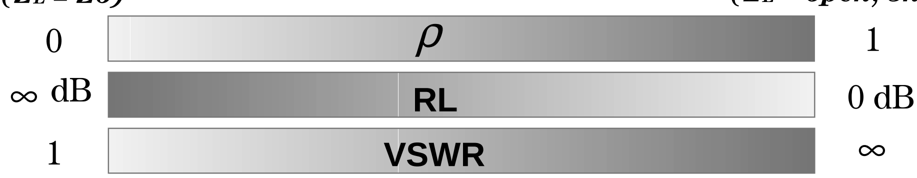



***Voltage Standing Wave Ratio***

**VSWR** $= \dfrac{E_{max}}{E_{min}} = \dfrac{1 + \rho}{1 - \rho}$

***No reflection***
***($Z_L$ = Zo)***

***Full reflection***
***($Z_L$ = open, short)***

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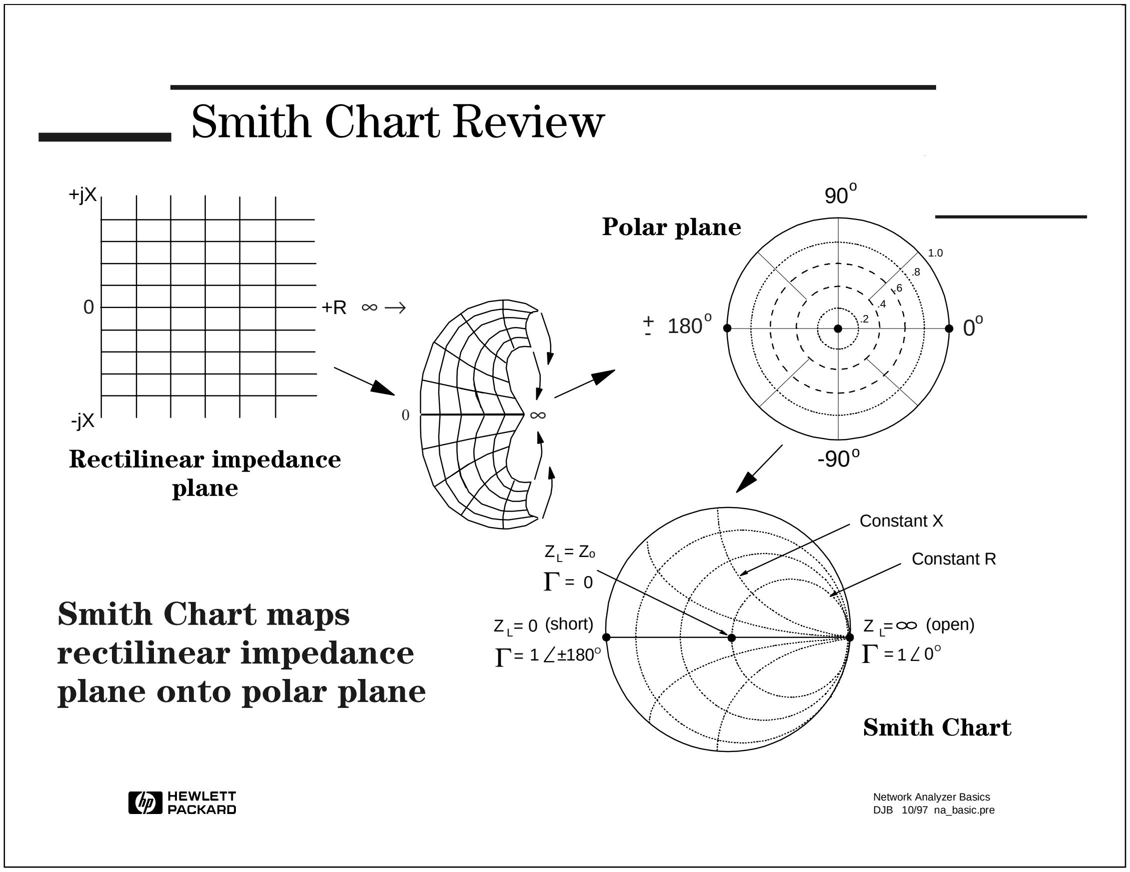
The first term for reflected waves is reflection coefficient gamma ($\Gamma$). The magnitude portion of gamma is called rho ($\rho$). Reflection coefficient is the ratio of the reflected signal voltage to the incident signal voltage. For example, a transmission line terminated in $Z_o$ will have all energy transferred to the load; hence $V_{refl}$ = 0 and $\rho$ = 0. When $Z_L$ is not equal to $Z_o$ , some energy is reflected and $\rho$ is greater than zero. When $Z_L$ = a short or open circuit, all energy is reflected and $\rho$ = 1. The range of possible values for $\rho$ is then zero to one.

Since it is often very convenient to show reflection on a logarithmic display, the second way to convey reflection is return loss. Return loss is expressed in terms of dB, and is a scalar quantity. The definition for return loss includes a negative sign so that the return loss value is always a positive number (when measuring reflection on a network analyzer with a log magnitude format, ignoring the minus sign gives the results in terms of return loss). Return loss can be thought of as the number of dB that the reflected signal is below the incident signal. Return loss varies between infinity for a $Z_o$ impedance and 0 dB for an open or short circuit.

As we have already seen, two waves traveling in opposite directions on the same media cause a "standing wave". This condition can be measured in terms of the voltage standing wave ratio (VSWR or SWR for short), and is defined as the maximum value of the RF envelope over the minimum value of the envelope. This value can be computed as (1+$\rho$)/(1-$\rho$). VSWR can take on values between one and infinity.

**Slide #15**




Our network analyzer gives us complex reflection coefficient. However, we typically want to know the impedance of the device under test. The previous slide shows the relationship between reflection coefficient and impedance, and we could perform the complex math to find the impedance. Although programmable calculators and computers take the drudgery out of doing the math, a single number does not always give us the complete picture. In addition, impedance almost certainly changes with frequency, so even if we did all the math, we would end up with a table of numbers that may be difficult to interpret.

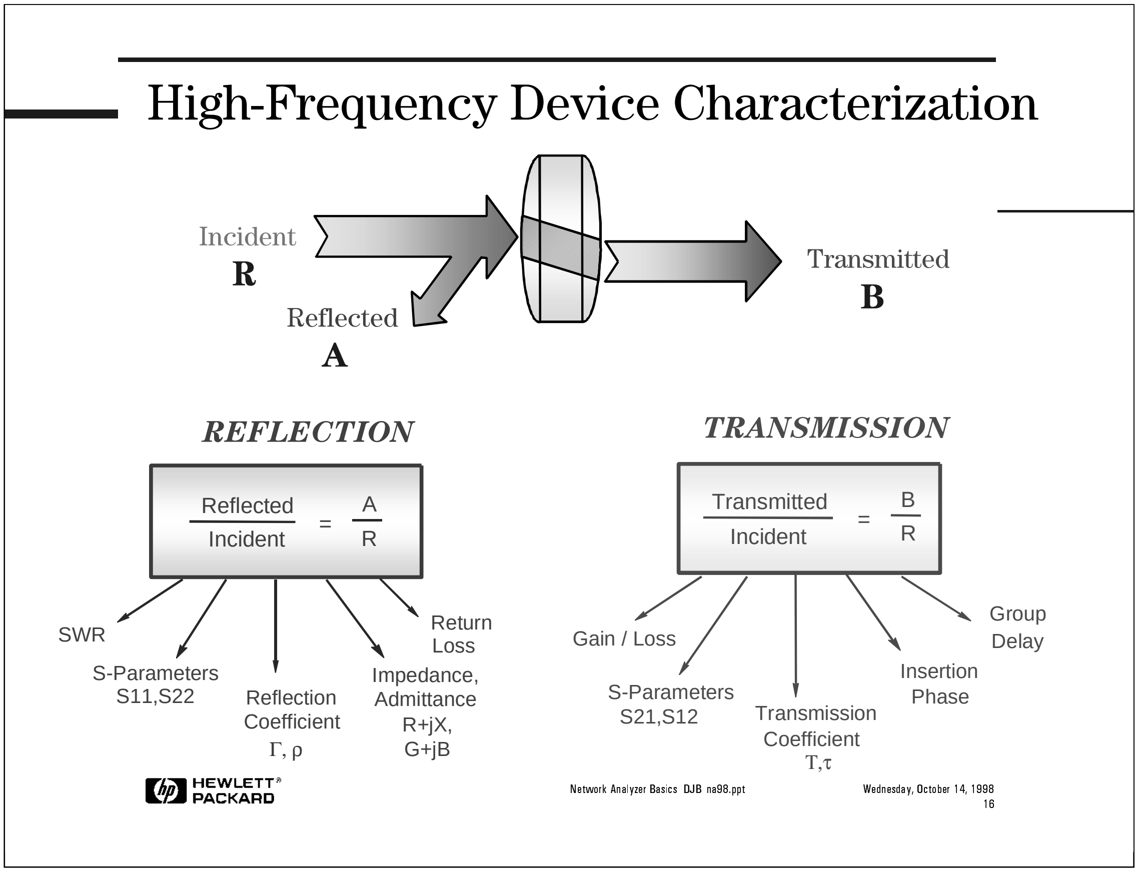
A simple, graphical method may be what we want. Let's first plot reflection coefficient. For positive resistance, the absolute magnitude of $\Gamma$ varies from zero (perfect load) to unity (full reflection) at some angle. So we have a unit circle, the polar plane shown here. An open would plot at 1 $\underline{/0}^{o}$ ; a short, at 1 $\underline{/180}^{o}$; a perfect load, at the center, and so on. How do we get from the polar data to impedance graphically? If we try to map the polar plane onto the rectilinear impedance plane, we find that we have problems. First of all, the rectilinear plane does not have values to infinity. Second, circles of constant reflection coefficient are concentric on the polar plane but not on the rectilinear plane, making it difficult to make judgments regarding two different impedances. Finally, phase angles plot as radii on the polar plane but plot as arcs on the rectilinear plane, making it difficult to pinpoint $\rho\angle\Theta$.

Plotting the polar plane onto the impedance plane is not the answer. Instead, in the 1930's, Phillip H. Smith mapped the impedance plane onto the polar plane, creating the chart that bears his name, the Smith Chart. Since unity at zero degrees on the polar plane represents infinite impedance, both plus and minus infinite reactances, as well as infinite resistance, plot to this point. The result is that the vertical lines on the rectilinear plane that indicate values of constant resistance map to circles on the polar plane; the horizontal lines that indicate values of constant reactance map to arcs on the polar plane. Zo maps to the exact center of the chart.

To use the chart, we can take the value of $\rho$ from our network analyzer and draw a circle on the Smith Chart. We then find the angle from our network analyzer and plot the appropriate radius on the chart. Finally, we read the impedance at the point of intersection of the circle and radius. In general, Smith Charts are normalized to Zo; that is, the impedance values are divided by Zo. The chart is then independent of the characteristic impedance of the system in question. Actual impedance values are derived by multiplying the indicated value by Zo. For example, in a 50-ohm system, a normalized value of 0.3 - j0.15 becomes 15 - j7.5 ohms; in a 75-ohm system, 22.5 - j11.25 ohms.

Fortunately, we no longer have to go through the exercise ourselves. Out network analyzer can display the Smith Chart, plot measured data on it, and provide adjustable markers and calculated impedance at the marked point.

**Slide #16**




Now that we fully understand the relationship of electromagnetic waves, we must also recognize the terms used to describe them. We will show how to measure the incident, reflected and the transmitted waves. Common network analyzer terminology has the incident wave measured with the R (for reference) channel. The reflected wave is measured with the A channel and the transmitted wave is measured with the B channel. With amplitude and phase information of these three waves, we can quantify the reflection and transmission characteristics of our device under test (DUT). Some of the common measured terms are scalar in nature (the phase part is ignored or not measured), while others are vector (both magnitude and phase are measured). For example, return loss is a scalar measurement of reflection, while impedance is a vector reflection measurement. Ratioed reflection is often shown as A/R and ratioed transmission is often shown as B/R, relating to the measurement channels used in the network analyzer.

**Slide #17**

# Transmission Parameters


*Transmission Coefficient* = $T = \dfrac{V_{\text{Transmitted}}}{V_{\text{Incident}}} = \tau \angle \phi$

*Insertion Loss (dB)* = $-20 \text{ Log} \left| \dfrac{V_{\text{Trans}}}{V_{\text{Inc}}} \right| = -20 \log \tau$

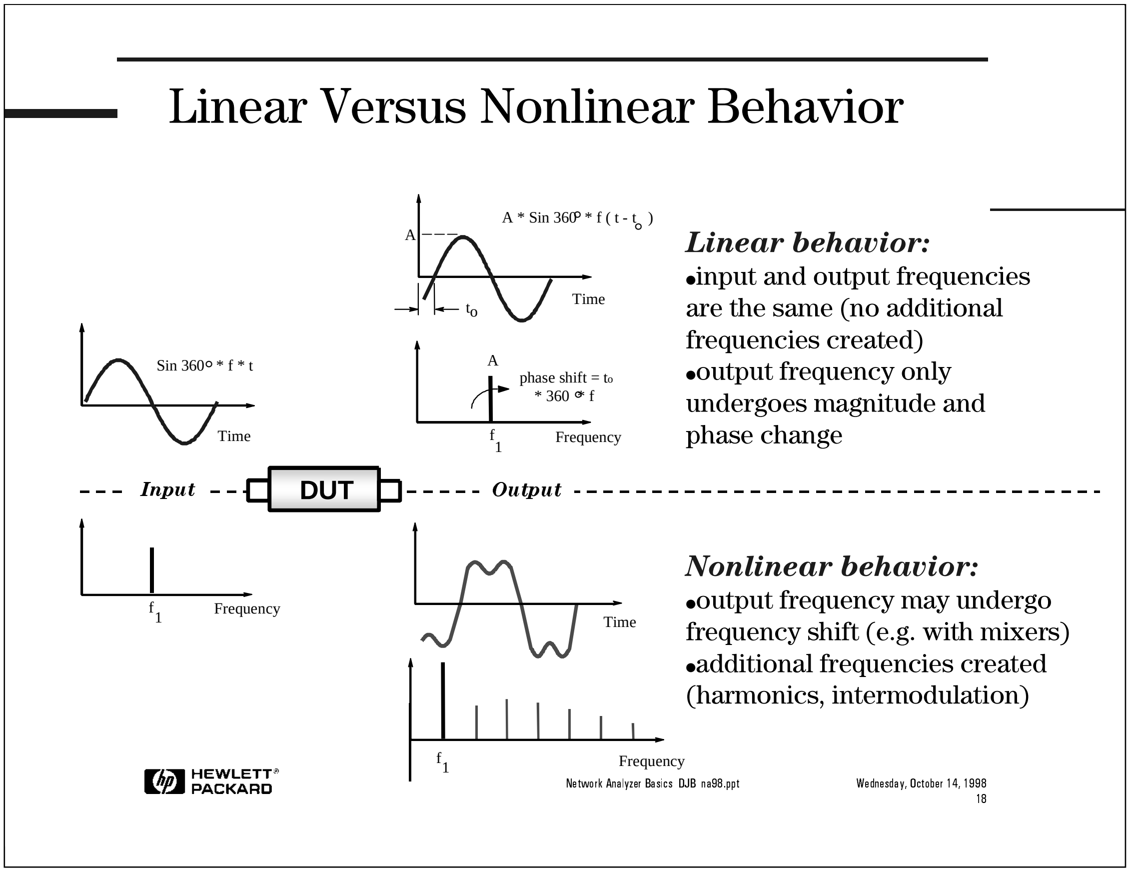
*Gain (dB)* = $20 \text{ Log} \left| \dfrac{V_{\text{Trans}}}{V_{\text{Inc}}} \right| = 20 \log \tau$

Network Analyzer Basics DJB na98.ppt    Wednesday, October 14, 1998
17

Transmission coefficient T is defined as the transmitted voltage divided by the incident voltage. If $|V_{trans}| > |V_{inc}|$, we have gain, and if $|V_{trans}| < |V_{inc}|$, we have attenuation or insertion loss. When insertion loss is expressed in dB, a negative sign is added in the definition so that the loss value is expressed as a positive number. The phase portion of the transmission coefficient is called insertion phase.

There is more to transmission than simple gain or loss. Most signals are time varying; that is, they occupy a given bandwidth and are made up of multiple components. It might be important, then, to know to what extent the device under test alters the makeup of the signal and thereby distorts the signal.

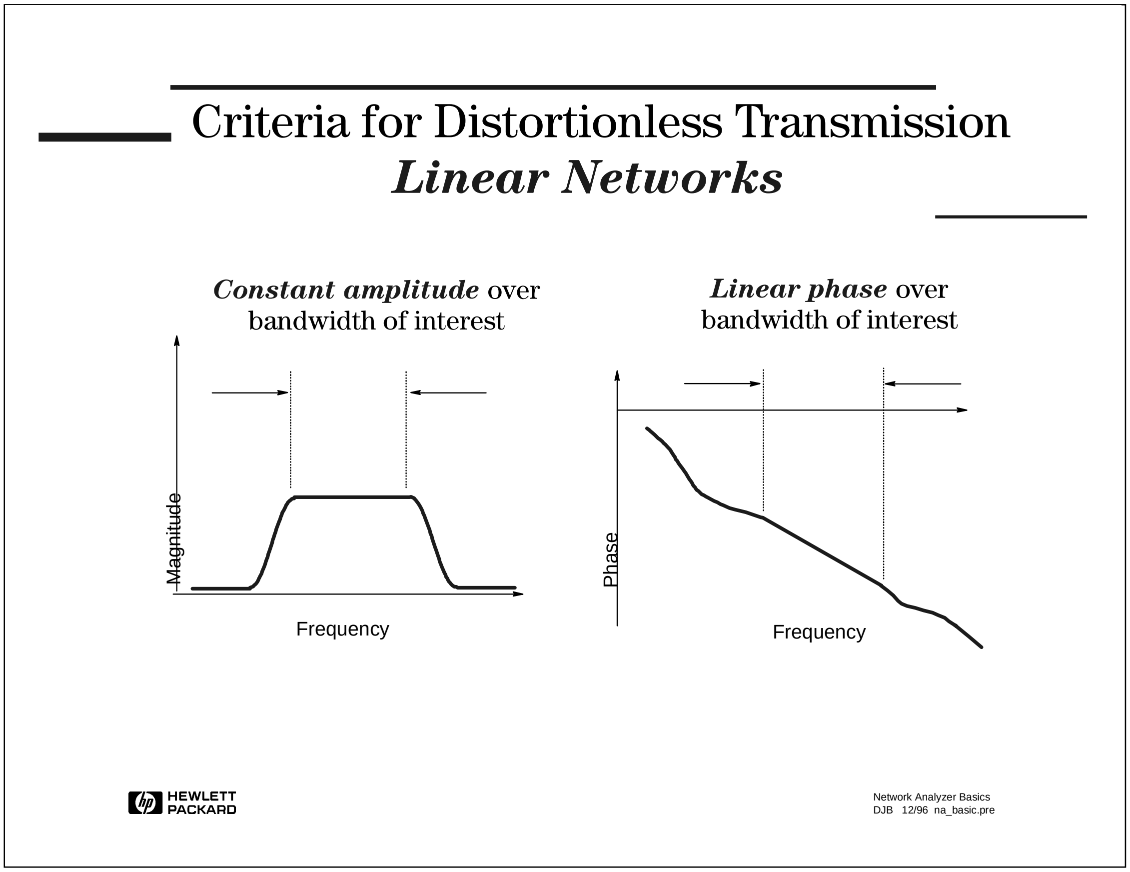
**Slide #18**




Before we explore the different types of signal distortion that can occur, lets review the differences between linear and nonlinear behavior. Devices that behave linearly only impose magnitude and phase changes on input signals. Any sinusoid appearing at the input will also appear at the output at the same frequency. No new signals are created.

Non-linear devices can shift input signals in frequency (a mixer for example) and/or create new signals in the form of harmonics or intermodulation products. Many components that behave linearly under most signal conditions can exhibit nonlinear behavior if driven with a large enough input signal. This is true for both passive devices like filters and active devices like amplifiers.

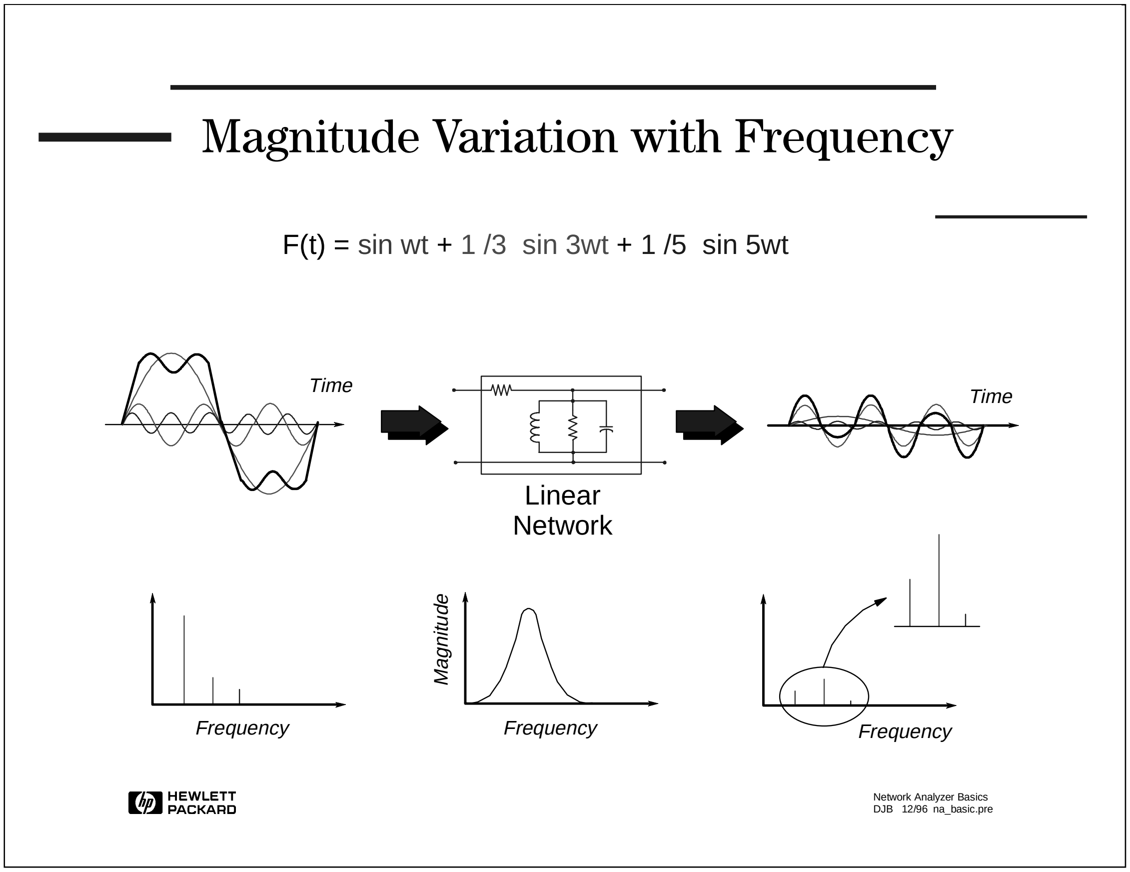
**Slide #19**




Now lets examine how linear networks can cause signal distortion. There are two criteria that must be satisfied for linear *distortionless* transmission. First, the amplitude (magnitude) response of the device or system must be flat over the bandwidth of interest. This means all frequencies within the bandwidth will be attenuated identically. Second, the phase response must be linear over the bandwidth of interest.

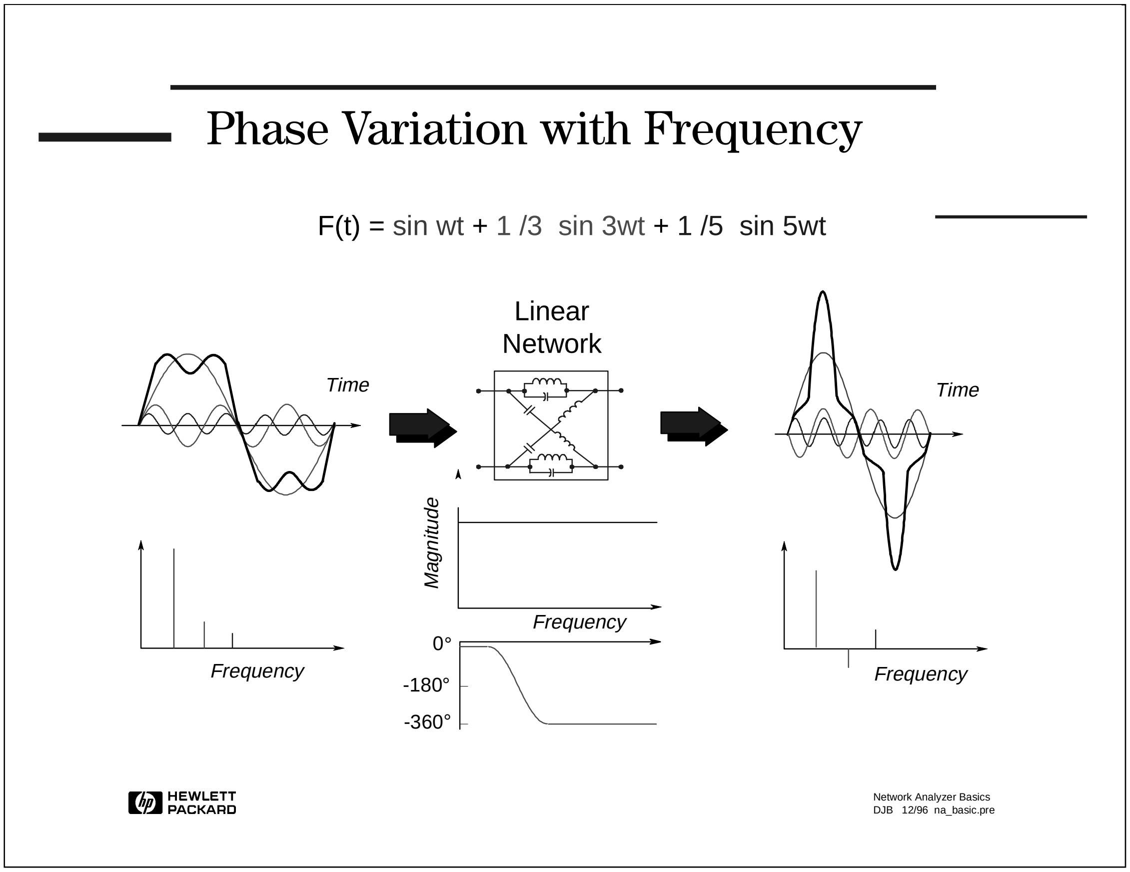
How can magnitude and phase distortion occur? The following two examples will illustrate how both magnitude and phase responses can introduce linear distortion onto signals.

**Slide #20**




Here is an example of a square wave (consisting of three sinusoids) applied to a bandpass filter. The filter imposes a non-uniform amplitude change to each frequency component. Even though no phase changes are introduced, the frequency components no longer sum to a square wave at the output. The square wave is now severely distorted, having become more sinusoidal in nature.

**Slide #21**

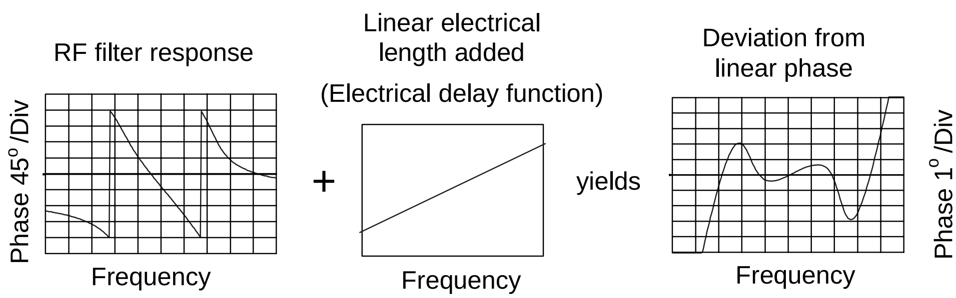



Let's apply the same square wave to another filter. Here, the third harmonic undergoes a 180° phase change, but the other components are not shifted in phase. All the amplitudes of the three spectral components remain the same (filters which only affect the phase of signals are called allpass filters). The output is again distorted, appearing very impulsive this time.

**Slide #22**

# Deviation from Linear Phase

***Use electrical delay to remove linear portion of phase response***



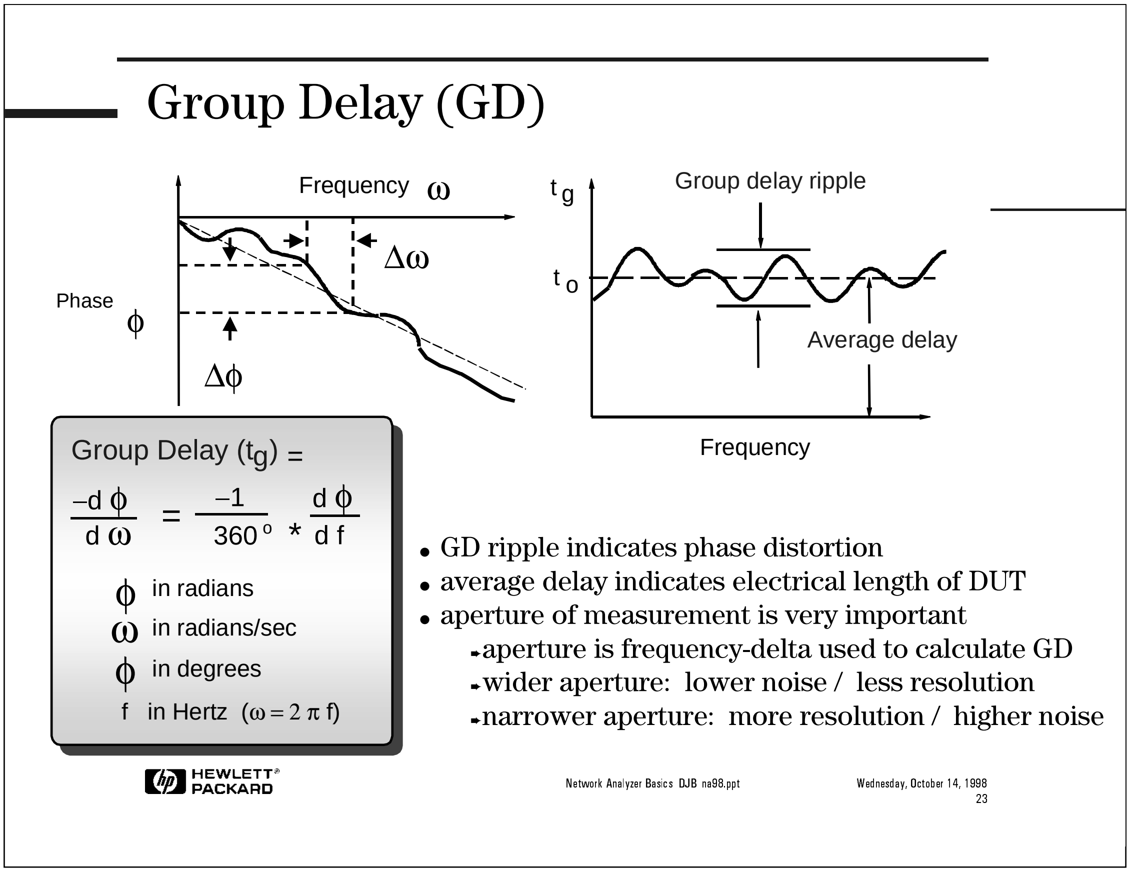

Low resolution

High resolution



Looking at insertion phase directly is usually not very useful. This is because the phase has a negative slope with respect to frequency due to the electrical length of the device (the longer the device, the greater the slope). Since it is only the deviation from linear phase which causes distortion, it is desirable to remove the linear portion of the phase response. This can be accomplished by using the electrical delay feature of the network analyzer to cancel the electrical length of the DUT. This results in a high-resolution display of phase distortion (deviation from linear phase).

**Slide #23**

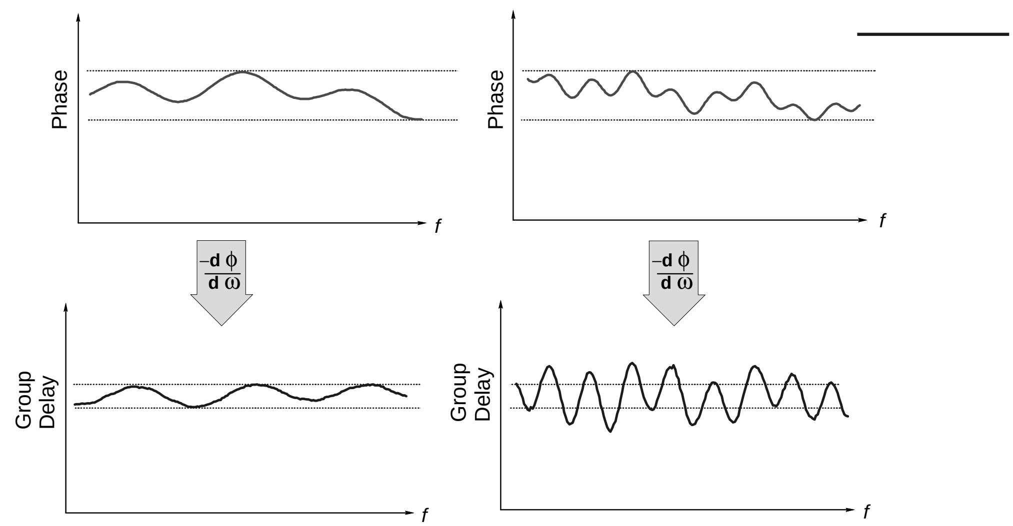



Another useful measure of phase distortion is group delay. Group delay is a measure of the transit time of a signal through the device under test, versus frequency. Group delay is calculated by differentiating the insertion-phase response of the DUT versus frequency. Another way to say this is that group delay is a measure of the slope of the transmission phase response. The linear portion of the phase response is converted to a constant value (representing the average signal-transit time) and deviations from linear phase are transformed into deviations from constant group delay. The variations in group delay cause signal distortion, just as deviations from linear phase cause distortion. Group delay is just another way to look at linear phase distortion.

When specifying or measuring group delay, it is important to quantify the aperture in which the measurement is made. The aperture is defined as the frequency delta used in the differentiation process (the denominator in the group-delay formula). As we widen the aperture, trace noise is reduced but less group-delay resolution is available (we are essentially averaging the phase response over a wider window). As we make the aperture more narrow, trace noise increases but we have more measurement resolution.

**Slide #24**

## Why Measure Group Delay?




**Same p-p phase ripple can result in different group delay**

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Why are both deviation from linear phase and group delay commonly measured? Depending on the device, both may be important. Specifying a maximum peak-to-peak value of phase ripple is not sufficient to completely characterize a device since the slope of the phase ripple is dependent on the number of ripples which occur per unit of frequency. Group delay takes this into account since it is the differentiated phase response. Group delay is often a more easily interpreted indication of phase distortion.

The plot above shows that the same value of peak-to-peak phase ripple can result in substantially different group delay responses. The response on the right with the larger group-delay variation would cause more signal distortion.

**Slide #25**

## Characterizing Unknown Devices

***Using parameters (H, Y, Z, S) to characterize devices:***

- gives us a linear behavioral model of our device
- measure parameters (e.g. voltage and current) versus frequency under various source and load conditions (e.g. short and open circuits)
- compute device parameters from measured data
- now we can predict circuit performance under any source and load conditions

| ***H-parameters*** | ***Y-parameters*** | ***Z-parameters*** |
|---|---|---|
| $V_1 = h_{11}I_1 + h_{12}V_2$ | $I_1 = y_{11}V_1 + y_{12}V_2$ | $V_1 = z_{11}I_1 + z_{12}I_2$ |
| $I_2 = h_{21}I_1 + h_{22}V_2$ | $I_2 = y_{21}V_1 + y_{22}V_2$ | $V_2 = z_{21}I_1 + z_{22}I_2$ |

$$h_{11} = \left.\frac{V_1}{I_1}\right|_{V_2=0}$$ *(requires **short** circuit)*

$$h_{12} = \left.\frac{V_1}{V_2}\right|_{I_1=0}$$ *(requires **open** circuit)*

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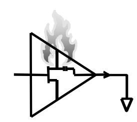
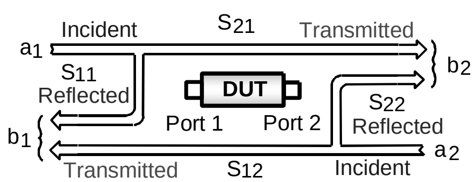
In order to completely characterize an unknown linear two-port device, we must make measurements under various conditions and compute a set of parameters. These parameters can be used to completely describe the electrical behavior of our device (or network), even under source and load conditions other than when we made our measurements. For low-frequency characterization of devices, the three most commonly measured parameters are the H, Y and Z-parameters. All of these parameters require measuring the total voltage or current as a function of frequency at the input or output nodes (ports) of the device. Furthermore, we have to apply either open or short circuits as part of the measurement. Extending measurements of these parameters to high frequencies is not very practical.

**Slide #26**

# Why Use S-Parameters?

- relatively easy to **obtain** at high frequencies
  - measure voltage traveling waves with a vector network analyzer
  - don't need shorts/opens which can cause active devices to oscillate or self-destruct
- relate to **familiar** measurements (gain, loss, reflection coefficient ...)
- can **cascade** S-parameters of multiple devices to predict system performance
- can **compute** H, Y, or Z parameters from S-parameters if desired
- can easily import and use S-parameter files in our **electronic-simulation** tools






$$b_1 = S_{11} a_1 + S_{12} a_2$$
$$b_2 = S_{21} a_1 + S_{22} a_2$$

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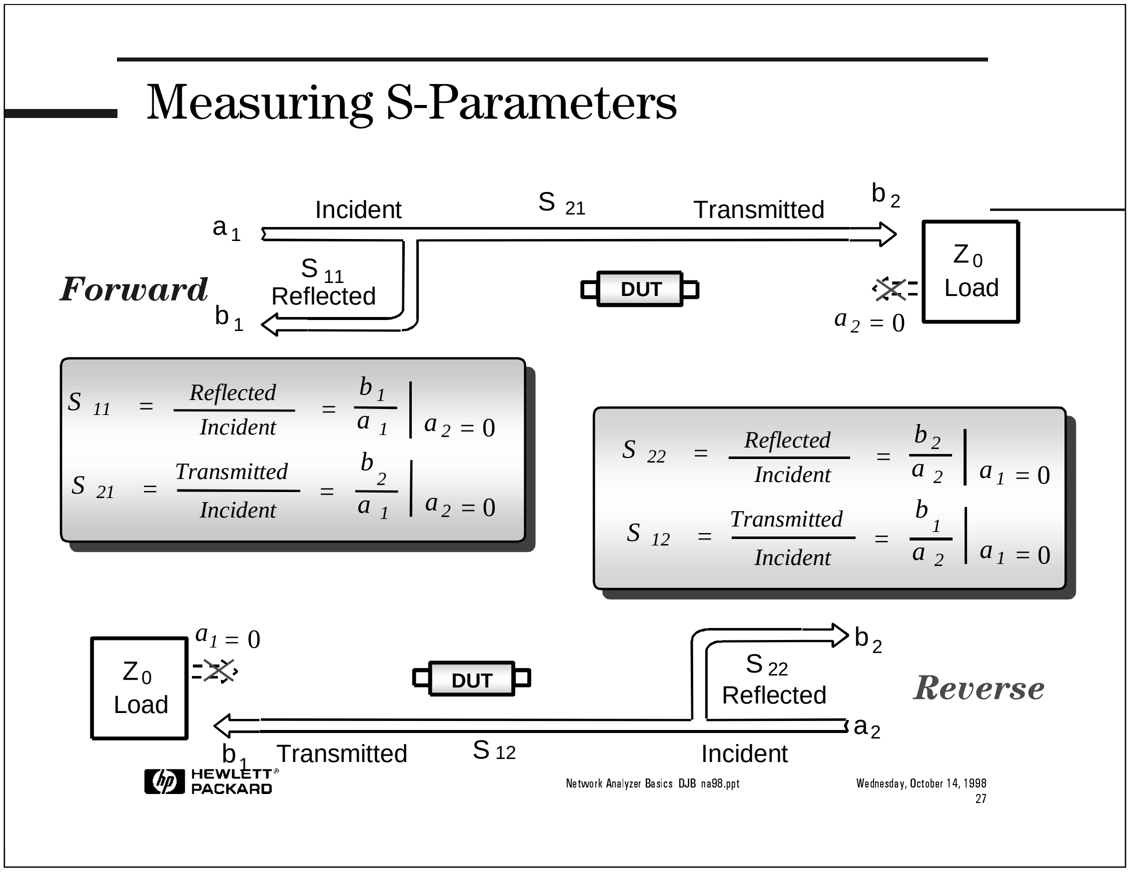
26

At high frequencies, it is very hard to measure total voltage and current at the device ports. One cannot simply connect a voltmeter or current probe and get accurate measurements due to the impedance of the probes themselves and the difficulty of placing the probes at the desired positions. In addition, active devices may oscillate or self-destruct with the connection of shorts and opens.

Clearly, some other way of characterizing high-frequency networks is needed that doesn't have these drawbacks. That is why scattering or S-parameters were developed. S-parameters have many advantages over the previously mentioned H, Y or Z-parameters. They relate to familiar measurements such as gain, loss, and reflection coefficient. They are defined in terms of voltage traveling waves, which are relatively easy to measure. S-parameters don't require connection of undesirable loads to the device under test. The measured S-parameters of multiple devices can be cascaded to predict overall system performance. If desired, H, Y, or Z-parameters can be derived from S-parameters. And very important for RF design, S-parameters are easily imported and used with electronic-simulation tools. S-parameters are the shared language between simulation and measurement.

An N-port DUT has $N^2$ S-parameters. So, a two-port device has four S-parameters. The numbering convention for S-parameters is that the first number following the "S" is the port where the signal emerges, and the second number is the port where the signal is applied. So, $S_{21}$ is a measure of the signal coming out port 2 relative to the RF stimulus entering port 1. When the numbers are the same (e.g., $S_{11}$), it indicates a reflection measurement, as the input and output ports are the same. The incident terms (a1, a2) and output terms (b1, b2) are expressed in terms of voltage traveling waves.

**Slide #27**




S11 and S21 are determined by measuring the magnitude and phase of the incident, reflected and transmitted voltage signals when the output is terminated in a perfect Zo (a load that equals the characteristic impedance of the test system). This condition guarantees that $a_2$ is zero, since there is no reflection from an ideal load. S11 is equivalent to the input complex reflection coefficient or impedance of the DUT, and S21 is the forward complex transmission coefficient. Likewise, by placing the source at port 2 and terminating port 1 in a perfect load (making $a_1$ zero), S22 and S12 measurements can be made. S22 is equivalent to the output complex reflection coefficient or output impedance of the DUT, and S12 is the reverse complex transmission coefficient.

The accuracy of S-parameter measurements depends greatly on how good a termination we apply to the port not being stimulated. Anything other than a perfect load will result in $a_1$ or $a_2$ not being zero (which violates the definition for S-parameters). When the DUT is connected to the test ports of a network analyzer and we don't account for imperfect test port match, we have not done a very good job satisfying the condition of a perfect termination. For this reason, two-port error correction, which corrects for source and load match, is very important for accurate S-parameter measurements (two-port correction is covered in the calibration section).

**Slide #28**

## Equating S-Parameters with Common Measurement Terms

$S_{11}$ = forward reflection coefficient ***(input match)***

$S_{22}$ = reverse reflection coefficient ***(output match)***

$S_{21}$ = forward transmission coefficient ***(gain or loss)***

$S_{12}$ = reverse transmission coefficient ***(isolation)***

***Remember, S-parameters are inherently linear quantities -- however, we often express them in a log-magnitude format***

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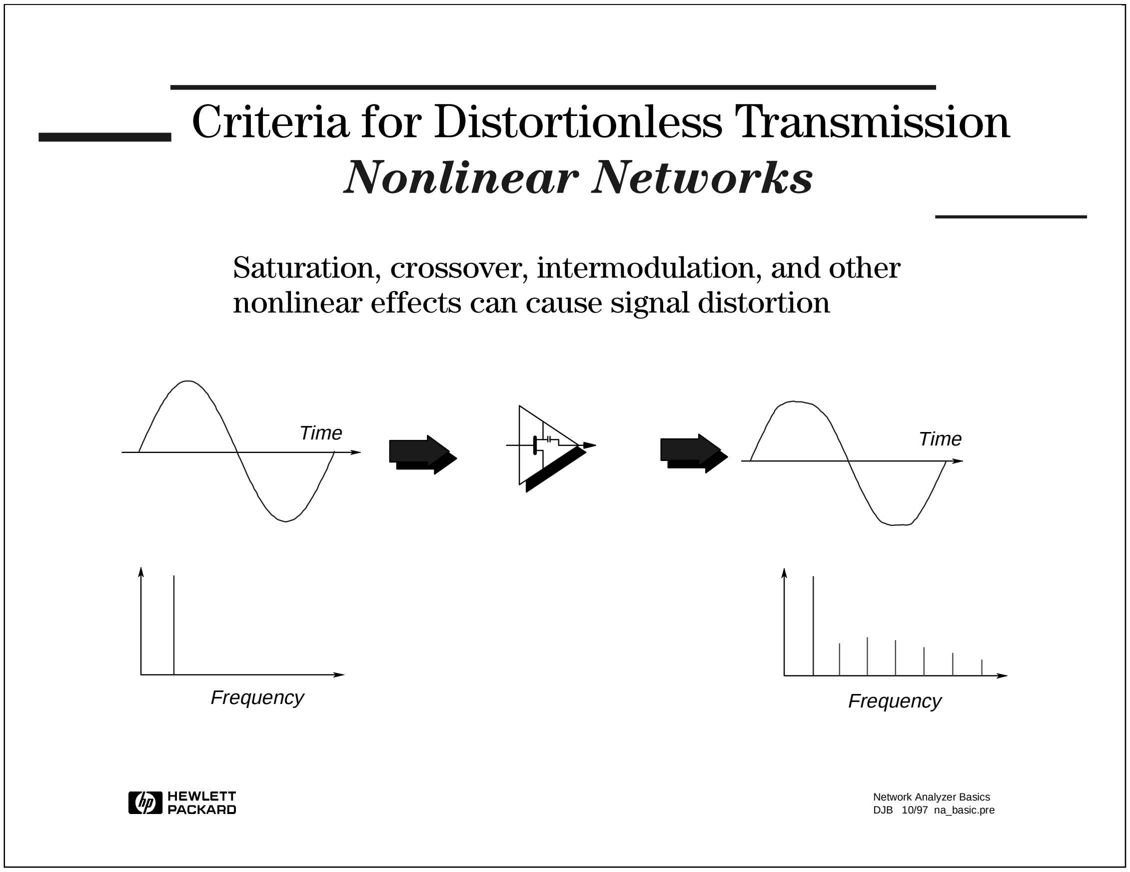
S-parameters are essentially the same parameters as some of the terms we have mentioned before, as described above. Remember, S-parameters are inherently linear quantities – however, we often express them in a log-magnitude format. S11 and S22 are often displayed on a Smith chart.

**Slide #29**




We have just seen how linear networks can cause distortion.  Devices which behave nonlinearly also introduce distortion. The example above shows an amplifier that is overdriven, causing the signal at the output to "clip" due to saturation in the amplifier. Because the output signal is no longer a pure sinusoid, harmonics are present at integer multiples of the input frequency.

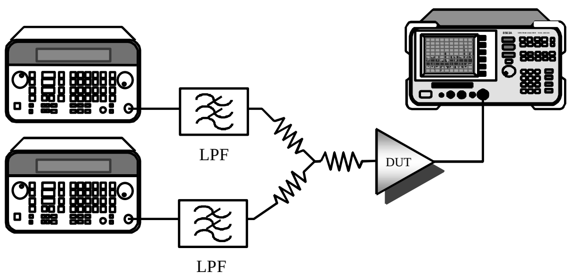
Passive devices can also exhibit nonlinear behavior at high power levels.  A common example is an L-C filter that uses inductors made with magnetic cores.  Magnetic materials often display hysteresis effects, which are highly nonlinear

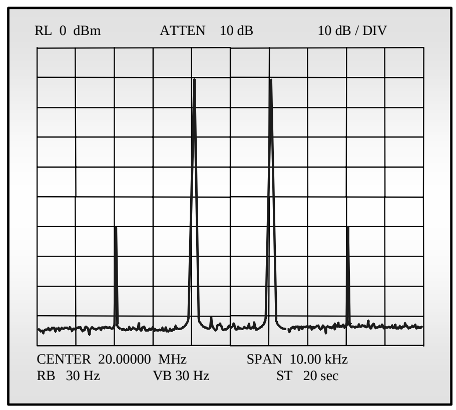
**Slide #30**

## Measuring Nonlinear Behavior

Most common measurements:

- using a ***spectrum analyzer*** + source(s)
  - harmonics, particularly second and third
  - intermodulation products resulting from two or more RF carriers
- using a ***network analyzer*** and power sweeps
  - gain compression
  - AM to PM conversion

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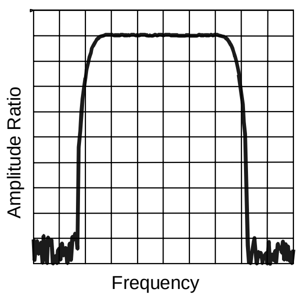
So far, we've focused our attention on linear swept-frequency characterization (used for passive and active devices). Other important characterizations for active devices are nonlinear behavior such as intermodulation distortion and gain compression.

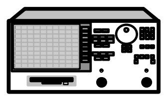
Nonlinear behavior is important to quantify as it can cause severe signal distortion. The most common nonlinear measurements are harmonic and intermodulation distortion (usually measured with spectrum analyzers and signal sources) and gain compression and AM-to-PM conversion (usually measured with network analyzers and power sweeps).

We will cover swept-power measurements using a network analyzer in the typical-measurements section of this presentation.

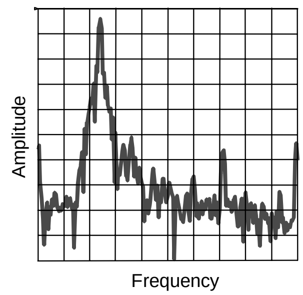
**Slide #31**

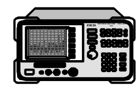
# What is the Difference Between ***Network*** and ***Spectrum*** Analyzers?



Measures known signal



Measures unknown signals

**Network analyzers:**
- measure components, devices, circuits, sub-assemblies
- contain source and receiver
- display ratioed amplitude and phase (frequency or power sweeps)
- offer advanced error correction

**Spectrum analyzers:**
- measure signal amplitude characteristics carrier level, sidebands, harmonics...)
- can demodulate (& measure) complex signals
- are receivers only (single channel)
- can be used for scalar component test (*no phase*) with tracking gen. or ext. source(s)

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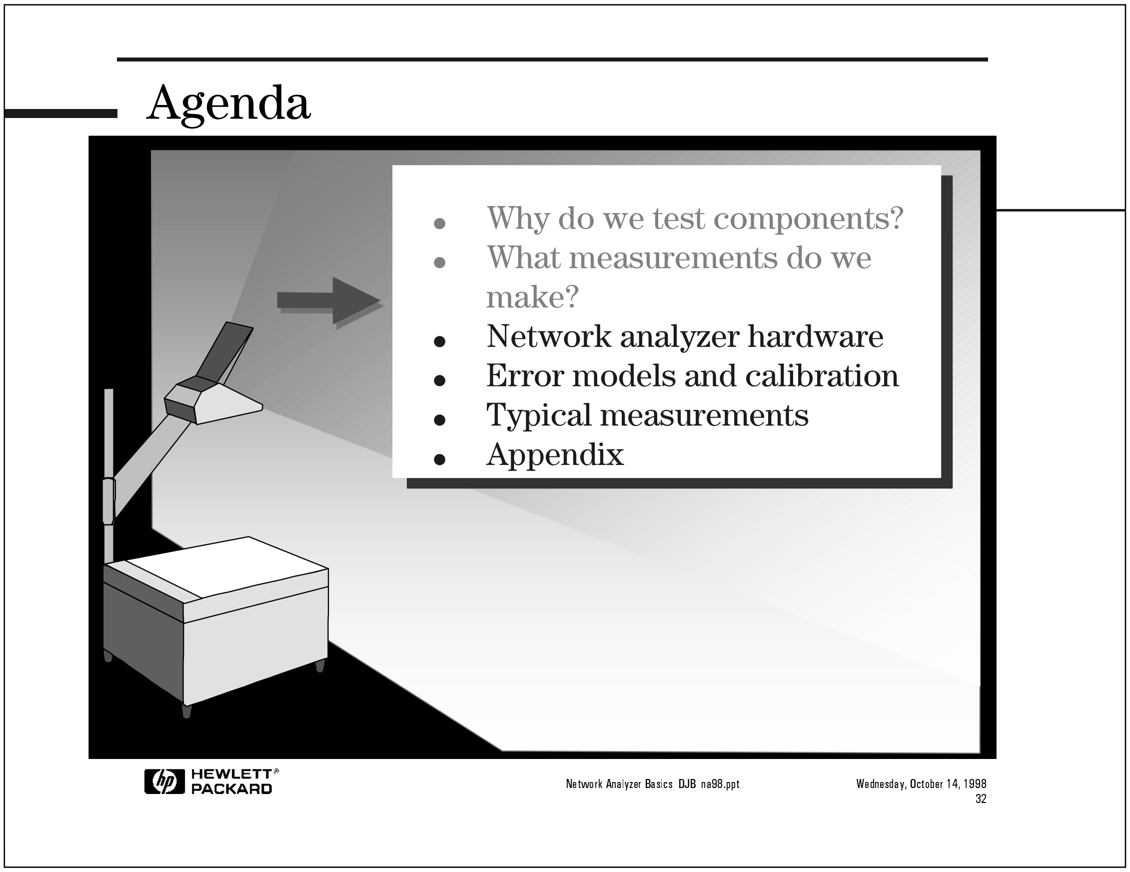
Now that we have seen some of the measurements that are commonly done with network and spectrum analyzers, it might be helpful to review the main differences between these instruments. Network analyzers are used to measure components, devices, circuits, and sub-assemblies. They contain both a source and multiple receivers, and generally display *ratioed* amplitude and phase information (frequency or power sweeps). A network analyzer is always looking at a *known* signal (in terms of frequency), since it is a stimulus-response system. With network analyzers, it is harder to get an (accurate) trace on the display, but very easy to interpret the results. With vector-error correction, network analyzers provide much higher measurement accuracy than spectrum analyzers.

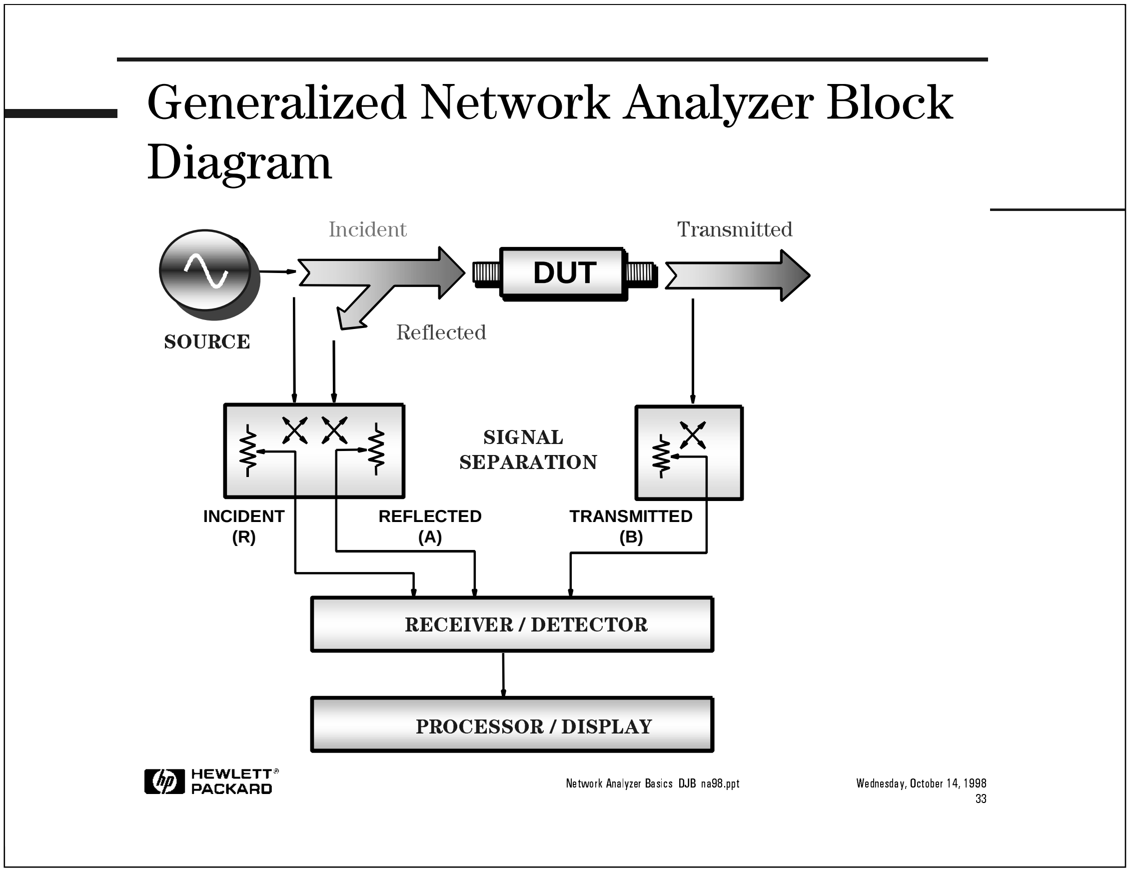
Spectrum analyzers are most often used to measure signal characteristics such as carrier level, sidebands, harmonics, phase noise, etc., on *unknown* signals. They are most commonly configured as a single-channel receiver, without a source. Because of the flexibility needed to analyze signals, spectrum analyzers generally have a much wider range of IF bandwidths available than most network analyzers. Spectrum analyzers are often used with external sources for nonlinear stimulus/response testing. When combined with a tracking generator, spectrum analyzers can be used for scalar component testing (magnitude versus frequency, but no phase measurements). With spectrum analyzers, it is easy to get a trace on the display, but interpreting the results can be much more difficult than with a network analyzer.

**Slide #32**




In this next section, we will look at the main parts of a network analyzer. More detailed information of the hardware inside the network analyzer is available in the Appendix section of this paper.

**Slide #33**




Here is a generalized block diagram of a network analyzer, showing the major signal processing sections. In order to measure the incident, reflected and transmitted signal, four sections are required:

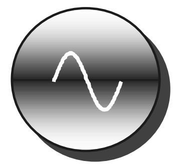
1. Source for stimulus
2. Signal-separation devices
3. Receiver that provides detection
4. Processor/display for calculating and reviewing the results

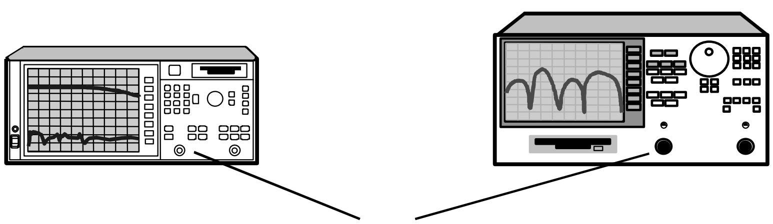
We will briefly examine the source, receiver, and processor sections. More detailed information about the signal separation devices and receiver section are in the Appendix.

**Slide #34**

# Source

- Supplies stimulus for system
- Swept frequency or power
- Traditionally NAs used separate source
  - Open-loop VCOs
  - Synthesized sweepers
- Most HP analyzers sold today have integrated, synthesized sources



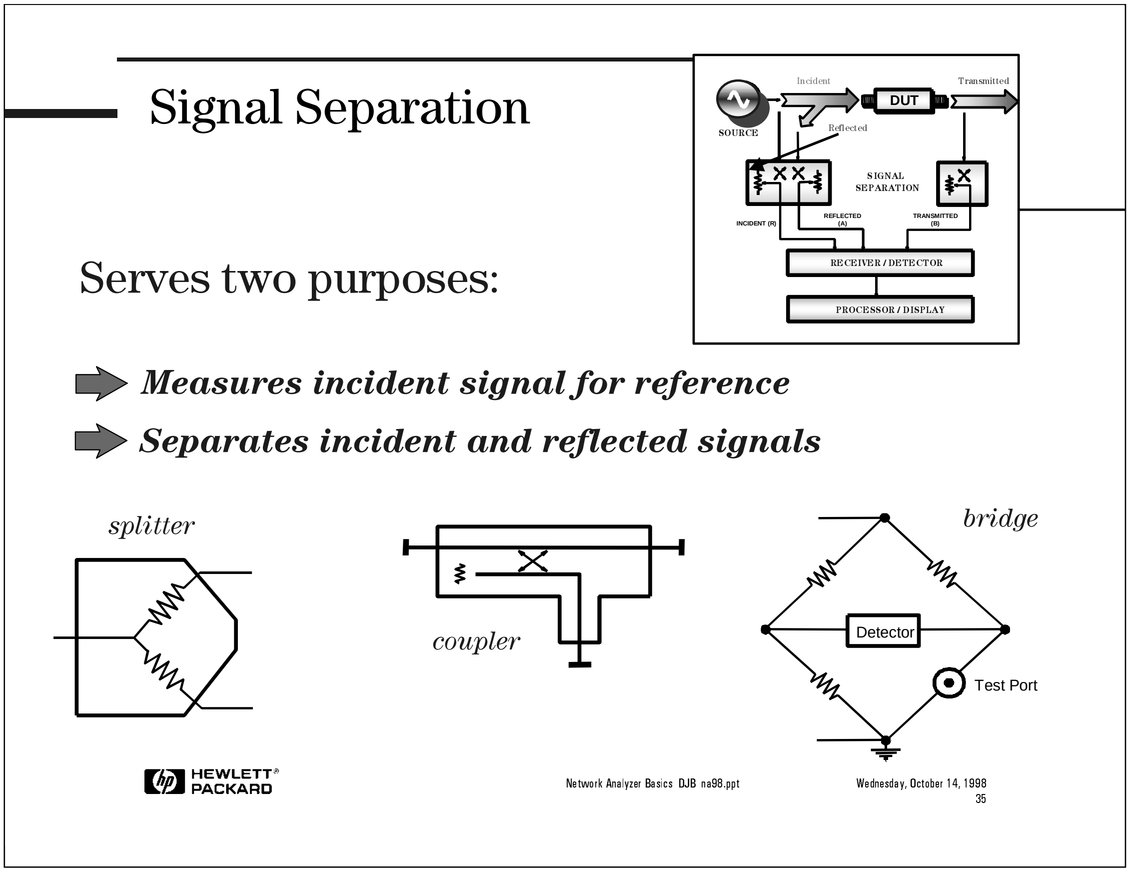


***Integrated, synthesized sources***



The signal source supplies the stimulus for our stimulus-response test system. We can either sweep the frequency of the source or sweep its power level. Traditionally, network analyzers used a separate source. These sources were either based on open-loop voltage-controlled oscillators (VCOs) which were cheaper, or more expensive synthesized sweepers which provided higher performance, especially for measuring narrowband devices. Excessive phase noise on open-loop VCOs degrades measurement accuracy considerably when measuring narrowband components over small frequency spans. Most network analyzers that HP sells today have integrated, synthesized sources.

**Slide #35**

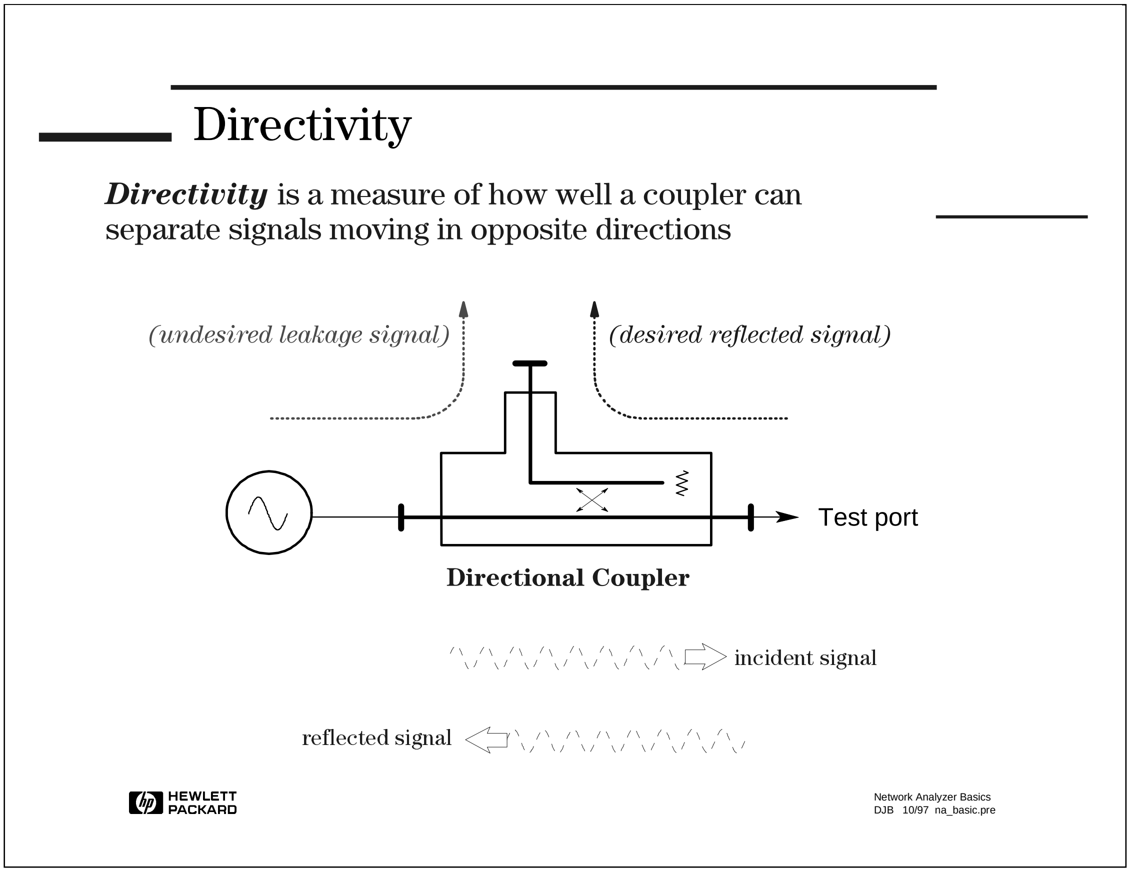



There are two functions that our signal-separation hardware must provide. The first is to measure a portion of the incident signal to provide a reference for ratioing. This can be done with splitters or directional couplers. Splitters are usually resistive. They are non-directional devices (more on directionality later) and can be very broadband. The trade-off is that they usually have 6 dB or more of loss in each arm. Directional couplers have very low insertion loss (through the main arm) and good isolation and directivity.

The second function of the signal-splitting hardware is to separate the incident (forward) and reflected (reverse) traveling waves at the input of our DUT. Again, couplers are ideal in that they are directional, have low loss, and high reverse isolation. However, due to the difficulty of making truly broadband couplers, bridges are often used.

More detailed information about these signal separation devices are provided in the Appendix Section.

**Slide #36**




Unfortunately, the signal separation devices are not perfect. For example, let's take a closer look at actual performance of a directional coupler.

Ideally, a signal traveling in the coupler's reverse direction will not appear at all at the coupled port, since its energy is either absorbed in the coupler's internal load or the external termination at the end of the main arm. In reality, however, some energy does leak through the coupled arm, as a result of finite isolation.

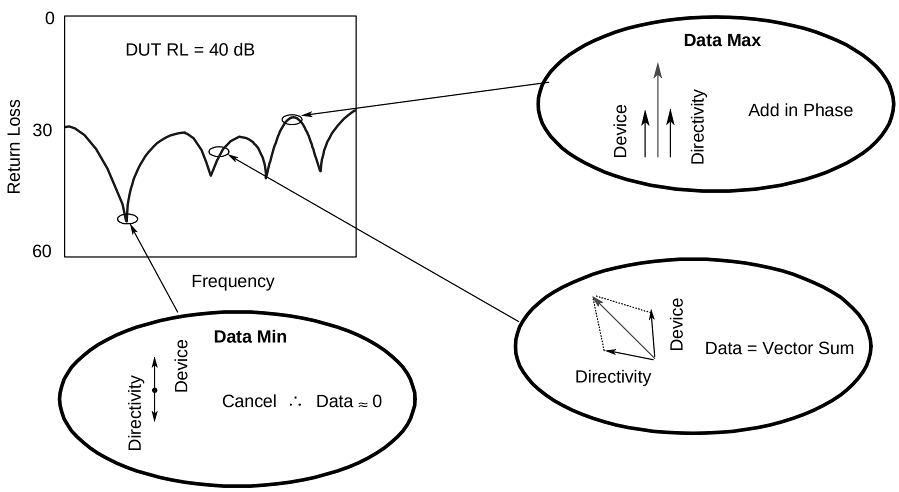
One of the most important measured parameter for couplers is their directivity. Directivity is a measure of a coupler's ability to separate signals flowing in opposite directions within the coupler. It can be thought of as the dynamic range available for reflection measurements. Directivity can be defined as:

Directivity (dB) = Isolation (dB) - Forward Coupling Factor (dB) - Loss (through-arm) (dB)

As we will see in the next slide, directivity adds error in our measured results.

**Slide #37**

## Interaction of Directivity with the DUT (Without Error Correction)

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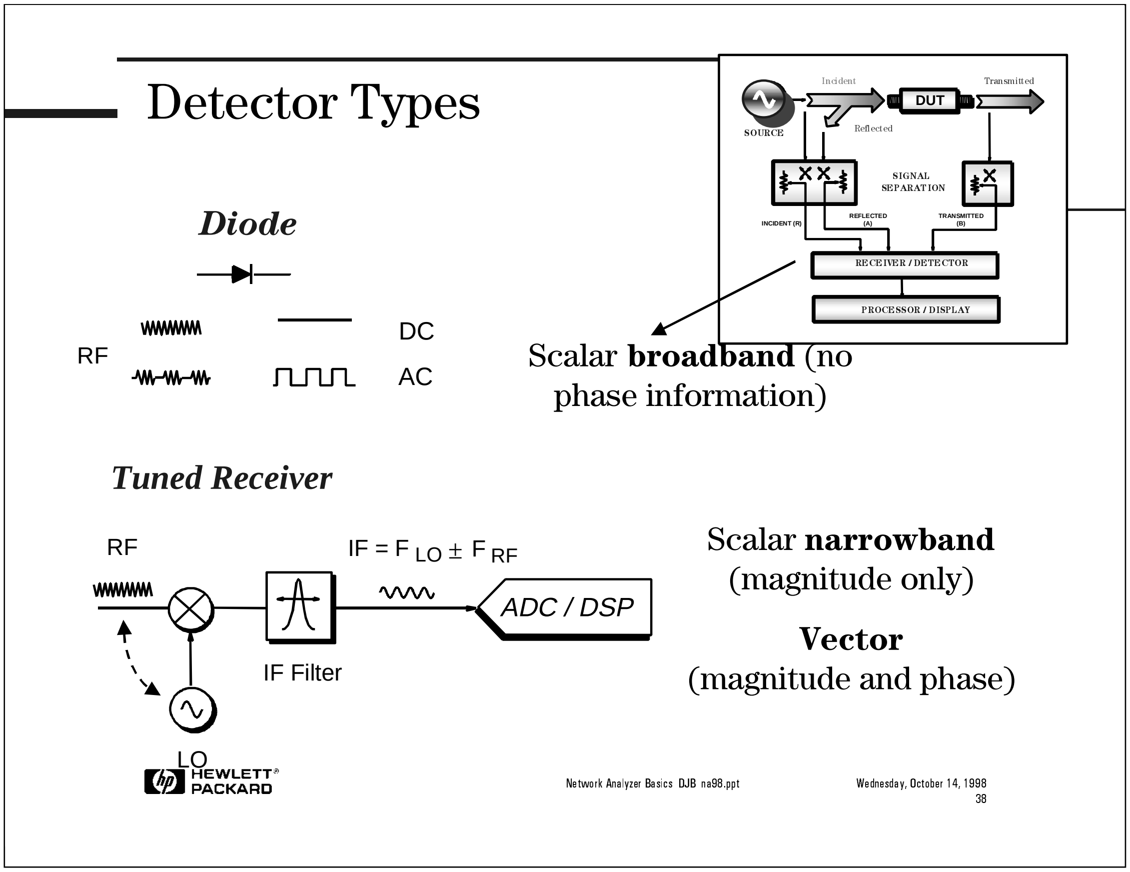
Directivity error is the main reason we see a large ripple pattern in many measurements of return loss. At the peaks of the ripple, directivity is adding in phase with the reflection from the DUT. In some cases, directivity will cancel the DUT's reflection, resulting in a sharp dip in the response.

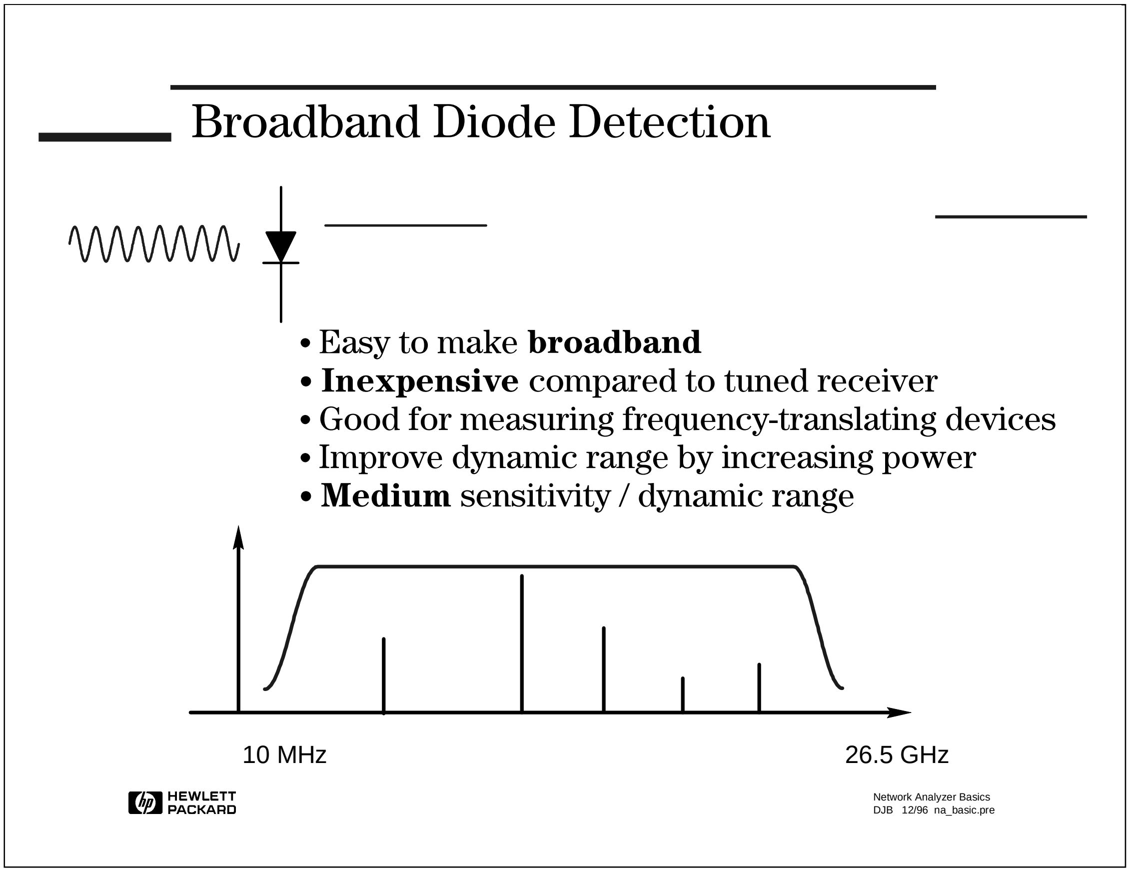
**Slide #38**




There are two basic ways of providing signal detection in network analyzers. Diode detectors convert the RF signal level to a proportional DC level. If the signal is amplitude modulated (AC detection), the diode strips the RF carrier from the modulation. Diode detection is inherently scalar, as phase information of the RF carrier is lost.

The tuned receiver uses a local oscillator (LO) to mix the RF down to a lower "intermediate" frequency (IF). The LO is either locked to the RF or the IF signal so that the receivers in the network analyzer are always tuned to the RF signal present at the input. The IF signal is bandpass filtered, which narrows the receiver bandwidth and greatly improves sensitivity and dynamic range. Modern analyzers use an analog-to-digital converter (ADC) and digital-signal processing (DSP) to extract magnitude and phase information from the IF signal. The tuned-receiver approach can be used in scalar or vector network analyzers.

**Slide #39**

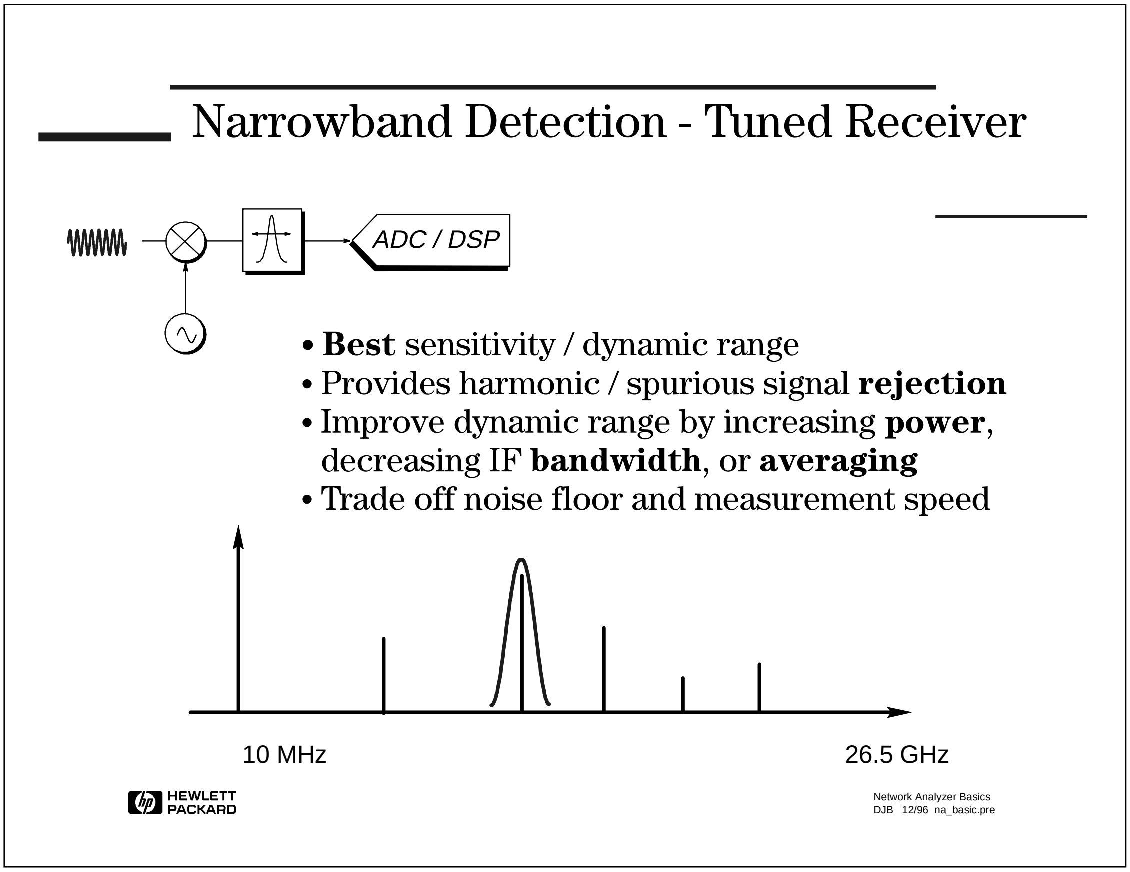



The main advantages of diode detectors are broadband frequency coverage ( < 10 MHz on the low end to > 26.5 GHz at the high end) and they are inexpensive compared to a tuned receiver. Diode detectors provide medium sensitivity and dynamic range: they can measure signals to -60 dBm or so and have a dynamic range around 60 to 75 dB, depending on the detector type. Their broadband nature limits their sensitivity and makes them sensitive to source harmonics and other spurious signals. Dynamic range is improved in measurements by increasing input power.

AC detection eliminates the DC drift of the diode as an error source, resulting in more accurate measurements. This scheme also reduces noise and other unwanted signals. The major benefit of DC detection is that there is no modulation of the RF signal, which can have adverse effects on the measurement of some devices. Examples include amplifiers with AGC or large DC gain, and narrowband filters.

One application where broadband diode detectors are very useful is measuring frequency-translating devices, particularly those with internal LOs. See slide 105 for a description of reference material.

**Slide #40**



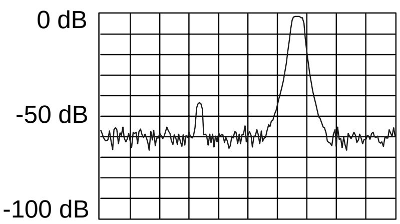

Tuned receivers provide the best sensitivity and dynamic range, and also provide harmonic and spurious-signal rejection. The narrow IF filter produces a considerably lower noise floor, resulting in a significant sensitivity improvement. For example, a microwave vector network analyzer (using a tuned receiver) might have a 3 kHz IF bandwidth, where a scalar analyzer's diode detector noise bandwidth might be 26.5 GHz. Measurement dynamic range is improved with tuned receivers by increasing input power, by decreasing IF bandwidth, or by averaging. The latter two techniques provide a trade off between noise floor and measurement speed. Averaging reduces the noise floor of the network analyzer (as opposed to just reducing the noise excursions as happens when averaging spectrum analyzer data) because we are averaging complex data. Without phase information, averaging does not improve analyzer sensitivity.

The same block diagram features that produce increased dynamic range also eliminate harmonic and spurious responses. As was mentioned earlier, the RF signal is downconverted and filtered before it is measured. The harmonics associated with the source are also downconverted, but they appear at frequencies outside the IF bandwidth and are therefore removed by filtering.

**Slide #41**

## Comparison of Receiver Techniques

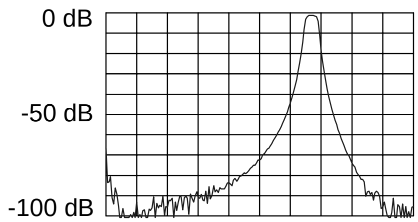
***Broadband (diode) detection***




-60 dBm Sensitivity

- higher noise floor
- false responses

***Narrowband (tuned-receiver) detection***




< -100 dBm Sensitivity

- high dynamic range
- harmonic immunity

***Dynamic range = maximum receiver power - receiver noise floor***

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Dynamic range is generally defined as the maximum power the receiver can accurately measure minus the receiver noise floor. There are many applications requiring large dynamic range. One of the most common are filter applications. As you can see here, at least 80 dB dynamic range is needed to properly characterize the rejection characteristics of this filter. The plots show a typical narrowband filter measured on an HP 8757 scalar network analyzer and on the HP 8510 vector network analyzer. Notice that the filter exhibits 90 dB of rejection but the scalar analyzer is unable to measure it because of its higher noise floor.

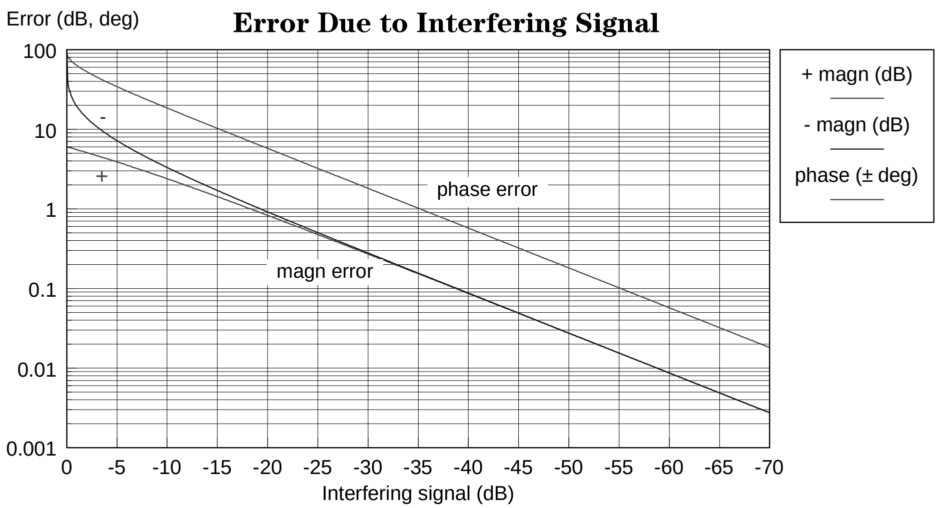
In the case where the scalar network analyzer was used with broadband diode detection, a harmonic or subharmonic from the source created a "false" response. For example, at some point on a broadband sweep, the second harmonic of the source might fall within the passband of the filter. If this occurs, the detector will register a response, even though the stopband of the filter is severely attenuating the frequency of the fundamental. This response from the second harmonic would show on the display at the frequency of the fundamental. On the tuned receiver, a spurious response such as this would be filtered away and would not appear on the display.

**Slide #42**

# Dynamic Range and Accuracy

***Dynamic range is very important for measurement accuracy!***

**Error Due to Interfering Signal**

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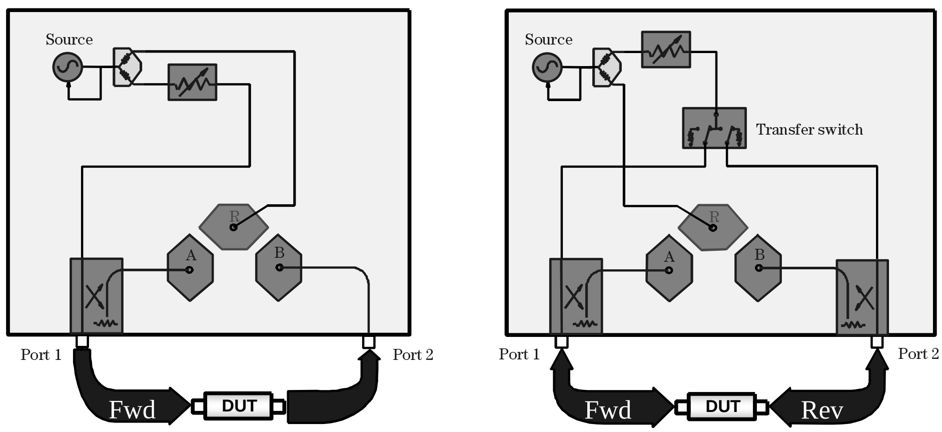
This plot shows the effect that interfering signals (sinusoids or noise) have on measurement accuracy. To get low measurement uncertainty due, more dynamic range is needed than the device exhibits. For example, to get less than 0.1 dB magnitude error and less than 0.5 degree phase error, our noise floor needs to be more than 39 dB below our measured power levels (note that there are other sources of error besides noise that may limit measurement accuracy). To achieve that level of accuracy while measuring 80 dB of rejection would require 115 dB of dynamic range. This could be accomplished by averaging test data with a tuned-receiver based network analyzer. HP network analyzers often have a clear competitive advantage by providing greater dynamic range than competitor's network analyzers.

**Slide #43**

# T/R Versus S-Parameter Test Sets

***Transmission/Reflection Test Set*** ***S-Parameter Test Set***




- RF always comes out port 1
- port 2 is always receiver
- **response**, **one-port** cal available

- RF comes out port 1 or port 2
- forward and reverse measurements
- **two-port** calibration possible

Network Analyzer Basics DJB na98.ppt Wednesday, October 14, 1998
43

There are two basic types of test sets that are used with network analyzers. For transmission/reflection (T/R) test sets, the RF power always comes out of test port one and test port two is always connected to a receiver in the analyzer. To measure reverse transmission or output reflection of the DUT, we must disconnect it, turn it around, and re-connect it to the analyzer. T/R-based network analyzers offer only response and one-port calibrations, so measurement accuracy is not as good as that which can be achieved with S-parameter test sets. However, T/R-based analyzers are more economical.

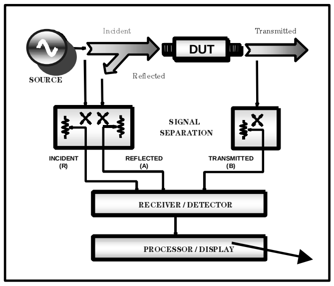
S-parameter test sets allow both forward and reverse measurements on the DUT, which are needed to characterize all four S-parameters. RF power can come out of either test port one or two, and either test port can be connected to a receiver. S-parameter test sets also allow full two-port (12-term) error correction, which is the most accurate form available. S-parameter network analyzers provide more performance than T/R-based analyzers, but cost more due to extra RF components in the test set.

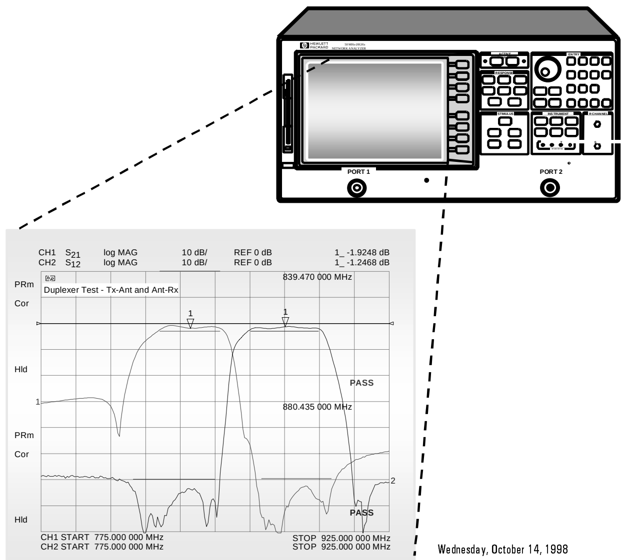
There are two different types of transfer switches that can be used in an S-parameter test set: solid-state and mechanical. Solid-state switches have the advantage of infinite lifetimes (assuming they are not damaged by too much power from the DUT). However, they are more lossy so they reduce the maximum output power of the network analyzer. Mechanical switches have very low loss and therefore allow higher output powers. Their main disadvantage is that eventually they wear out (after 5 million cycles or so). When using a network analyzer with mechanical switches, measurements are generally done in single-sweep mode, so the transfer switch is not continuously switching.

S-parameter test sets have two types of archectitures, 3-samplers and 4-samplers. More detailed information of the two archectitures is available in the Appendix section.

**Slide #44**

# Processor / Display




- markers
- limit lines
- pass/fail indicators
- linear/log formats
- grid/polar/Smith charts

Wednesday, October 14, 1998
44

The last major block of hardware in the network analyzer is the display/processor section. This is where the reflection and transmission data is formatted in ways that make it easy to interpret the measurement results. Most network analyzers have similar features such as linear and logarithmic sweeps, linear and log formats, polar plots, Smith charts, etc. Other common features are trace markers, limit lines, and pass/fail testing. Many of HP's network analyzers have specialized measurement features tailored to a particular market or application. One example is the HP 8730A tuner analyzer

**Slide #45**

# Internal Measurement Automation

Simple: **recall states**
More powerful:

- ***Test sequencing***
  - available on HP 8753 / 8720 families
  - keystroke recording
  - some advanced functions
- ***IBASIC***
  - available on HP 8711 family
  - sophisticated programs
  - custom user interfaces

```
ABCDEFGHIJKLMNOPQRSTUVWXYZ0123456789 + - / * = < > ( ) & ""  " , . / ? ; : ' [ ]
 1  ASSIGN @Hp8714 TO 800
 2  OUTPUT @Hp8714;"SYST:PRES; *WAI"
 3  OUTPUT @Hp8714;"ABOR;:INIT1:CONT OFF;*WAI"
 4  OUTPUT @Hp8714;"DISP:ANN:FREQ1:MODE SSTOP"
 5  OUTPUT @Hp8714;"DISP:ANN:FREQ1:MODE CSPAN"
 6  OUTPUT @Hp8714;"SENS1:FREQ:CENT 175000000 HZ;*WAI"
 7  OUTPUT @Hp8714;"ABOR;:INIT1:CONT OFF;:INIT1;*WAI"
 8  OUTPUT @Hp8714;"DISP:WIND1:TRAC:Y:AUTO ONCE"
 9  OUTPUT @Hp8714;"CALC1:MARK1 ON"
10  OUTPUT @Hp8714;"CALC1:MARK:FUNC BWID"
11  OUTPUT @Hp8714;"SENS2:STAT ON; *WAI"
12  OUTPUT @Hp8714;"SENS2:FUNC 'XFR:POW:RAT 1,0';DET NBAN; *WAI"
13  OUTPUT @Hp8714;"ABOR;:INIT1:CONT OFF;:INIT1;*WAI"
14  OUTPUT @Hp8714;"DISP:WIND2:TRAC:Y:AUTO ONCE"
15  OUTPUT @Hp8714;"ABOR;:INIT1:CONT ON;*WAI"
16  END
```

HEWLETT® PACKARD    Network Analyzer Basics DJB na98.ppt    Wednesday, October 14, 1998
45

All of HP's network analyzers offer some form of internal measurement automation. The most simple form is recall states. This is an easy way to set up the analyzer to a pre-configured measurement state, with all of the necessary instrument parameters.

More powerful automation can be achieved with test sequencing or Instrument BASIC (IBASIC). Test sequencing is available on the HP 8753/8720 families and provides keystroke recording and some advanced functions. IBASIC is available on the HP 8711C family (as an option) and provides the user with sophisticated programs and custom user interfaces and measurement personalities.

**Slide #46**

# HP Families of HF Vector Analyzers

## Microwave

**HP 8720D family**
- 40 GHz
- economical
- fast, small
- test mixers, high-power amps
- S-parameter

**HP 8510C family**
- 110 GHz *in coax*
- pulse systems
- antenna meas.
- Tx/Rx module test
- highest accuracy
- 4 S-parameter display

## RF

**HP 8712/14C**
- 3 GHz
- low cost, fast
- narrowband *and* broadband detection
- T/R test set only

**HP 8753E family**
- 6 GHz
- 52C: T/R test set
- 53E: S-parameter
- highest RF accuracy
- Offset and harmonic RF sweeps

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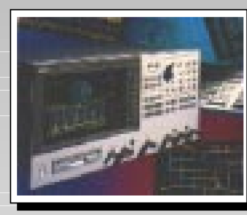
Shown here is a summary of HP's high-frequency families of vector network analyzers.

**Slide #47**

# HP Families of LF/RF Vector Analyzers

**Combination**

**HP 4395A**
- 500 MHz
- network/spectrum/ impedance (option)
- hot S.A. performance
- pc friendly



**HP 4396B**
- 1.8 GHz
- network/spectrum/ impedance (option)
- fast, highest accuracy
- time-gated spectrum (option)

**LF**

**HP E5100A/B**
- 300 MHz (180 MHz)
- economical
- fast, small
- xtals, resonators, filters
- equivalent circuit

**HP 8751A**
- 500 MHz
- fast
- accurate

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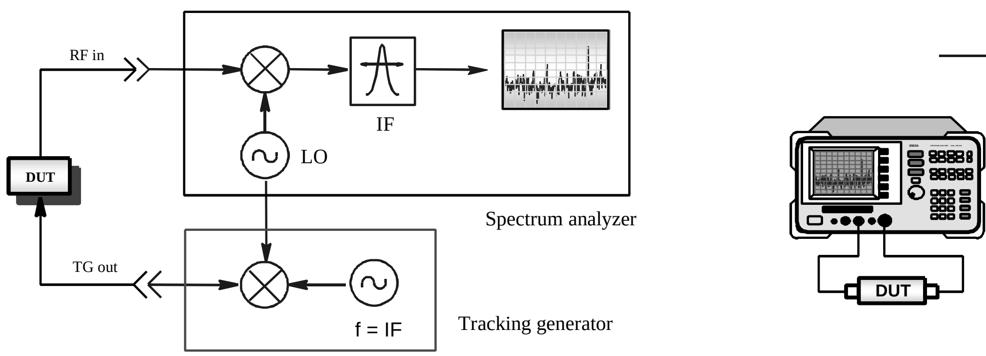
Shown here is a summary of HP's low-frequency vector network analyzers.

**Slide #48**

# Spectrum Analyzer / Tracking Generator




***Key differences from network analyzer:***

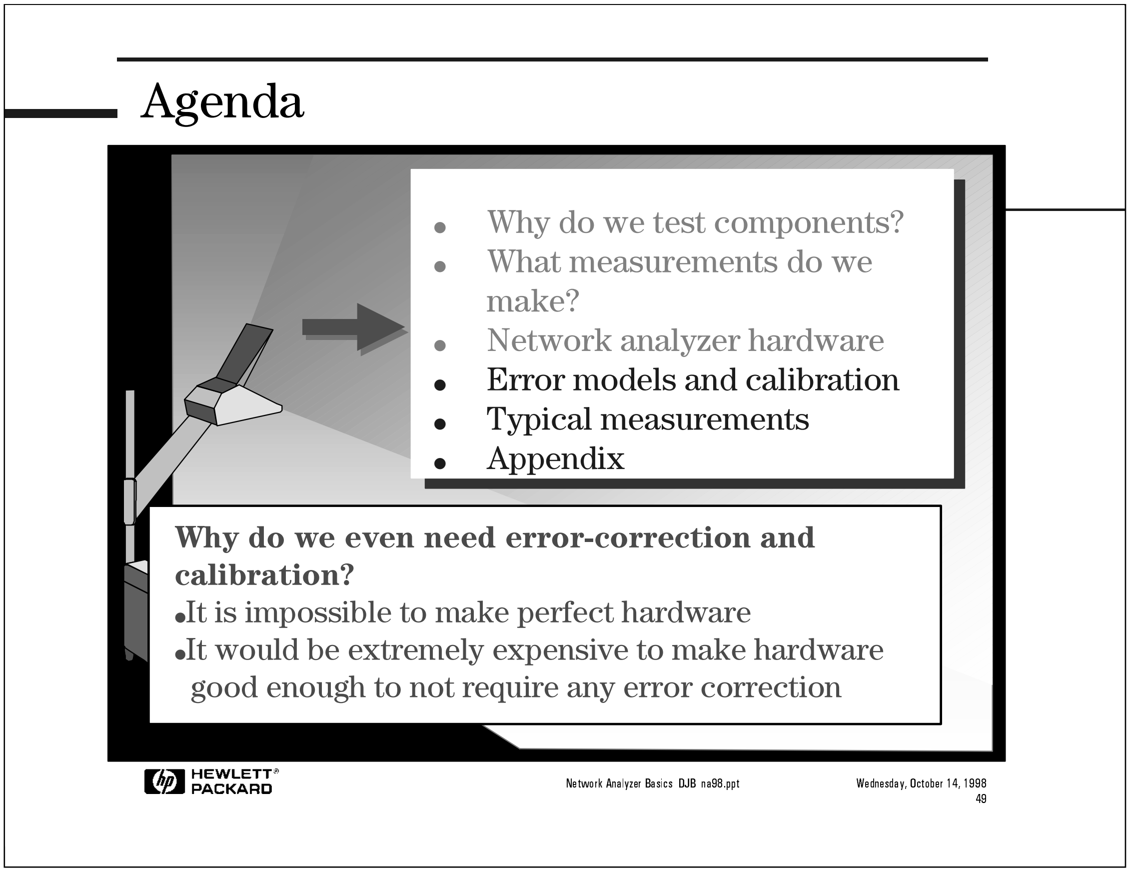
- **one channel** -- no ratioed or phase measurements
- More **expensive** than scalar NA
- Only error correction available is **normalization** (and possibly open-short averaging)
- Poorer **accuracy**
- Small **incremental cost** if SA is required for other measurements



If the main difference between spectrum and network analyzers is a source, why don't we add a tracking generator to our spectrum analyzer . . . then is it a network analyzer? Well, sort of.

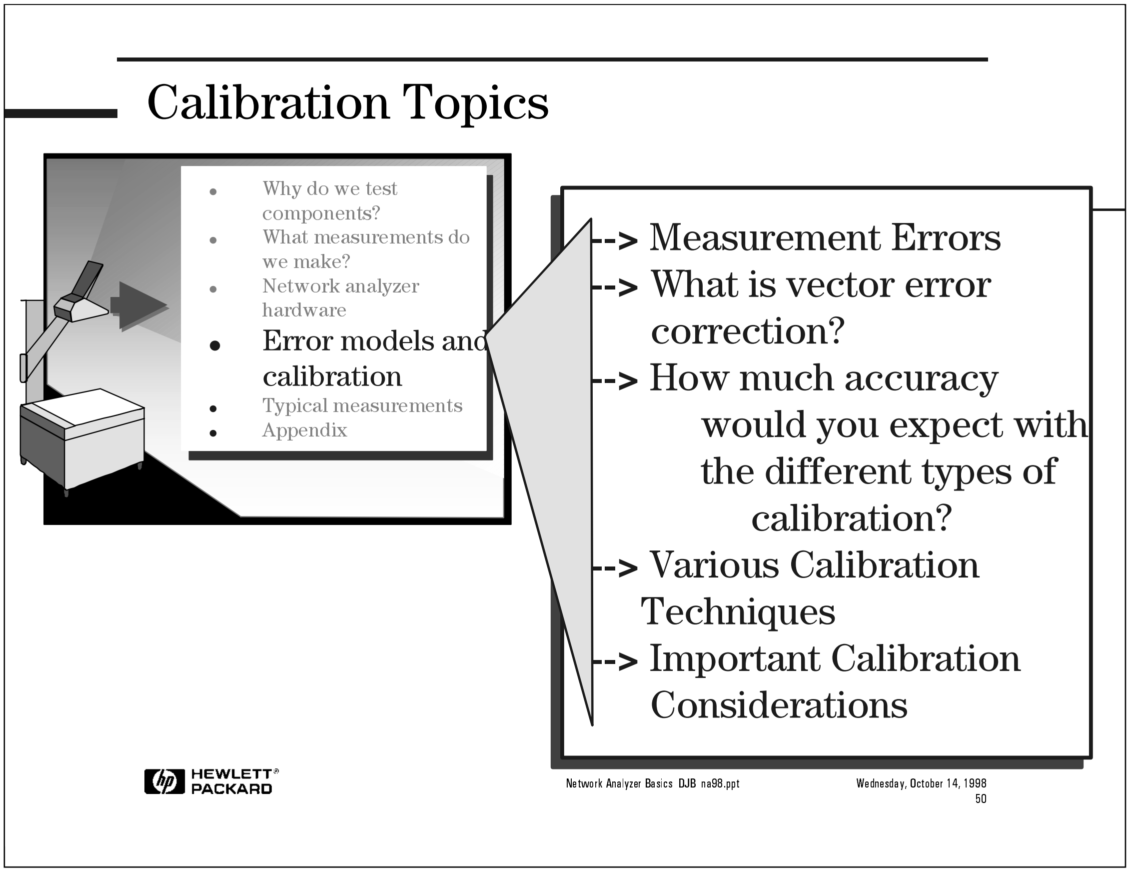
A spectrum analyzer with a tracking generator can make swept scalar measurements, but it is still a single-channel receiver. Therefore it cannot make ratio or phase measurements. Also, the only error correction available is normalization (and possible open-short averaging). We shall see later that scalar network analyzers such as the HP 8711C offer more advanced error-correction options. The amplitude accuracy with a spectrum analyzer is roughly an order of magnitude worse than on a scalar network analyzer (dB vs. tenths of dB). Finally, a spectrum analyzer with a tracking generator costs more than a scalar network analyzer, but it may be a small incremental cost to add a tracking generator if the spectrum analyzer is needed for other measurements.

**Slide #49**




In this next section, we will talk about the need for error correction and how it is accomplished. Why do we even need error-correction and calibration? It is impossible to make perfect hardware which obviously would not need any form of error correction. Even making the hardware good enough to not require any error correction for most devices would be extremely expensive. The best balance is to make the hardware as good as practically possible, balancing performance and cost. Error correction is then very useful to improve measurement accuracy.

**Slide #50**


Calibration Topics
Why do we test components?
What measurements do we make?
Network analyzer hardware
Error models and calibration
Typical measurements
Appendix
--> Measurement Errors
--> What is vector error correction?
--> How much accuracy would you expect with the different types of calibration?
--> Various Calibration Techniques
--> Important Calibration Considerations
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Network Analyzer Basics DJB na98.ppt
Wednesday, October 14, 1998
50

**Slide #51**

# Measurement Error Modeling

***Systematic errors***

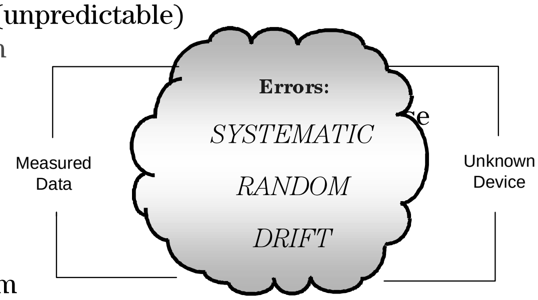
- due to **imperfections** in the analyzer and test setup
- are assumed to be **time invariant** (predictable)
- <u>can</u> be characterized (during calibration process) and **mathematically removed** during measurements

***Random errors***

- **vary** with time in random fashion (unpredictable)
- **<u>cannot be removed</u>** by calibration
- main contributors:
  - **instrument noise** (source noise, IF noise floor, etc.)
  - **switch** repeatability
  - **connector** repeatability

***Drift errors***

- are due to instrument or test-system performance changing ***after*** a calibration has been done
- are primarily caused by **temperature variation**
- **<u>can be removed</u>** by further calibration(s)




HEWLETT PACKARD Network Analyzer Basics DJB na98.ppt Wednesday, October 14, 1998 51

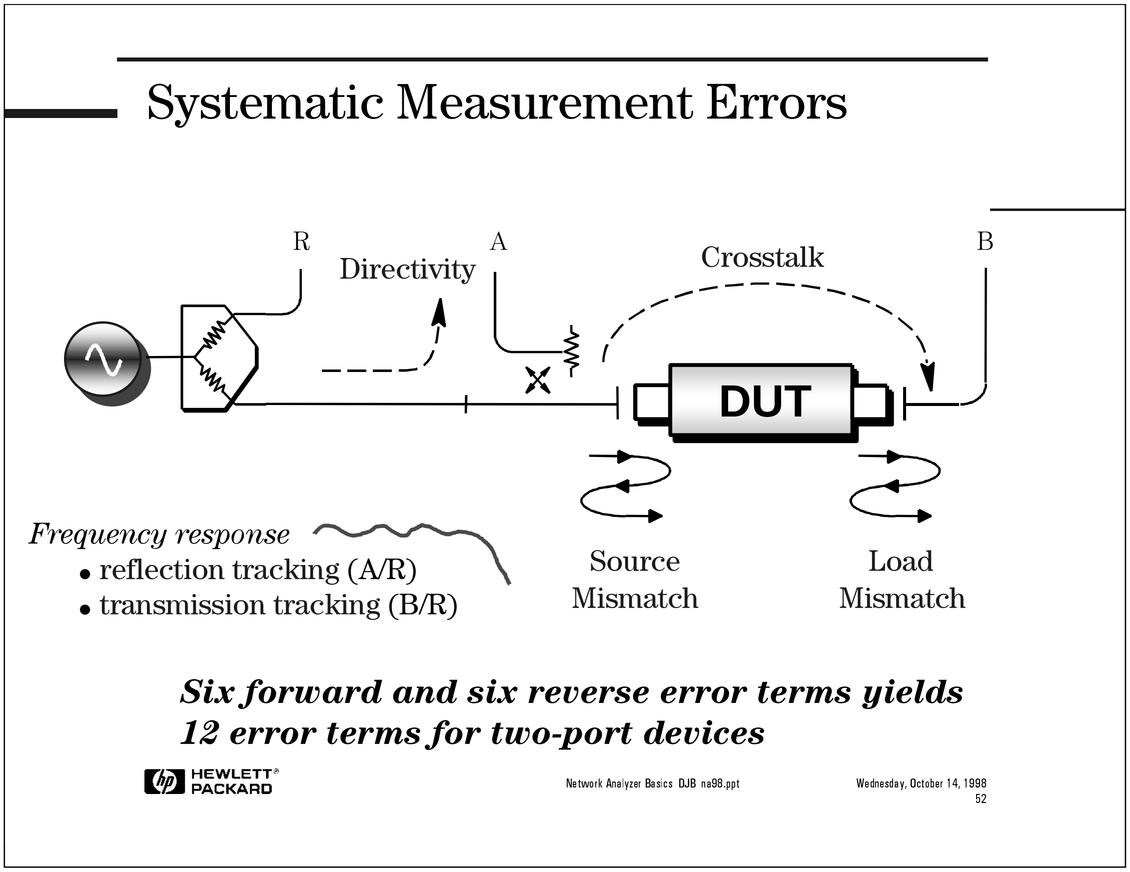
Let us look at the three basic sources of measurement error: systematic, random and drift.

Systematic errors are due to imperfections in the analyzer and test setup. They are repeatable (therefore predictable), and assumed to be time invariant. Systematic errors are characterized during the calibration process and mathematically removed during measurements.

Random errors are unpredictable since they vary with time in a random fashion. Therefore, they cannot be removed by calibration. The main contributors to random error are instrument noise (source phase noise, sampler

Drift errors are due to the instrument or test-system performance changing *after* a calibration has been done. Drift is primarily caused by temperature variation and it can be removed by further calibration(s). The timeframe over which a calibration remains accurate is dependent on the rate of drift that the test system undergoes in the user's test environment. Providing a stable ambient temperature usually goes a long way towards minimizing drift.

**Slide #52**



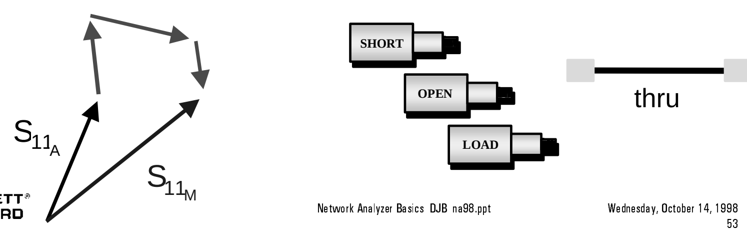

Shown here are the major systematic errors associated with network measurements. The errors relating to signal leakage are directivity and crosstalk. Errors related to signal reflections are source and load match. The final class of errors are related to frequency response of the receivers, and are called reflection and transmission tracking. The full two-port error model includes all six of these terms for the forward direction and the same six (with different data) in the reverse direction, for a total of twelve error terms. This is why we often refer to two-port calibration as twelve-term error correction.

**Slide #53**

# Types of Error Correction

Two main types of error correction:

- **response (normalization)**
  - simple to perform
  - only corrects for tracking errors
  - stores reference trace in memory, then does data divided by memory
- **vector**
  - requires more standards
  - requires an analyzer that can measure phase
  - accounts for all major sources of systematic error




The two main types of error correction that can be done are response (normalization) corrections and vector corrections. Response calibration is simple to perform, but only corrects for a few of the twelve possible systematic error terms (the tracking terms). Response calibration is essentially a normalized measurement where a reference trace is stored in memory, and subsequent measurement data is divided by this memory trace. A more advanced form of response calibration is open/short averaging for reflection measurements using broadband diode detectors. In this case, two traces are averaged together to derive the reference trace.

Vector-error correction requires an analyzer that can measure (but not necessarily display) both magnitude and phase data. It also requires measurements of more calibration standards. Vector-error correction can account for all major sources of systematic error and can give very accurate measurements.

Note that a response calibration can be performed on a vector network analyzer, in which case we store a complex (vector) reference trace in memory, so that we can display normalized magnitude or phase data. This is not the same as vector-error correction however (and not as accurate), because we are not measuring and removing the individual systematic errors, all of which are complex or vector quantities.

**Slide #54**

# What is Vector-Error Correction?

- Process of characterizing systematic error terms
  - measure **known standards**
  - remove effects from subsequent measurements.
- **1-port calibration** *(reflection measurements)*
  - only 3 systematic error terms measured
  - directivity, source match, and reflection tracking
- **Full 2-port calibration** *(reflection and transmission measurements)*
  - 12 systematic error terms measured
  - usually requires 12 measurements on four known standards (SOLT)
- Some standards can be measured **multiple** times (e.g., THRU is usually measured four times)
- Standards defined in **cal kit definition** file
  - network analyzer contains standard cal kit definitions
  - **CAL KIT DEFINITION MUST MATCH ACTUAL CAL KIT USED!**

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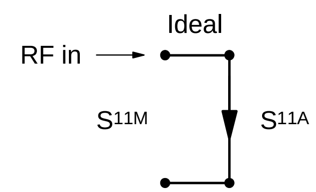
Vector-error correction is the process of characterizing systematic error terms by measuring known calibration standards, and then removing the effects of these errors from subsequent measurements.

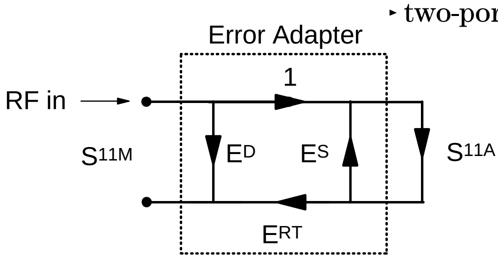
One-port calibration is used for reflection measurements and can measure and remove three systematic error terms (directivity, source match, and reflection tracking). Full two-port calibration can be used for both reflection and transmission measurements, and all twelve systematic error terms are measured and removed. Two-port calibration usually requires twelve measurements on four known standards (short-open-load-thru or SOLT). Some standards are measured multiple times (e.g., the thru standard is usually measured four times). The standards themselves are defined in a cal-kit definition file, which is stored in the network analyzer. HP network analyzers contain all of the cal-kit definitions for our standard calibration kits. In order to make accurate measurements, the cal-kit definition MUST MATCH THE ACTUAL CALIBRATION KIT USED!

**Slide #55**

## Reflection: One-Port Model




- If you know the systematic error terms, you can solve for the actual S-parameter
- Assumes good termination at port two if testing two-port devices
- If port 2 is connected to the network analyzer *and* DUT reverse isolation is low (e.g., filter passband):
  - assumption of good termination is not valid
  - two-port error correction yields better results




$E_D$ = Directivity
$E_{RT}$ = Reflection tracking
$E_S$ = Source Match
$S_{11M}$ = Measured
$S_{11A}$ = Actual

$$S_{11M} = E_D + E_{RT} \frac{S_{11A}}{1 - E_S S_{11A}}$$

***To solve for $S_{11A}$, we have 3 equations and 3 unknowns***

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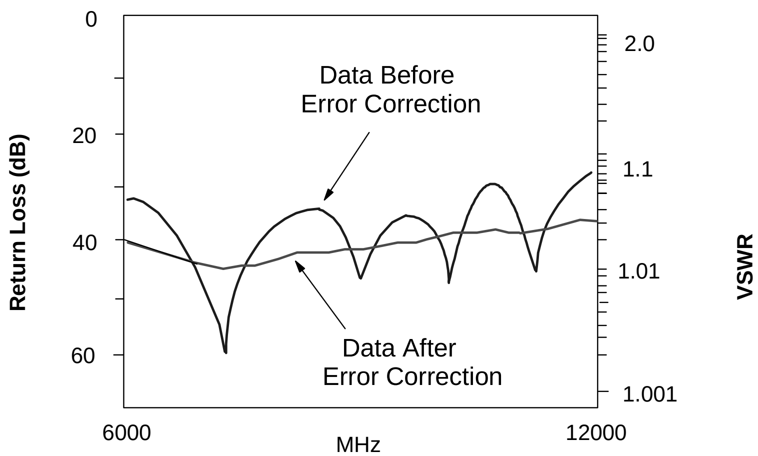
Taking the simplest case of a one-port reflection measurement, we have three systematic errors and one equation to solve. In order to do this, we must create three equations with three unknowns and solve them simultaneously. To do this, we measure three known standards, for example, a short, an open, and a Zo load. Solving the equations will yield the systematic error terms and allow us to derive the actual reflection S-parameters of the device from our measurements.

When measuring two-port devices, a one-port calibration assumes a good termination at port two of the device. If this condition is met (by connecting a load calibration standard for example), the one-port calibration is quite accurate. If port two of the device is connected to the network analyzer and the reverse isolation of the DUT is low (for example, filter passbands or cables), the assumption of a good load termination is not valid. In these cases, two-port error correction provides more accurate measurements. An example of a two-port device where load match is not important is an amplifier. The reverse isolation of the amplifier allows one-port calibration to be used effectively. An example of the measurement error that can occur when measuring a two-port filter using a one-port calibration will be shown shortly.

**Slide #56**

## Before and After One-Port Calibration

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Shown here is a plot of reflection with and without one-port calibration. Without error correction, we see the classic ripple pattern caused by the systematic errors interfering with the measured signal. The error-corrected trace is much smoother and better represents the device's actual reflection performance.

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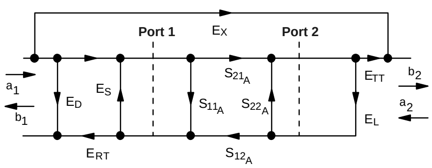
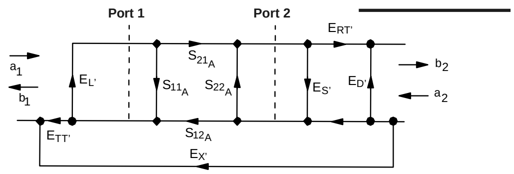
## Slide #57

# Two-Port Error Correction

Forward model




Reverse model




$E_D$ = Fwd Directivity
$E_S$ = Fwd Source Match
$E_{RT}$ = Fwd Reflection Tracking
$E_L$ = Fwd Load Match
$E_{TT}$ = Fwd Transmission Tracking
$E_X$ = Fwd Isolation

$E_{D'}$ = Rev Directivity
$E_{S'}$ = Rev Source Match
$E_{RT'}$ = Rev Reflection Tracking
$E_{L'}$ = Rev Load Match
$E_{TT'}$ = Rev Transmission Tracking
$E_{X'}$ = Rev Isolation

- Notice that each actual S-parameter is a function of all four measured S-parameters
- Analyzer must make forward *and* reverse sweep to update any one S-parameter
- Luckily, you don't need to know these equations to ***use*** network analyzers!!!

$$S_{11a} = \frac{(\frac{S_{11m}-E_D}{E_{RT}})(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}{(1+\frac{S_{11m}-E_D'}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

$$S_{21a} = \frac{(\frac{S_{21m}-E_X}{E_{TT}})(1+\frac{S_{22m}-E_D'}{E_{RT}'}(E_S'-E_L))}{(1+\frac{S_{11m}-E_D}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

$$S_{12a} = \frac{(\frac{S_{12m}-E_X'}{E_{TT}'})(1+\frac{S_{11m}-E_D}{E_{RT}}(E_S-E_L'))}{(1+\frac{S_{11m}-E_D}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

$$S_{22a} = \frac{(\frac{S_{22m}-E_D'}{E_{RT}'})(1+\frac{S_{11m}-E_D}{E_{RT}}E_S) - E_L'(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}{(1+\frac{S_{11m}-E_D}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

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Two-port error correction is the most accurate form of error correction since it accounts for all of the major sources of systematic error. The error model for a two-port device is shown above. Shown below are the equations to derive the actual device S-parameters from the measured S-parameters, once the systematic error terms have been characterized. Notice that each actual S-parameter is a function of all four measured S-parameters. The network analyzer must make a forward and reverse sweep to update any one S-parameter. Luckily, you don't need to know these equations to use network analyzers!!!

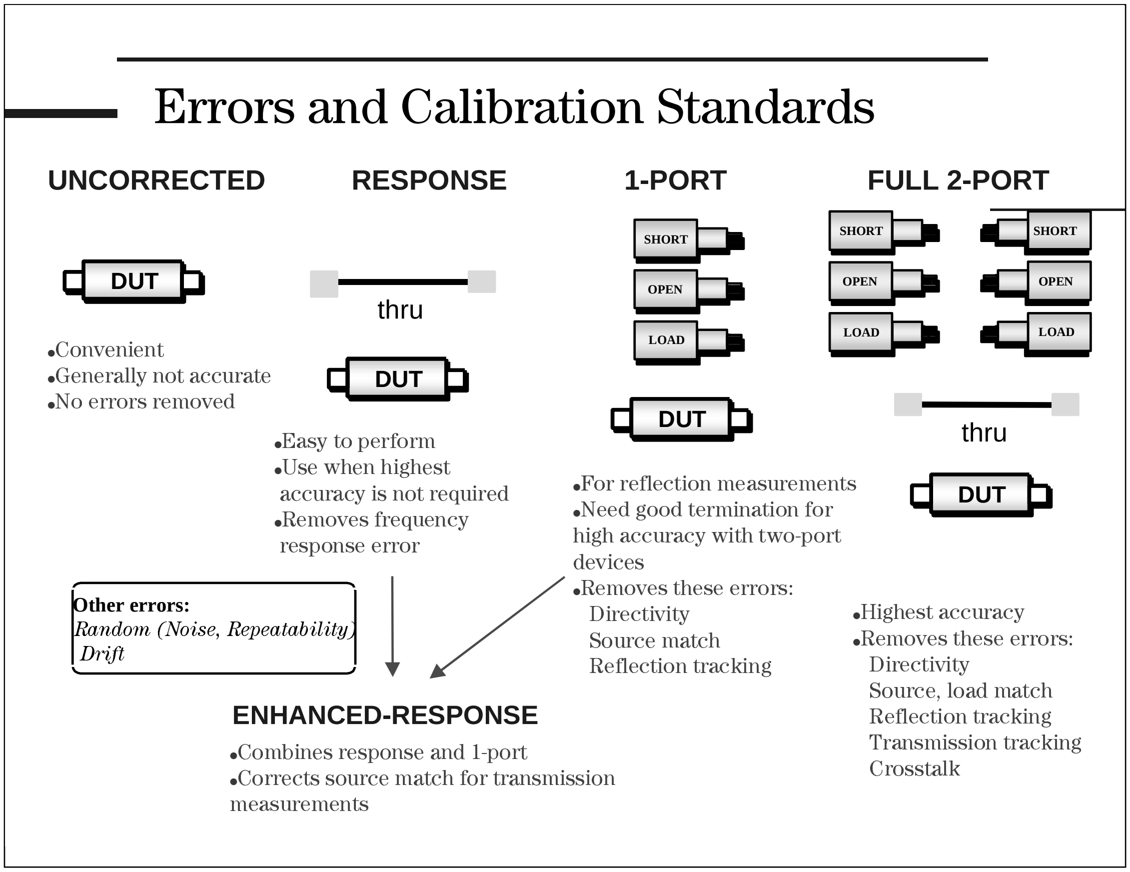
$$S_{11a} = \frac{(\frac{S_{11m}-E_D}{E_{RT}})(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}{(1+\frac{S_{11m}-E_D'}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

$$S_{21a} = \frac{(\frac{S_{21m}-E_X}{E_{TT}})(1+\frac{S_{22m}-E_D'}{E_{RT}'}(E_S'-E_L))}{(1+\frac{S_{11m}-E_D}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

$$S_{12a} = \frac{(\frac{S_{12m}-E_X'}{E_{TT}'})(1+\frac{S_{11m}-E_D}{E_{RT}}(E_S-E_L'))}{(1+\frac{S_{11m}-E_D}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

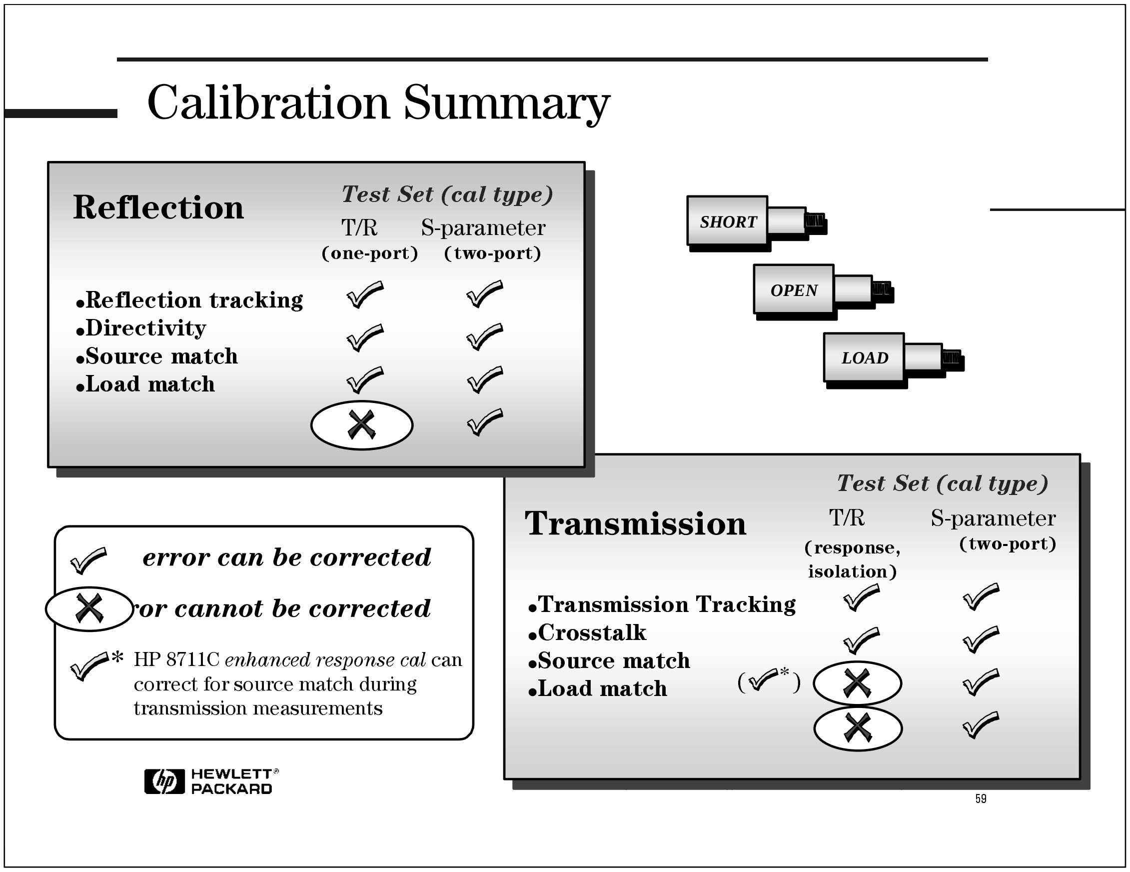
$$S_{22a} = \frac{(\frac{S_{22m}-E_D'}{E_{RT}'})(1+\frac{S_{11m}-E_D}{E_{RT}}E_S) - E_L'(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}{(1+\frac{S_{11m}-E_D}{E_{RT}}E_S)(1+\frac{S_{22m}-E_D'}{E_{RT}'}E_S') - E_L'E_L(\frac{S_{21m}-E_X}{E_{TT}})(\frac{S_{12m}-E_X'}{E_{TT}'})}$$

**Slide #58**




A network analyzer can be used for uncorrected measurements, or with any one of a number of calibration options, including response calibrations and one- or two-port vector calibrations. A summary of these calibrations is shown above. We will explore the measurement uncertainties associated with the various calibration schemes in this section.

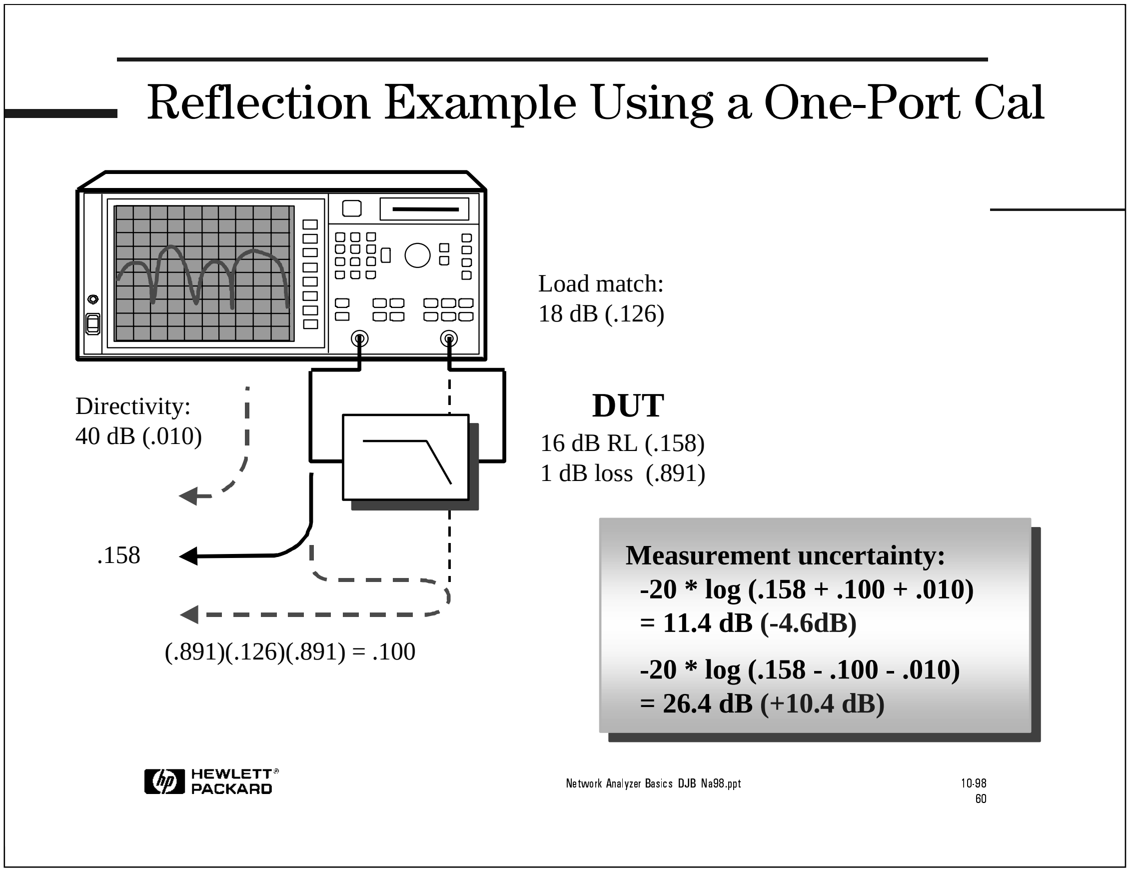
**Slide #59**




This summary shows which error terms are accounted for when using analyzers with T/R test sets (such as the HP 8711C family) and S-parameter test sets (such as the HP 8753/8720 families).

The following examples show how measurement uncertainty can be estimated when measuring two-port devices with a T/R-based network analyzer.

**Slide #60**

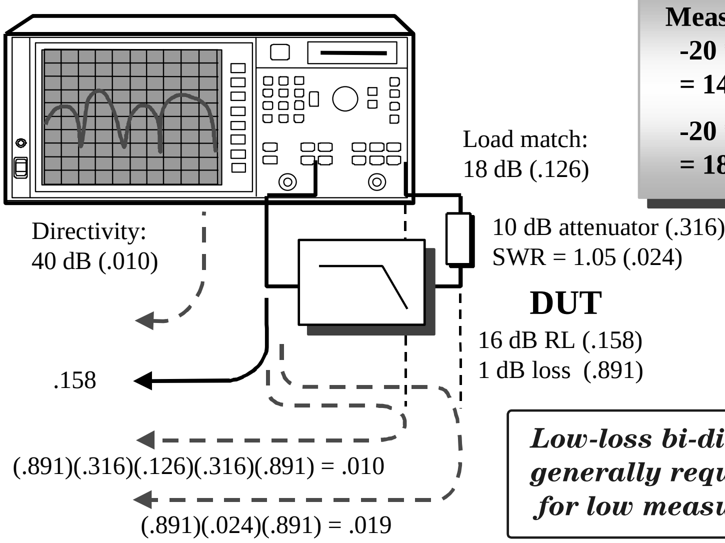



Here is an example of how much measurement uncertainty we might encounter when measuring the input match of a filter after a one-port calibration. In this example, our filter has a return loss of 16 dB, and 1 dB of insertion loss. The raw load match of an 8711C network analyzer is specified to be 18 dB (although it's often significantly better than this). The reflection from the test port connected to the filter's output is attenuated by twice the filter loss, which is only 2 dB total in this case. This value is not adequate to sufficiently suppress the effects of this error signal, which illustrates why low-loss devices are difficult to measure accurately. To determine the measurement uncertainty of this example, it is necessary to add and subtract the undesired reflection signal (with a reflection coefficient of 0.100) with the signal reflecting from the DUT (0.158) (to be consistent with the next example, we will also include the effect of the directivity error signal). The measured return loss of the 16-dB filter may appear to be anywhere from 11.4 dB to 26.4 dB, allowing too much room for error. In production testing, these errors could easily cause filters which met specification to fail, while filters that actually did not meet specification might pass. In tuning applications, filters could be mistuned as operators try to compensate for the measurement error.

When measuring an amplifier with good isolation between output and input (i.e., where the isolation is much greater than the gain), there is much less measurement uncertainty. This is because the reflection caused by the load match is severely attenuated by the product of the amplifier's isolation and gain. To improve measurement uncertainty for a filter, the output of the filter must be disconnected from the analyzer and terminated with a high-quality load, or a high-quality attenuator can be inserted between the filter and port two of the network analyzer. Both techniques improve the analyzer's effective load match.

**Slide #61**

# Using a One-Port Cal + Attenuator




**Measurement uncertainty:**
**-20 * log (.158 + .039)**
**= 14.1 dB (-1.9 dB)**

**-20 * log (.158 - .039)**
**= 18.5 dB (+2.5 dB)**

***Low-loss bi-directional devices generally require two-port calibration for low measurement uncertainty***

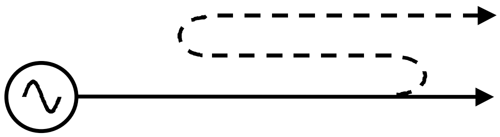
Worst-case error = .010 + .010 + .019 = .039

Network Analyzer Basics DJB Na98.ppt 10-98
61

Let's see how much improvement we get by adding an attenuator between the output of the filter and our network analyzer. If we insert a 10 dB attenuator with a SWR of 1.05 between port two of the network analyzer and the filter used in the previous example, our effective load match improves to 28.6 dB (-20*log[10exp(-32.3/20)+10exp(-38/20)]). This value is the combination of a 32.3 dB match from the attenuator (SWR = 1.05) and a 38 dB match from the network analyzer (since the error signal travels through the attenuator twice, the analyzer's load match is improved by twice the value of the attenuator). Our worst-case uncertainty is now reduced to +2.5 dB, -1.9 dB, instead of the +10.4 dB, -4.6 dB we had without the 10 dB attenuator. While not as good as what could be achieved with two-port calibration, this level of accuracy may be sufficient for manufacturing applications.

**Slide #62**

## Transmission Example Using Response Cal



RL = 18 dB (.126)

RL = 14 dB (.200)

Thru calibration (normalization) builds error into measurement due to source and load match interaction

**Calibration Uncertainty**

$$= (1 \pm \rho_S \rho_L)$$

$$= (1 \pm (.200)(.126))$$

$$= \pm 0.22 \text{ dB}$$

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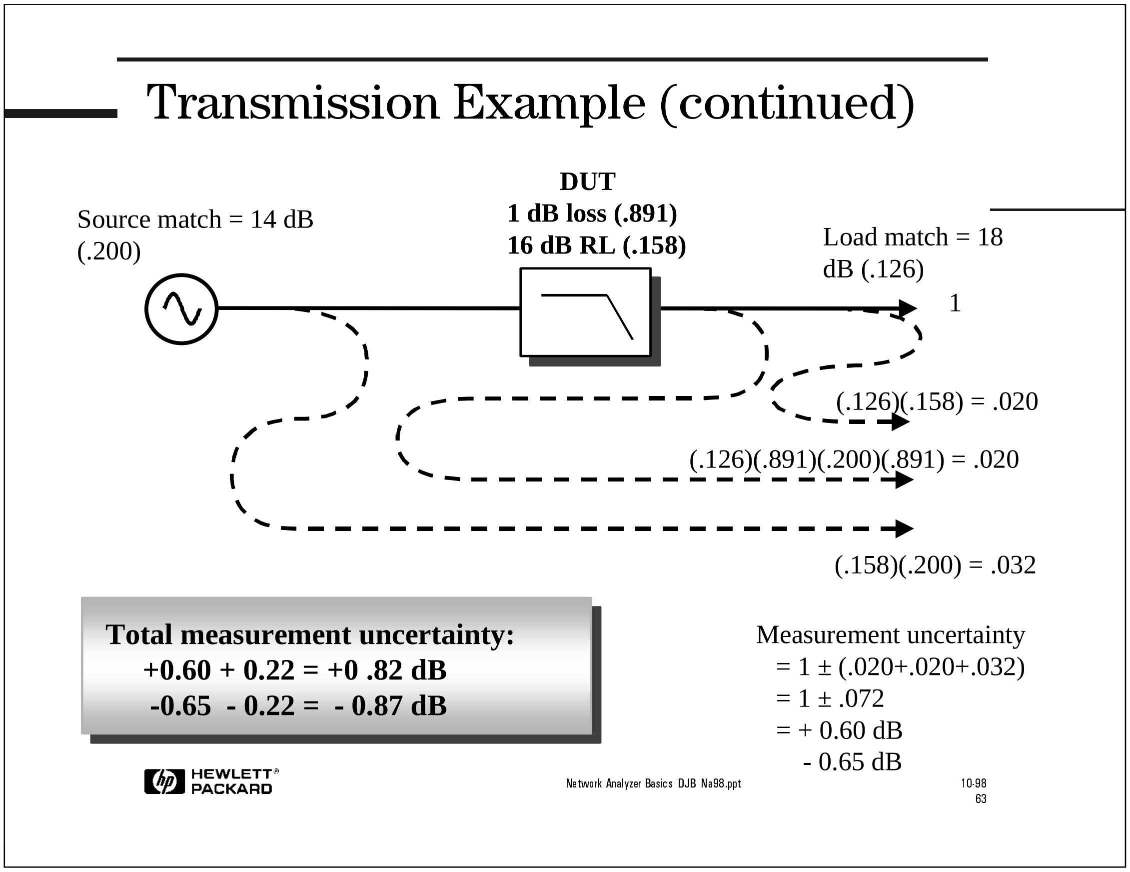
10-98
62

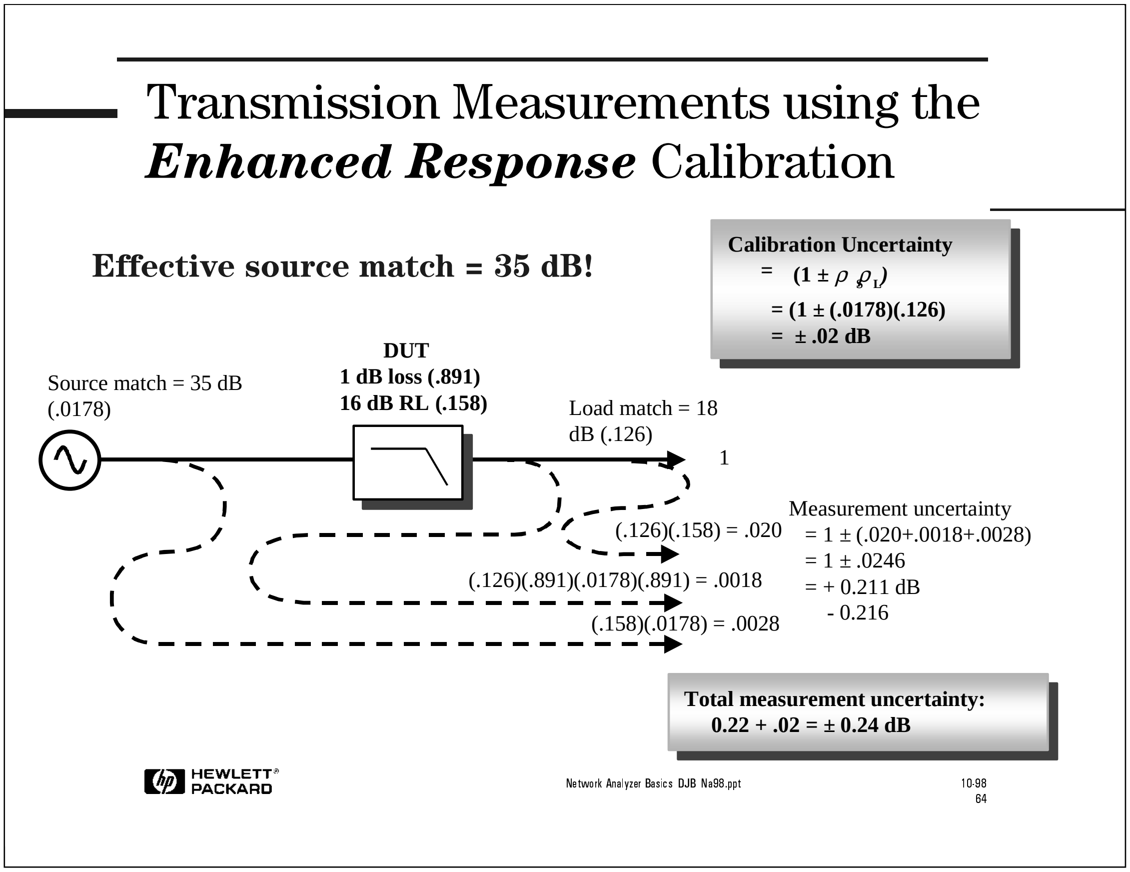
Let's do an example transmission measurement using only response calibration. Response calibrations offer simplicity, but with some compromise in measurement accuracy. In making a filter transmission measurement using only response calibration, the first step is to make a through connection between the two test port cables (with no DUT in place). For this example, test port specifications for the HP 8711C network analyzer will be used (note: the HP 8713C and 8714C have considerably better uncorrected source match than the 8711C). The ripple caused by this amount of mismatch is calculated as ±0.22 dB, and is now present in the reference data. It must be added to the uncertainty when the DUT is measured in order to compute worst-case overall measurement uncertainty.

**Slide #63**




Now let's look at the measurement uncertainty when the DUT is inserted. We will use the same loss and mismatch specifications for the DUT and analyzer as before. We have three main error signals due to reflections between the ports of the analyzer and the DUT. There are higher-order reflections present as well, but they don't add any significant error since they are small compared to the three main terms. One of the signals passes through the DUT twice, so it is attenuated by twice the loss of the DUT. The worst case is when all of the reflected error signals add together in-phase (.020 + .020 + .032 = .072). In that case, we get a measurement uncertainty of +0.60 dB, -0.65 dB. The total measurement uncertainty, which must include the 0.22 dB of error incorporated into our calibration measurement, is about ± 0.85 dB.

**Slide #64**




A new feature of the 8711C family of RF economy network analyzers is the *enhanced* response calibration. This calibration requires the measurement of short, open, load, and thru standards for transmission measurements. Essentially, it combines a one-port cal and a response cal to allow correction of source match during transmission measurements. Recall that a standard response calibration (such as found in the HP 8711B, 8752C, or 8753D) cannot correct for the source and load match error terms.

The enhanced response calibration improves the effective source match during transmission measurements to around 35 dB, instead of the 14 dB we used in the previous examples. This greatly reduces the calibration error (± 0.02 dB instead of ± 0.22 dB), as well as the two measurement error terms that involve interaction with the effective source match. Our total measurement error is now ± 0.24 dB, instead of the previously calculated ± 0.85 dB.

**Slide #65**

# Using the ***Enhanced Response*** Calibration Plus an Attenuator

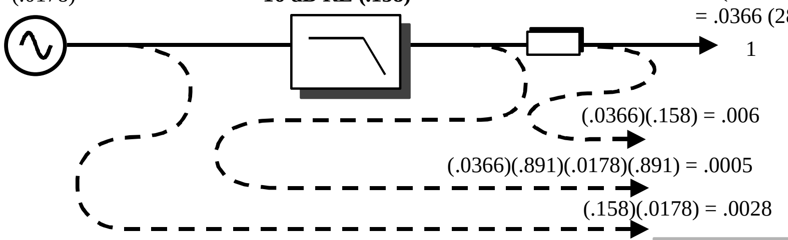
10 dB attenuator (.316)
SWR = 1.05 (.024 or 32.4 dB)
Analyzer load match =18 dB (.126)

**Calibration Uncertainty**
**= (1 ± $\rho_S \rho_L$)**
**= (1 ± (.0178)(.0366)**
**= ± .01 dB**

Source match = 35 dB (.0178)

**DUT**
**1 dB loss (.891)**
**16 dB RL (.158)**

Effective load match = (.316 x .316 x .126 + .024)
= .0366 (28.7dB)




Measurement uncertainty
= 1 ± (.006+.0005+.0028)
= 1 ± .0093
= ± 0.08 dB

**Total measurement uncertainty:**
**0.01 + .08 = ± 0.09 dB**

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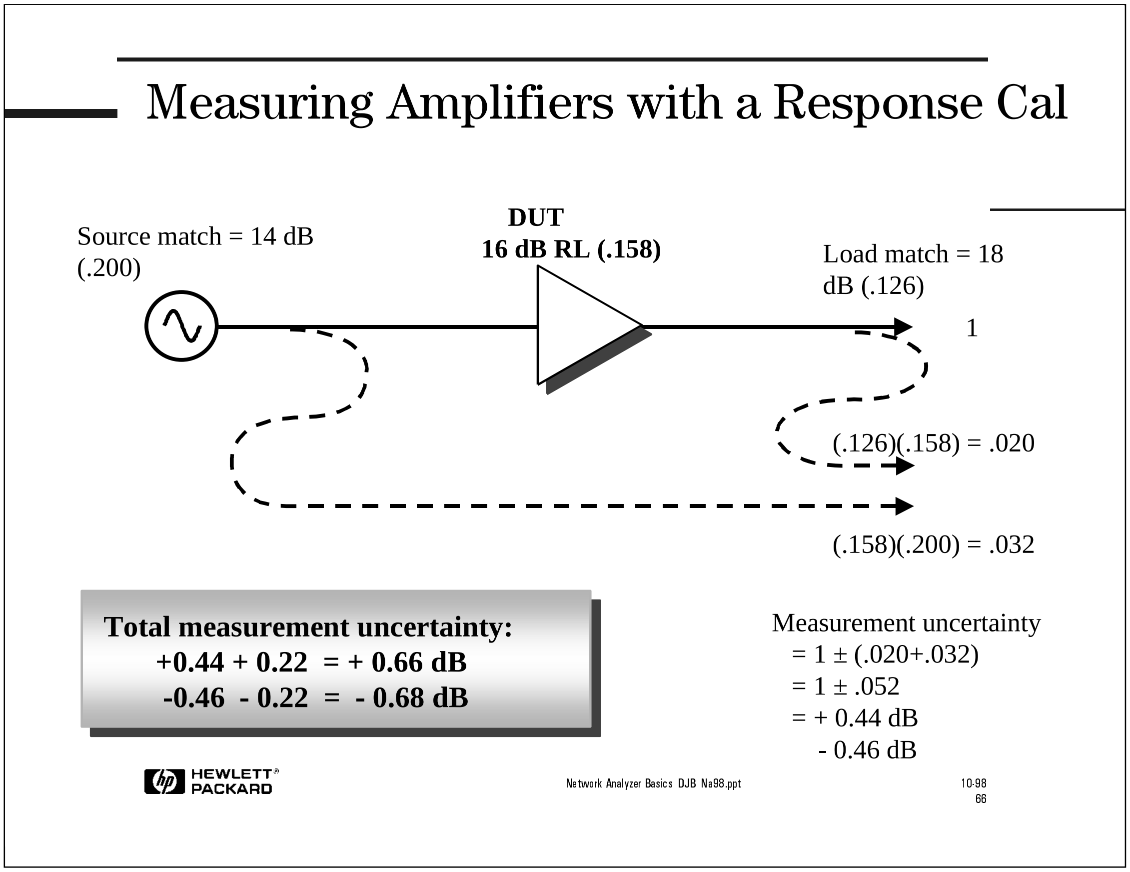
10-98
65

We can further improve transmission measurements by using the enhanced response calibration and by inserting a high-quality attenuator between the output port of the device and test port two of the network analyzer. In this example, we will use a 10 dB attenuator with a SWR of 1.05 (as we did with the reflection example). This makes the effective load match of the analyzer 28.7 dB, about a 10 dB improvement. Our calibration error is minuscule (±.01 dB), and our total measurement uncertainty has been reduced to ±0.09 dB. This is very close to what can be achieved with two-port error correction. As we have seen, adding a high-quality attenuator to port two of a T/R network analyzer can significantly improve measurement accuracy, with only a modest loss in dynamic range.

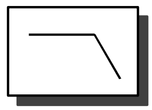
**Slide #66**




Now let's look at an example of measuring an amplifier that has port matches of 16 dB. The match of our test ports remains the same as our initial transmission response example. We see that the middle error term is no longer present, due to the reverse isolation of the amplifier. This fact has reduced our measurement uncertainty to about ± 0.45 dB. Our total measurement error now has been reduced to about ± 0.67 dB, versus the ± 0.85 dB we had when measuring the filter.

**Slide #67**

## Calculating Measurement Uncertainty After a Two-Port Calibration



**DUT**
**1 dB loss (.891)**
**16 dB RL (.158)**

***Corrected error terms:***
***(8753D 1.3-3 GHz Type-N)***

| | | |
|---|---|---|
| Directivity | = | 47 dB |
| Source match | = | 36 dB |
| Load match | = | 47 dB |
| Refl. tracking | = | .019 dB |
| Trans. tracking | = | .026 dB |
| Isolation | = | 100 dB |

**Reflection uncertainty**

$$S_{11m} = S_{11a} \pm (E_D + S_{11a}{}^2 E_S + S_{21a} S_{12a} E_L + S_{11a}(1 - E_{RT}))$$
$$= 0.158 \pm (.0045 + 0.158^2 * .0158 + 0.891^2 * .0045 + 0.158 * .0022)$$
$$= 0.158 \pm .0088 = 16 \text{ dB } \mathbf{+0.53\ dB, -0.44\ dB\ (worst\text{-}case)}$$

**Transmission uncertainty**

$$S_{21m} = S_{21a} \pm S_{21a}(E_I / S_{21a} + S_{11a} E_S + S_{21a} S_{12a} E_S E_L + S_{22a} E_L + (1 - E_{TT}))$$
$$= 0.891 \pm 0.891(10^{-6} / 0.891 + 0.158 * .0158 + 0.891^2 * .0158 * .0045 + 0.158 * .0045 + .003)$$
$$= 0.891 \pm .0056 = 1 \text{ dB } \mathbf{\pm 0.05\ dB\ (worst\text{-}case)}$$

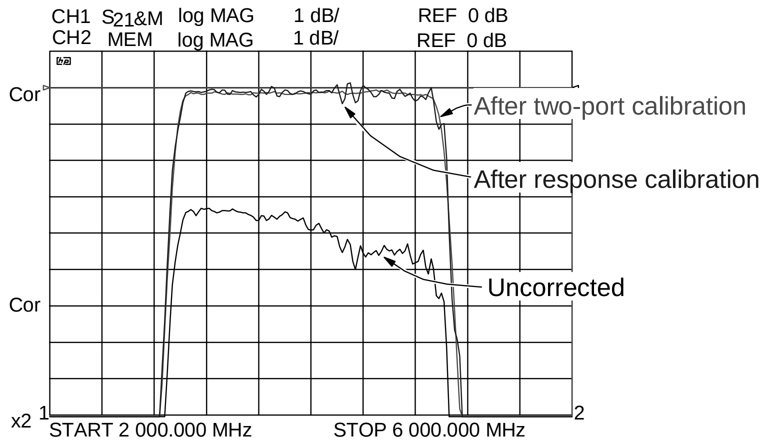
Here is an example of calculating measurement error after a two-port calibration has been done. HP provides values on network analyzer data sheets for effective directivity, source and load match, tracking, and isolation, sometimes for several different calibration kits. The errors when measuring our example filter have been greatly reduced (±0.5 dB reflection error, ±0.05 dB transmission error). Phase errors would be similarly small.

Note that this is a worst-case analysis since we assume that all of the errors would add in-phase. For many narrowband measurements, the error terms will not all align with one another. A less conservative approach to calculating measurement uncertainty would be to use a root-sum-squares (RSS) method. The best technique for estimating measurement uncertainty is to use a statistical approach (which requires knowing or estimating the probability-distribution function of the error terms) and calculating the $\pm 3\sigma$ (sigma) limits.

**Slide #68**

# Response versus Two-Port Calibration

## Measuring filter insertion loss

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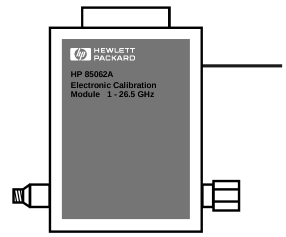
Let's look at some actual measurements done on a bandpass filter with different levels of error correction. The uncorrected trace shows considerable loss and ripple. In fact, the passband response varies about ± 1 dB around filter's center frequency. Is the filter really this bad? No. What we are actually measuring is the sum of the filter's response and that of our test system.

Performing a normalization prior to the measurement of the filter will remove the frequency response of the system (tracking error) from the measurement. The loss that was removed was most likely caused by the test cables. After normalization, the frequency response of the filter still contains ripple caused by an interaction between the system's source and load match. This ripple even goes above the 0 dB reference line, indicating gain! However, we know that a passive device cannot amplify signals. This apparent anomaly is due to mismatch measurement error.

The measurement shown after a two-port calibration is the most accurate of the three measurements shown. Using vector-error correction, the filter's passband response shows variation of about ± 0.1 dB around its center frequency. This increased level of measurement flatness will ensure minimum amplitude distortion, increase confidence in the filter's design, and ultimately increase manufacturing yields due to lower test-failure rates.

**Slide #69**

## ECal: Electronic Calibration (HP 85060 series)

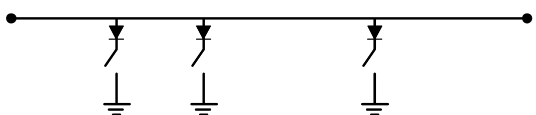


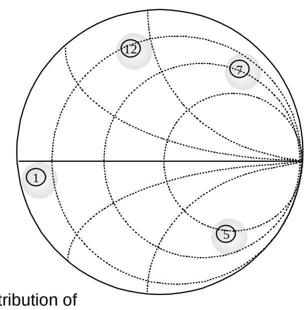

**Impedance States**

- achieved by shunting transmission line with PIN-diode switches in various combinations
- 13 reflective states, from low to high reflection
- two thru states plus one isolation state
- programmable and highly repeatable
- characterized by TRL-calibrated network analyzer

**Calibration**

- four known impedance states presented at each frequency (providing redundant information)
- uses least-squares fit to calculate error terms
- yields accuracy between SOLT and TRL






Example distribution of impedance states for reflection calibration at one frequency

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Although the previous slide showed mechanical calibration standards, HP offers a solid-state calibration solution which makes two-port calibration fast, easy, and less prone to operator errors. This system consists of a control unit and various calibration modules. The calibration modules are solid-state devices with programmable, repeatable impedance states. These states are characterized at the HP factory using a TRL-calibrated network analyzer, making the ECal modules transfer standards (rather than direct standards). ECal provides accuracy better than SOLT, but somewhat less than TRL.

For the microwave calibration modules, the various impedance states are achieved by PIN-diode switches which shunt the transmission line to ground. The number of diodes and their location vary depending upon the module's frequency range. A multitude of reflection coefficients can be generated by applying various combinations of the shunts. With no shunts, the network acts as a low loss transmission line. High isolation between the ports is obtained by driving several of the PIN shunts simultaneously. Four different states are used to compute the error terms at each frequency point. Four states are used because this gives the best trade-off between high accuracy and the time required for the calibration. With four reflection states, we have four equations but only three unknowns. To achieve the best accuracy from this over-determined set of equations, a least-squares-fit algorithm is used. Adding more impedance states at each frequency point would further improve accuracy but at the expense of more calibration time.

**Slide #70**

# Calibrating Non-Insertable Devices

**When doing a thru cal, normally test ports mate directly**
- cables can be connected directly without an adapter
- result is a zero-length thru

**What is an insertable device?**
- has same type of connector, but different sex on each port
- has same type of sexless connector on each port (e.g. APC-7)

**What is a non-insertable device?**
- one that cannot be inserted in place of a zero-length thru
- has same connectors on each port (type and sex)
- has different type of connector on each port
- (e.g., waveguide on one port, coaxial on the other)


**What calibration choices do I have for non-insertable devices?**
- Use an uncharacterized thru adapter
- Use a characterized thru adapter (modify cal-kit definition)
- Swap equal adapters
- Adapter removal

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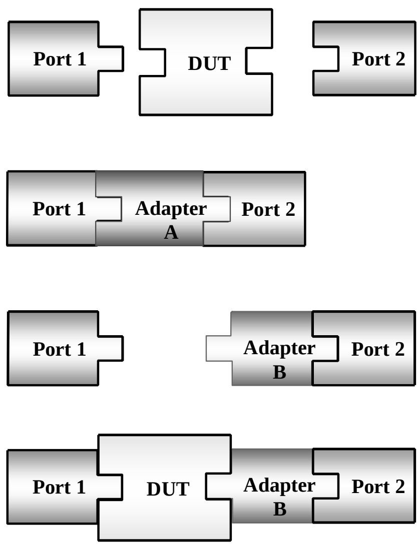
When performing a through calibration, normally the test ports mate directly. For example, two cables with the appropriate connectors can be joined without a through adapter, resulting in a zero-length through path. An insertable device is one that can be substituted for a zero-length through. This device has the same connector type on each port but of the opposite sex, or the same sexless connector on each port, either of which makes connection to the test ports quite simple. A noninsertable device is one that can not be substituted for a zero-length through. It has the same type and sex connectors on each port or a different type of connector on each port, such as 7/16 at one end and SMA on the other end.

There are several calibration choices available for noninsertable devices. The first is to use a characterized through adapter (electrical length and loss specified), which requires modifying the calibration-kit definition. This will reduce (but not eliminate) source and load match errors. A high-quality through adapter (with good match) should be used since reflections from the adapter cannot be removed. The other two choices (swapping equal adapters and adapter removal) will be discussed next.

**Slide #71**

# Swap Equal Adapters Method




***Accuracy depends on how well the adapters are matched - loss, electrical length, match and impedance should all be equal***

1. Transmission cal using adapter A.
2. Reflection cal using adapter B.
3. Measure DUT using adapter B.

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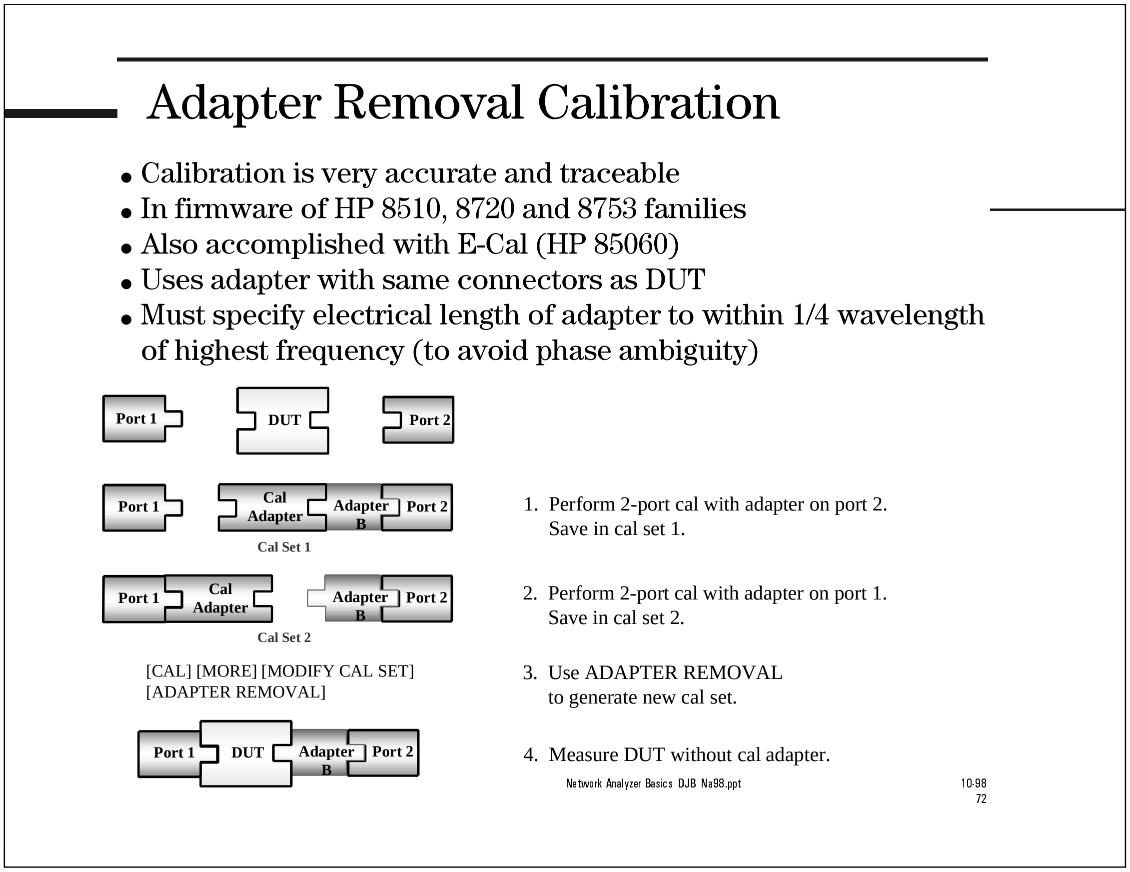
The swap-equal-adapters method is very useful for devices with the same connector type and sex (female SMA on both ends for example). It requires the use of two precision matched adapters that are equal in performance but have connectors of different sexes. For example, for measuring a device with female SMA connectors on both ends using APC-7 mm test cables, the adapters could be 7 mm to male 3.5 mm and 7 mm to female 3.5 mm. To be equal, the adapters must have the same match, characteristic impedance, insertion loss, and electrical delay. Many of Hewlett-Packard's calibration kits include matched adapters for this purpose.

The first step in the swap-equal-adapters method is to perform the transmission portion of a two-port calibration with the adapter needed to make the through connection. This adapter is then removed and the second adapter is used in its place during the reflection portion of the calibration, which is performed on both test ports. This swap changes the sex of one of the test ports so that the DUT can be inserted and measured (with the second adapter still in place) after the calibration procedure is finished. The errors remaining after calibration are equal to the difference between the two adapters. The technique provides a high level of accuracy, but not quite as high as the more complicated adapter-removal technique.

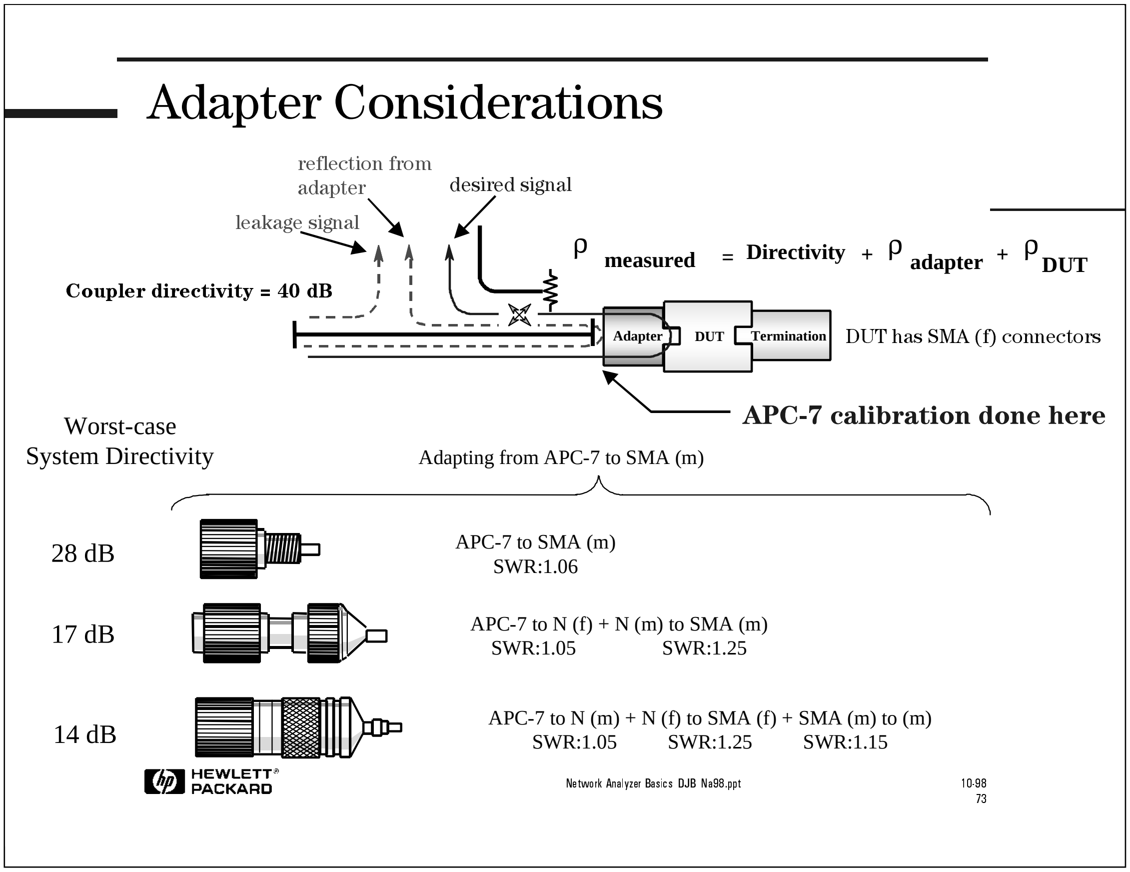
**Slide #72**




Adapter-removal calibration provides the most complete and accurate calibration procedure for noninsertable devices. It is available in the HP 8510, 8720 and 8753 families of vector network analyzers. This method uses a through adapter that has the same connectors as the noninsertable DUT (this adapter is sometimes referred to as the calibration adapter). The electrical length of the adapter must be specified within one-quarter wavelength at each calibration frequency. Type N, 3.5-mm, and 2.4-mm calibration kits for the HP 8510 contain adapters specified for this purpose. For other adapters, the user can simply enter the electrical length.

Two full two-port calibrations are needed for an adapter-removal calibration. In the first calibration, the through adapter is placed on test port two , and the results are saved into a calibration set. In the second calibration, the adapter is moved to test port one and the resulting data is saved into a second calibration set. Two different calibration kits may be used during this process to accommodate devices with different connector types. To complete the adapter-removal calibration, the network analyzer uses the two sets of calibration data to generate a new set of error coefficients that completely eliminate the effects of the calibration adapter. At this point, the adapter can be removed and measurements can be made directly on the DUT.

**Slide #73**




Whenever possible, reflection calibrations should be done with a cal kit that matches the connector type of the DUT. If adapters need to be used to mate the calibrated test system to the DUT, the effect of these adapters on measurement accuracy can be very large. This error is often ignored, which may or may not be acceptable. As the slide shows, the adapter causes an error signal which can add or subtract with the desired signal from the DUT. Worst-case effective directivity (in dB) is now:

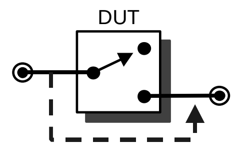
$$-20 \log (\text{Corrected-coupler-directivity} + \rho_{\text{adapters}})$$

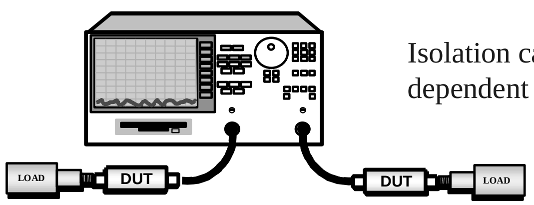
If the adapter has a SWR of say 1.5 (the less-expensive variety), the effective directivity of the coupler drops to around 14 dB worst case, even if the coupler itself had infinite directivity! In other words, with a perfect Zo load ($\rho L = 0$) on the output of the adapter; the reflected signal appearing at the coupled port would only be 14 dB less than the reflection from a short or open circuit. Stacking adapters compounds the problem, as is illustrated above. Consequently, it is very important to use quality adapters (or preferably, no adapters at all) in your measurement system, so system directivity is not excessively degraded. While error-correction can mitigate the effect of adapters on the test port, our system is more susceptible to drift with degraded raw (uncorrected) directivity.

**Slide #74**

## Crosstalk (Isolation)

- Crosstalk definition: signal **leakage** between ports
- Can be a problem with:
  - High-isolation devices (e.g., switch in open position)
  - High-dynamic range devices (some filter stopbands)
- Isolation calibration
  - Adds noise to error model (measuring noise floor of system)
  - Only perform if really needed (use averaging)
  - if crosstalk is **independent** of DUT match, use two terminations
  - if **dependent** on DUT match, use DUT with termination on output




Isolation cal when crosstalk is dependent on match of DUT



When performing a two-port calibration, the user has the option of omitting the part of the calibration that characterizes crosstalk or isolation. The definition of crosstalk is the signal leakage between test ports when no device is present. Crosstalk can be a problem with high-isolation devices (e.g., switch in open position) and high-dynamic range devices (some filter stopbands). The isolation calibration adds noise to the error model since we usually are measuring near the noise floor of the system. For this reason, one should only perform the isolation calibration if it is really needed. If the isolation portion of the calibration is done, trace averaging should be used to ensure that the system crosstalk is not obscured by noise. In some network analyzers, crosstalk can be minimized by using the alternate sweep mode instead of the chop mode (the chop mode makes measurements on both the reflection (A) and transmission (B) channels at each frequency point, whereas the alternate mode turns off the reflection receiver during the transmission measurement).

The best way to perform an isolation calibration is by placing the devices that will be measured on each test port of the network analyzer, with terminations on the other two device ports. Using this technique, the network analyzer sees the same impedance versus frequency during the isolation calibration as it will during subsequent measurements of the DUT. If this method is impractical (in test fixtures, or if only one DUT is available, for example), than placing a terminated DUT on the source port and a termination on the load port of the network analyzer is the next best alternative (the DUT and termination must be swapped for the reverse measurement). If no DUT is available or if the DUT will be tuned (which will change its port matches), then terminations should be placed on each network analyzer test port for the isolation calibration.

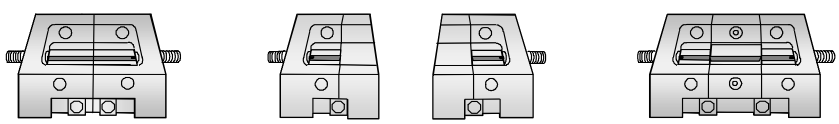
**Slide #75**

# Thru-Reflect-Line (TRL) Calibration

We know about Short-Open-Load-Thru (SOLT) calibration...
What is TRL?

- A two-port calibration technique
- Good for noncoaxial environments (waveguide, fixtures, wafer probing)
- Uses the same 12-term error model as the  more common SOLT cal
- Uses practical calibration standards that are easily fabricated and characterized
- Two variations: TRL (requires 4 receivers) and TRL* (only three receivers needed)
- Other variations: Line-Reflect-Match (LRM), Match (TRM), plus many others

TRL was developed for non-coaxial microwave measurements



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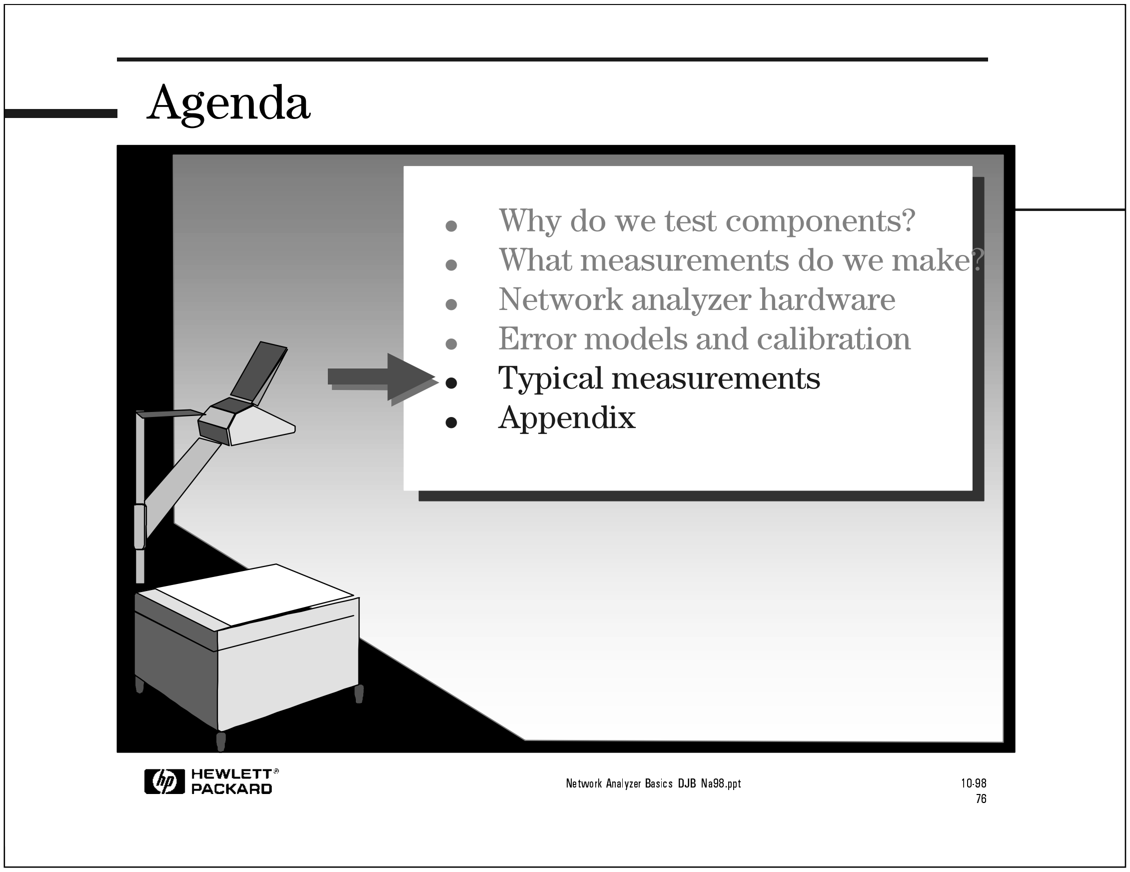
10-98
75

So far, we have discussed coaxial calibration techniques.  Let's briefly look at TRL, a calibration that is especially useful for the noncoaxial environment.

When performing a two-port calibration, we have some choices based on the type of calibration standards we want to use. The second-most common type of two-port calibration (after SOLT) is thru-reflect-line (TRL). TRL is a two-port calibration technique that is primarily used in noncoaxial environments (waveguide, fixtures, wafer probing). It solves for the same 12 error terms as the more common SOLT calibration, using a slightly different error model. TRL calibration uses practical calibration standards that are easily fabricated and characterized.

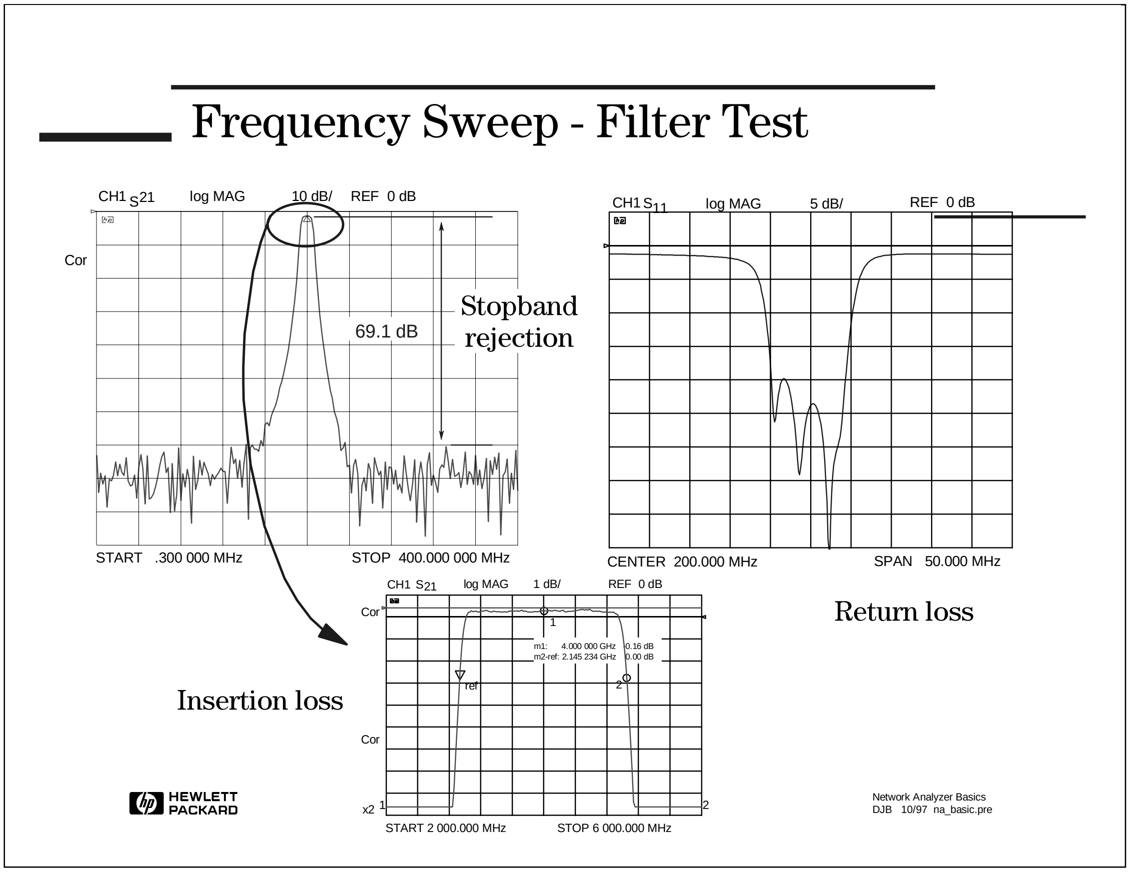
There are two variations of TRL. True TRL calibration requires a 4-receiver network analyzer. The version for three-receiver analyzers is called TRL*.    Other variations of this type of calibration (that share a common error model) are Line-Reflect-Line (LRL), Line-Reflect-Match (LRM), Thru-Reflect-Match (TRM), plus many others.

**Slide #76**




This section will cover some typical measurements. We will look at swept-frequency testing of a filter and swept-power testing of an amplifier.

**Slide #77**



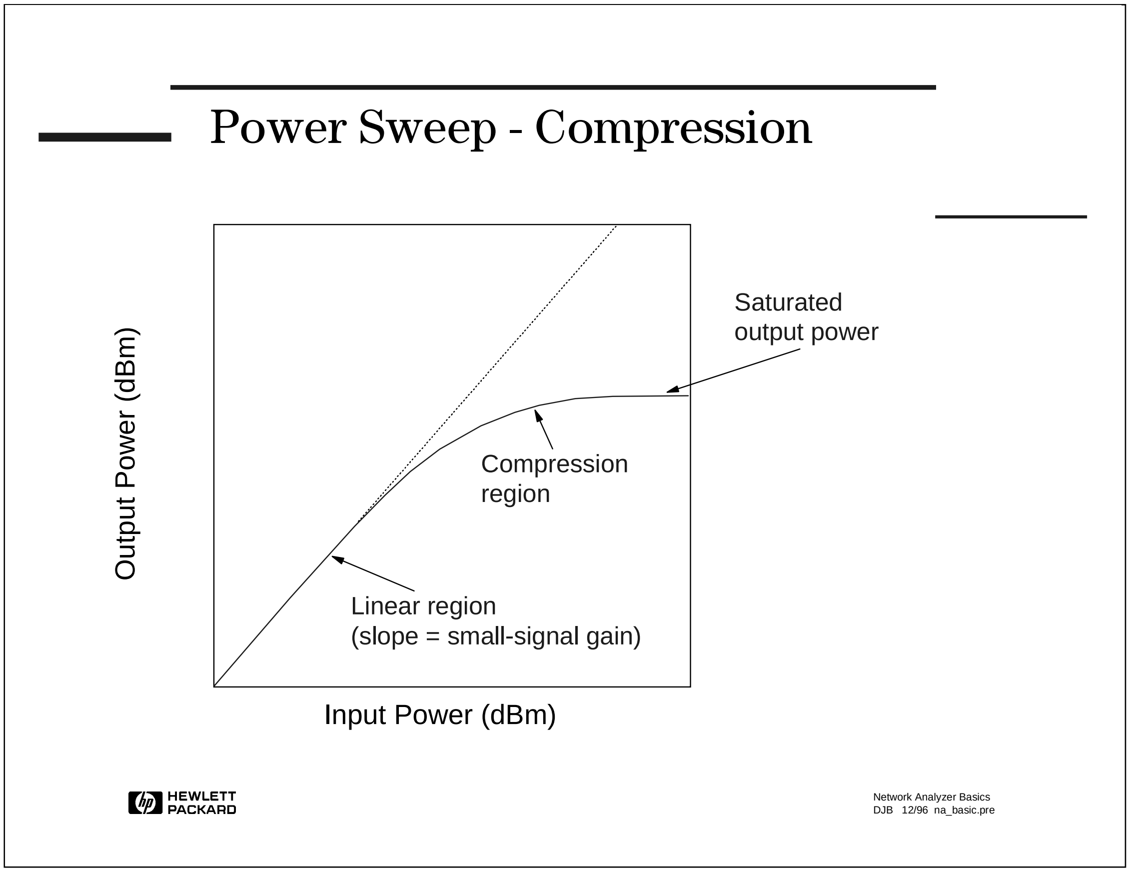

Shown above are the frequency responses of a filter. On the left and bottom we see the transmission response in log magnitude format, and on the right we see the reflection response (return loss).

The most commonly measured filter characteristics are insertion loss and bandwidth, shown on the lower plot with an expanded vertical scale. Another common parameter we might measure is out-of-band rejection. This is a measure of how well a filter passes signals within its bandwidth while simultaneously rejecting all other signals outside of that same bandwidth. The ability of a test system to measure out-of-band rejection is directly dependent on its system dynamic-range specification.

The return loss plot is very typical, showing high reflection (near 0 dB) in the stopbands, and reasonable match in the passband. Most passive filters work in this manner. A special class of filters exist that are absorptive in both the passband and stopband. These filters exhibit a good match over a broad frequency range.

For very narrowband devices, such as crystal filters, the network analyzer must sweep slow enough to allow the filter to respond properly. If the default sweep speed is too fast for the device, significant measurement errors can occur. This can also happen with devices that are electrically very long. The large time delay of the device can result in the receiver being tuned to frequencies that are higher than those coming out of the device, which also can cause significant measurement errors.

**Slide #78**

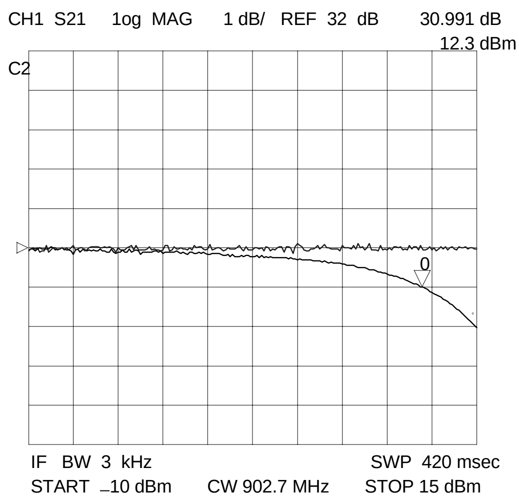



Many network analyzers have the ability to do power sweeps as well as frequency sweeps. Power sweeps help characterize the nonlinear performance of an amplifier. Shown above is a plot of an amplifier's output power versus input power at a single frequency. Amplifier gain at any particular power level is the slope of this curve. Notice that the amplifier has a linear region of operation where gain is constant and is independent of power level. The gain in this region is commonly referred to as "small-signal gain". At some point as the input power is increased, the amplifier gain appears to decrease, and the amplifier is said to be in compression. Under this nonlinear condition, the amplifier output is no longer sinusoidal -- some of the output power is present in harmonics, rather than occurring only at the fundamental frequency. As input power is increased even more, the amplifier becomes saturated, and output power remains constant. At this point the amplifier gain is essentially zero, since further increases in input power result in no change in output power. Saturated output power can be read directly from the above plot.

In order to measure the saturated output power of an amplifier, the network analyzer must be able to provide a power sweep with sufficient output power to drive the amplifier from its linear region into saturation. A preamp at the input of the amplifier under test may be necessary to achieve this.

**Slide #79**

# Power Sweep -Gain Compression




- **1 dB compression**: input power resulting in 1 dB ***drop*** in gain
- Ratioed measurement
- Output power available (non-ratioed measurement)

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DJB 12/96 na_basic.pre

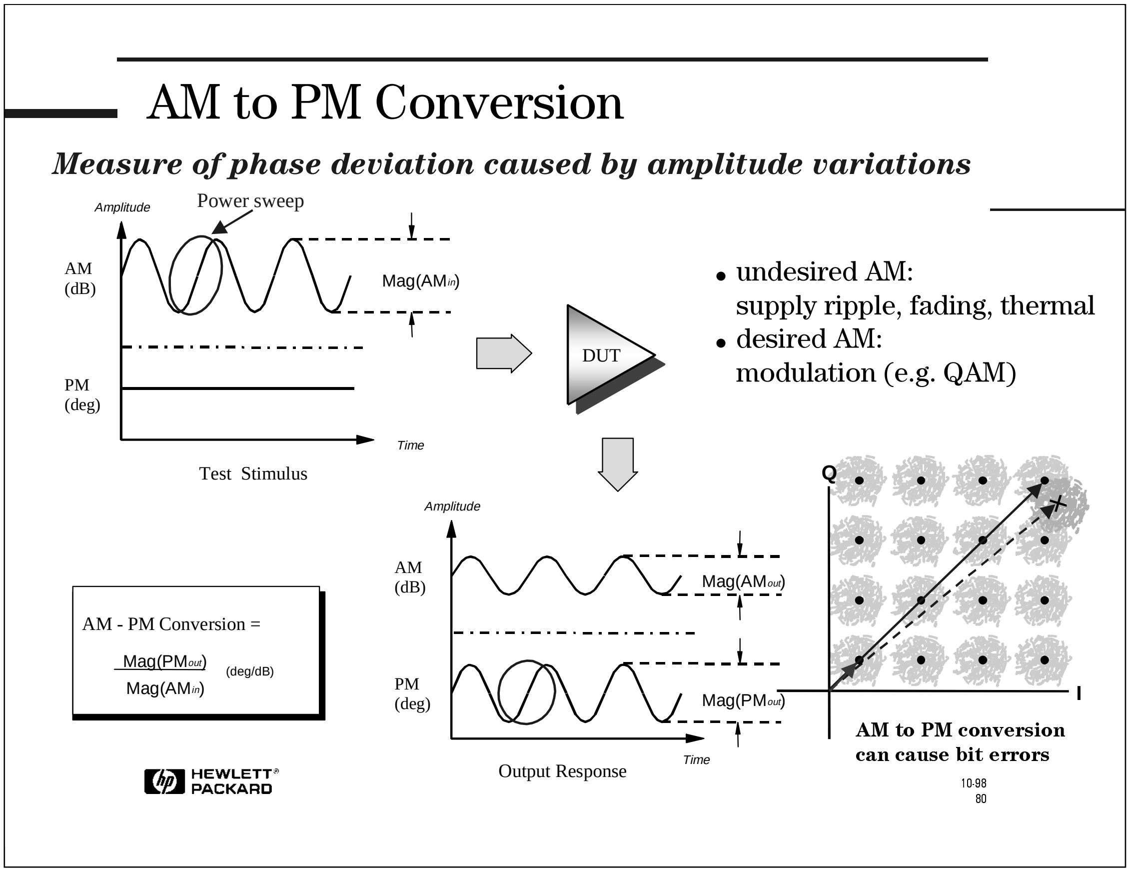
The most common measurement of amplifier compression is the 1-dB-compression point, defined here as the input power* which results in a 1-dB decrease in amplifier gain (referenced to the amplifier's small-signal gain). The easiest way to measure the 1-dB-compression point is to directly display normalized gain (B/R) from a power sweep. The flat part of the trace is the linear, small-signal region, and the curved part on the right side corresponds to compression caused by higher input power. As shown above, the 1-dB-compression point of the amplifier-under-test is 12.3 dBm, at a CW frequency of 902.7 MHz.

It is often helpful to also know the output power corresponding to the 1-dB-compression point. Using the dual-channel feature found on most modern network analyzers, absolute power and normalized gain can be displayed simultaneously. Display markers can read out both the output power and the input power where 1-dB-compression occurs. Alternatively, the gain of the amplifier at the 1-dB-compression point can simply be added to the 1-dB-compression power to compute the corresponding output power. As seen above, the output power at the 1-dB-compression point is 12.3 dBm + 31.0 dB = 43.3 dBm.

It should be noted that the power-sweep range needs to be large enough to ensure that the amplifier under test is driven from its linear region into compression. Modern network analyzers typically provide power sweeps with 15 to 20 dB of range, which is more than adequate for most amplifiers. It is also very important to sufficiently attenuate the output of high-power amplifiers to prevent damage to the network analyzer's receiver.

* The 1-dB-compression point is sometimes defined as the *output* power resulting in a 1-dB decrease in amplifier gain (as opposed to the *input* power).

**Slide #80**




Another common measurement which helps characterize the nonlinear behavior of amplifiers is AM-to-PM conversion, which is a measure of the amount of undesired phase deviation (the PM) which is induced by amplitude variations inherent in the system (the AM). In a communications system, this unwanted PM is caused by unintentional amplitude variations such as power supply ripple, thermal drift, or multipath fading, or by intentional amplitude change that is a result of the type of modulation used, such as the case with QAM or burst modulation.

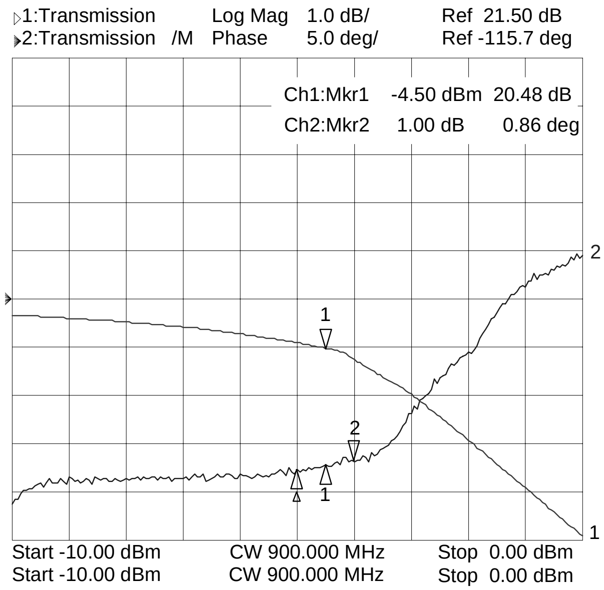
AM-to-PM conversion is a particularly critical parameter in systems where phase (angular) modulation is employed, because undesired phase distortion causes analog signal degradation, or increased bit-error rates (BER) in digital systems. Examples of common modulation types that use phase modulation are FM, QPSK, and 16QAM. While it is easy to measure the BER of a digital communication system, this measurement alone does not provide any insight into the underlying phenomena which cause bit errors. AM-to-PM conversion is one of the fundamental contributors to BER, and therefore it is important to quantify this parameter in communication systems.

The I/Q diagram shown above shows how AM-to-PM conversion can cause bit errors. Let's say the desirable state change is from the small solid vector to the large solid vector. With AM-PM conversion, the actual large vector may be as shown with the dotted line. This is due to phase shift that results from a change in power level. For a 64QAM signal as shown (only one quadrant is drawn), we see that the noise circles that surround each state actually overlap, which means that statistically, some bit errors will occur.

AM-to-PM conversion is usually defined as the change in output phase for a 1-dB increment in the input power to an amplifier, expressed in degrees-per-dB (°/dB). An ideal amplifier would have no interaction between its phase response and the level of the input signal.

**Slide #81**

# Power Sweep - AM to PM Conversion




- Use transmission setup with a power sweep
- Display phase of S21
- AM - PM = 0.86 deg/dB

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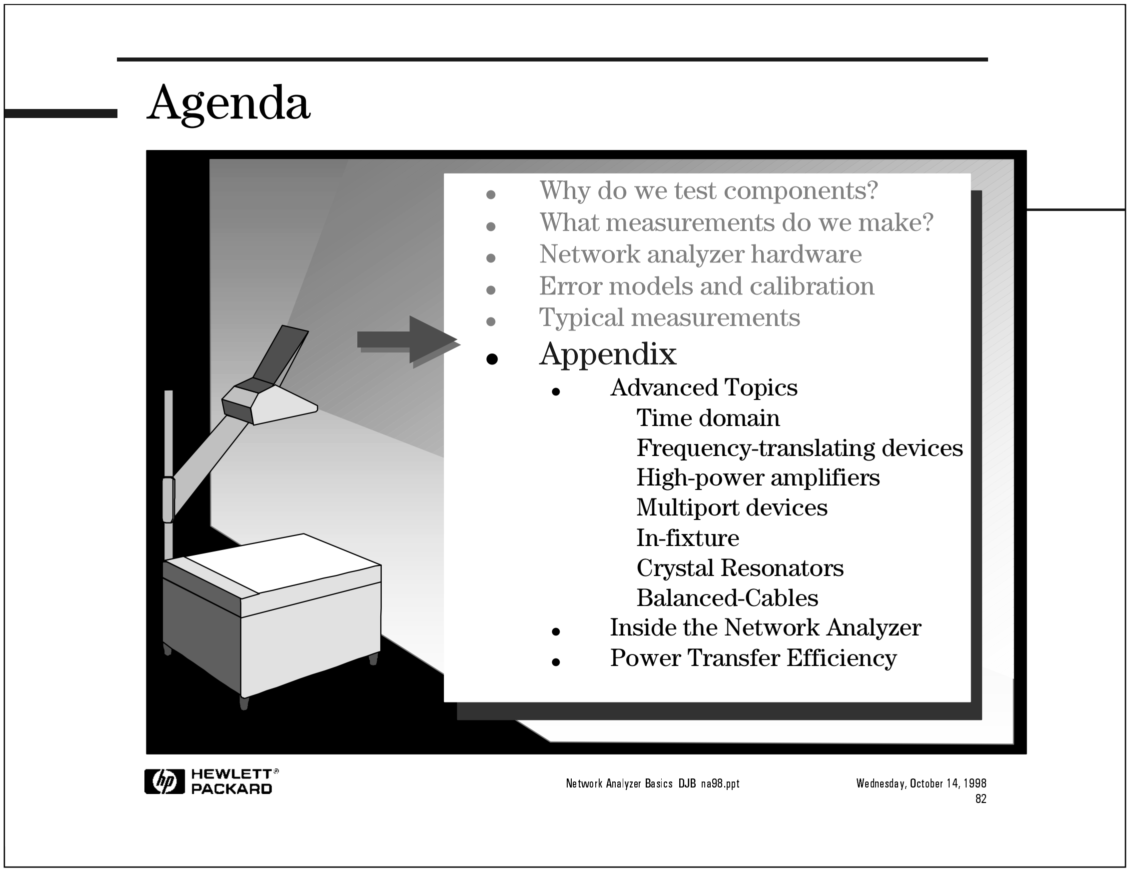
AM-PM conversion can be measured by performing a power sweep with a vector network analyzer, using the same transmission setup that we used for gain compression. The displayed data is formatted as the phase of $S_{21}$ (transmission) versus power. AM-PM conversion can be computed by choosing a small amplitude increment (typically 1 dB) centered at a particular RF power level, and noting the resultant change in phase. The easiest way to read out the amplitude and phase deltas is to use trace markers. Dividing the phase change by the amplitude change yields AM-PM conversion. The plot above shows AM-PM conversion of 0.86 °/dB, centered at an input power of -4.5 dBm and an output power of 16.0 dBm. Had we chosen to measure AM-PM conversion at a higher power level, we would have seen a much larger value (around 7 °/dB).

**Slide #82**

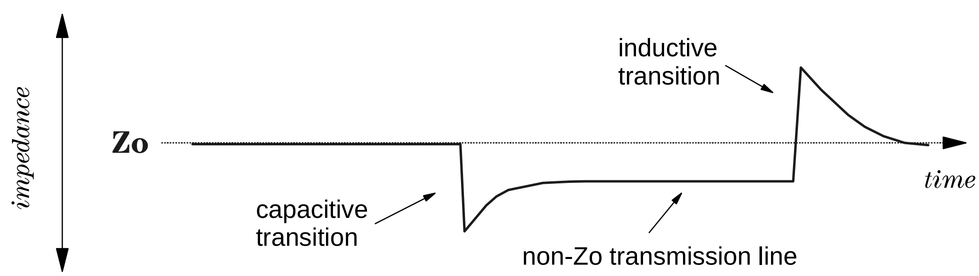



The Advanced Topics section is intended to provide pointers to reference material which covers these topics in more detail. Inside the Network Analyzer section provides a more detailed discussion about the signal separation devices and receiver sections of the network analyzer.

**Slide #83**

## Time-Domain Reflectometry (TDR)

- what is TDR?
  - time-domain reflectometry
  - analyze impedance versus time
  - distinguish between inductive and capacitive transitions
- with gating:
  - analyze transitions
  - analyzer standards






Time-domain reflectometry is a very useful tool that allows us to measure impedance versus distance. One good application for TDR is fixture design and the design of corresponding in-fixture calibration standards. We can distinguish between capacitive and inductive transitions, and see non-Zo transmission lines. TDR can help us determine the magnitude and position of reflections from transitions within the fixture, and we can measure the quality of the calibration standards. As long as we have enough spatial resolution, we can see the reflections of the connector launches independently from the reflections of the calibration standards. It is very easy to determine which transition is which, as the designer can place a probe on a transition and look for a large spike on the TDR trace.

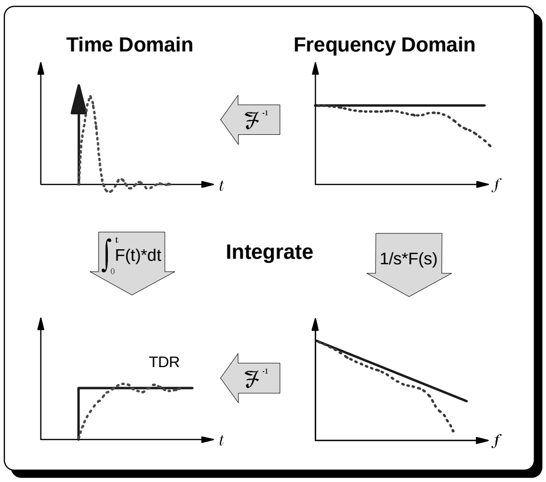
With time-domain gating, we can isolate various sections of the fixture and see the effects in the frequency domain. For example, we can choose to look at just the connector launches (without interference from the reflections of the calibration standards), or just the calibration standards by themselves.

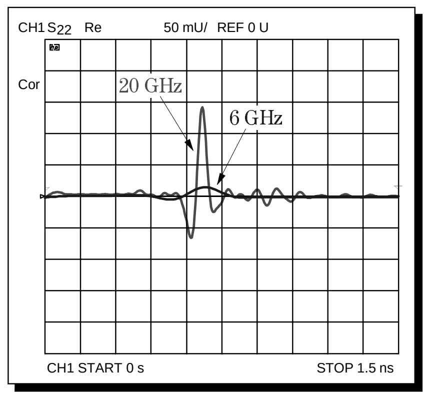
Another application for TDR is fault-location for coaxial cables in cellular and CATV installations. We can use TDR in these cases to precisely determine the location of cable faults such as crimps, poor connections, shorts, opens – anything that causes a portion of the incident signal to be reflected.

**Slide #84**

## TDR Basics Using a Network Analyzer

- start with broadband frequency sweep (often requires microwave VNA)
- use inverse-Fourier transform to compute time-domain
- resolution inversely proportionate to frequency span




Network Analyzer Basics
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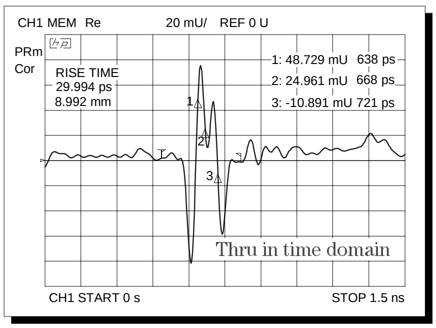
TDR measurements using a vector network analyzer start with a broadband sweep in the frequency domain. The inverse-Fourier transform is used to transform frequency-domain data to the time domain. The figure on the left of the slide shows a simplified conceptual model of how a network analyzer derives time-domain traces. For step-response TDR, we want to end up at the lower left-hand plot. The network analyzer gathers data in the frequency domain (upper right) from a broadband sweep (note: all the data is collected from a reflection measurement). In effect, we are stimulating the DUT with a flat frequency input, which is equivalent to an impulse in the time domain. The output response of our DUT is therefore the frequency response of its impulse response. Since a step in the time domain is the integral of an impulse, if we integrate the frequency-response data of our DUT, we will have frequency-domain data corresponding to the step response in the time domain. Now, we simply perform an inverse-Fourier transform to get from the frequency domain to the time domain, and *voilá*, we have the step response . Note that we could also perform the inverse-Fourier transform first, and then integrate the time-domain data. The result would be the same. The actual math used in the network analyzer is somewhat more complicated than described above, in order to take care of other effects (one example is extrapolating a value for the DC term, since the analyzer doesn't measure all the way down to 0 Hz).

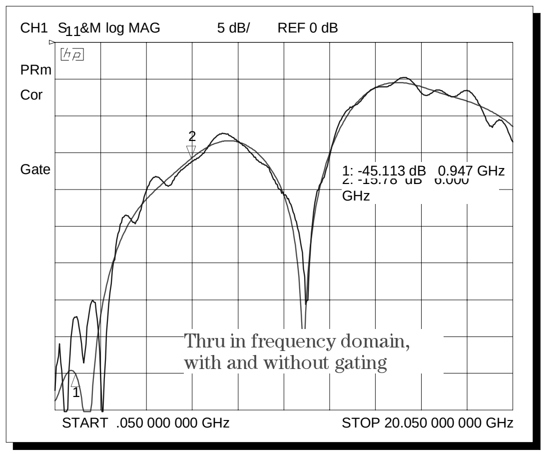
To get more resolution in the time domain (to separate transitions), we need a faster effective rise time for our step response. This translates to a sharper (narrower) effective impulse, which means a broader input-frequency range must be applied to our DUT. In other words, the higher the stop frequency, the smaller the distance that can be resolved. For this reason, it is generally necessary to make microwave measurements on the fixture to get sufficient resolution to analyze the various transitions. Providing sufficient spacing between transitions may eliminate the need for microwave characterization, but can result in very large fixtures. The plot above of a fixtured-load standard shows the extra resolution obtained with a 20 GHz sweep versus only a 6 GHz sweep.

**Slide #85**

## Time-Domain Gating

- TDR and gating can **remove** undesired reflections (a form of error **correction)**
- Only useful for **broadband** devices (a load or thru for example)
- Define **gate** to only include DUT
- Use two-port calibration




Network Analyzer Basics
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Gating can be used in conjunction with time-domain measurements to separate and remove undesirable reflections from those of interest. For example, gating can isolate the reflections of a DUT in a fixture from those of the fixture itself. This is a form of error correction. For time-domain gating to work effectively, the time domain responses need to be well-separated in time (and therefore distance). The gate itself looks like a filter in time, and has a finite transition range between passing and rejecting a reflection (similar to the skirts of a filter in the frequency domain).

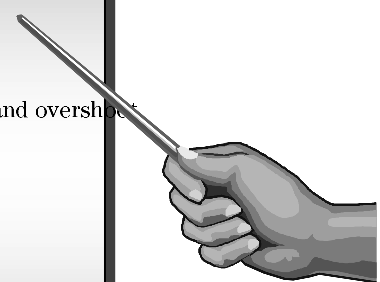
The plots above show the performance of an in-fixture thru standard (without normalization). We see about a 7 dB improvement in return loss at 947 MHz using time-domain gating, resulting in a return loss of 45 dB. The gating effectively removes the effects of the SMA connectors at either end of the test fixture.

**Slide #86**

# Ten Steps for Performing TDR

1. Set up desired frequency range
(need wide span for good spatial resolution)
2. Under SYSTEM, transform menu, press "set freq low pass"
3. Perform one- or two-port calibration
4. Select S11 measurement *
5. Turn on transform (low pass step) *
6. Set format to real *
7. Adjust transform window to trade off rise time with ringing and overshoot *
8. Adjust start and stop times if desired
9. For gating:
   - set start and stop frequencies for gate
   - turn gating on *
   - adjust gate shape to trade off resolution with ripple *
10. To display gated response in frequency domain
   - turn transform off (leave gating on) *
   - change format to log-magnitude *

* *If using two channels (even if coupled), these parameters must be set independently for second channel*



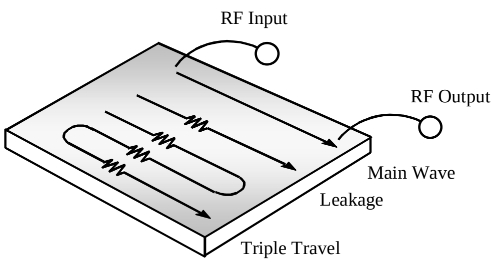
Network Analyzer Basics DJB Na98.ppt

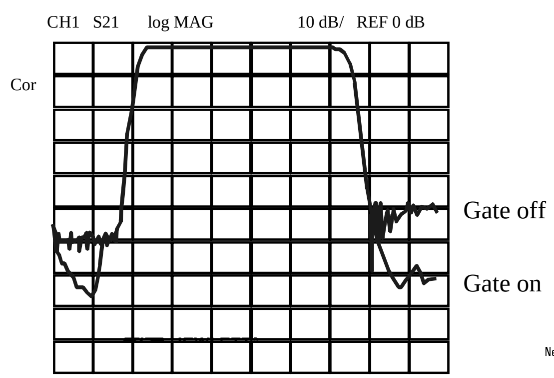
10-98
86

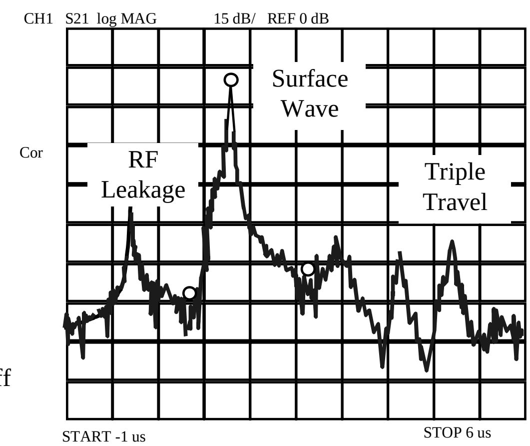
Here is a summary of how to perform TDR measurements. Without such a checklist, it is easy to overlook some of the more subtle steps, resulting in confusing or misleading measurements. A one-port calibration is all that is needed when characterizing connectors and the open, short and load standards. A two-port calibration is needed to characterize the reflection or line impedance of the thru standard.

**Slide #87**

# Time-Domain Transmission

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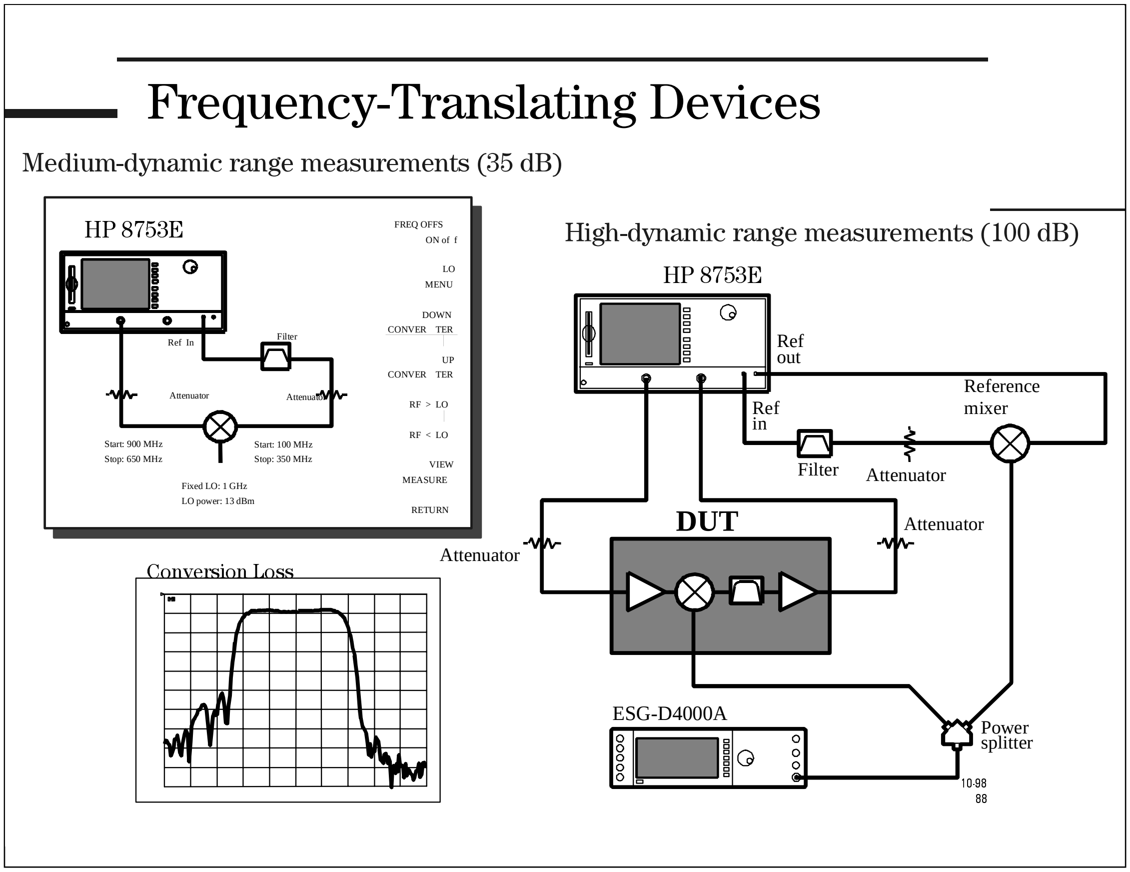
87

Time-domain transmission (TDT) is a similar tool which uses the transmission response instead of the reflection response. It is useful in analyzing signal timing in devices such as SAW filters. Gating is also useful for TDT. In the above example, a designer could look at the frequency response of the main surface wave without the effect of the leakage and triple-travel error signals.

**Slide #88**




More information about measuring mixers and tuners can be obtained from the following sources:

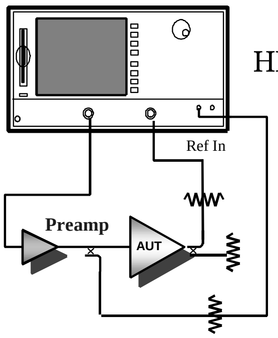
"*Improving Network Analyzer Measurements of Frequency-Translating Devices*", Application Note 1287-7, (Literature Number 5966-3318E, 2/98)

"*Measurement Challenges of Frequency-Translating Devices*", 1995 Device Test Seminar handout, 5963-5191E (12/94)

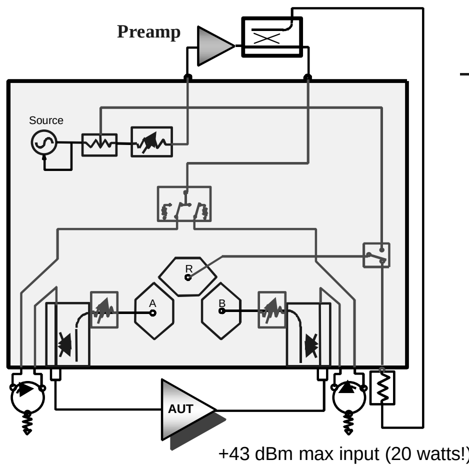
"*Mixer Measurements using the HP 8753 Network Analyzer*", Hewlett-Packard Product Note 8753-2A (5952-2771, 8/90)

**Slide #89**

# High-Power Amplifiers




HP 8753E




HP 8720D Option 085

HP 85118A High-Power Amplifier Test System

Network Analyzer Basics DJB Na98.ppt 10-98
89

More information about measuring high-power amplifiers can be obtained from the following sources:

"*Using a Network Analyzer to Characterize High-Power Devices*," *Application Note 1287-6, 5966-3319E (2/98)*

"*Measurement Solutions for Test Base-Station Amplifiers*", 1996 Device Test Seminar handout, 5964-9803E (4/96)

"*Modern Solutions for Testing RF Communications Amplifiers*", 1995 Device Test Seminar handout, 5963-5191E (12/94)

"*Amplifier Measurements using the HP 8753 Network Analyzer*", Hewlett-Packard Product Note 8753-1 (5956-4361, 5/88)

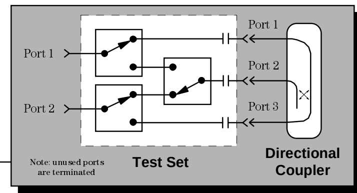
"*Testing Amplifiers and Active Devices with the HP 8720 Network Analyzer*", Hewlett-Packard Product Note 8720-1 (5091-1942E, 8/91)

"*HP 85108 Series Network Analyzer Systems for Isothermal, High-Power, and Pulsed Applications*", Product Overview, (5091-8965E, '94)

"*HP 85118 Series High Power Amplifier Test System*", Product Overview (5963-9930E, 5/95)

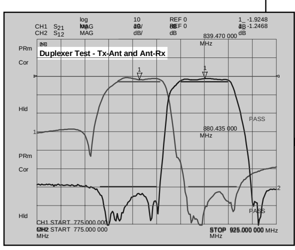
**Slide #90**

# Multiport Device Test




Multiport test sets:

- improve **throughput** by reducing the number of connections to DUTs with more than 2 ports
- allow **simultaneous** viewing of two paths (good for tuning duplexers)
- include **mechanical** or **solid-state** switches, **50** or **75** ohms
- degrade raw performance so **calibration** is a ***must*** (use two-port cals whenever possible)

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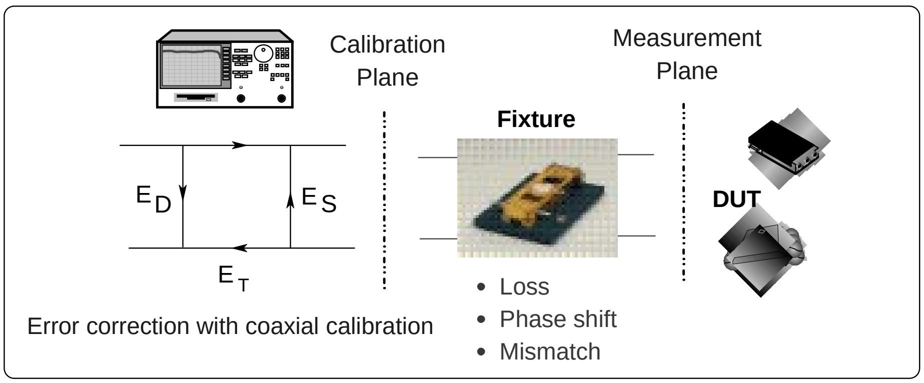
More information about measuring multiport devices can be obtained from the following source:

"*Improve Test Throughput for Duplexers and Other Multiport Devices*", 1996 Device Test Seminar handout, 5964-9803E (4/96)

**Slide #91**

## In-Fixture Measurements

**Measurement problem:** coaxial calibration plane is not the same as the in-fixture measurement plane






More information about in-fixture measurements can be obtained from the following sources:

"*Designing and Calibrating RF Fixtures for Surface-Mount Devices*", 1996 Device Test Seminar handout, 5964-9803E (4/96)

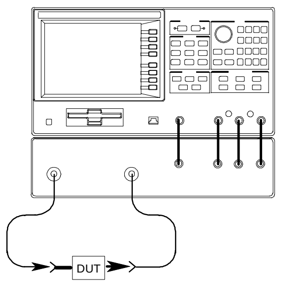
"*Accurate Measurements of Packaged RF Devices*", 1995 Device Test Seminar handout, 5963-5191E (12/94)

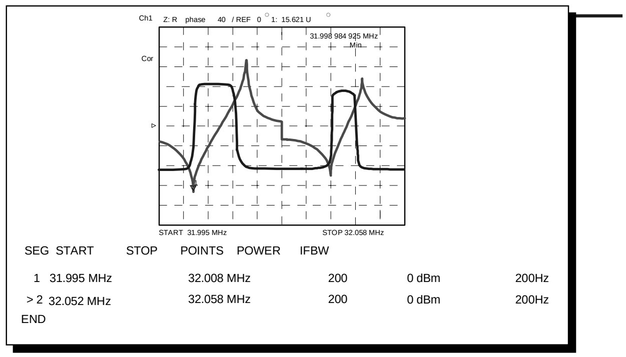
"*Specifying calibration standards for the HP 8510 network analyzer*", Product Note 8510-5A, 08510-90352 (1/88)

"*Applying the HP 8510 TRL calibration for non-coaxial measurements*", Product Note 8510-8A, 5091-3645E (2/92)

## Slide #92

# Characterizing Crystal Resonators/Filters




Example of crystal resonator measurement

HP E5100A Network Analyzer

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DJB 12/96 na_basic.pre

More information about measuring crystal resonators and filters can be obtained from the following sources:

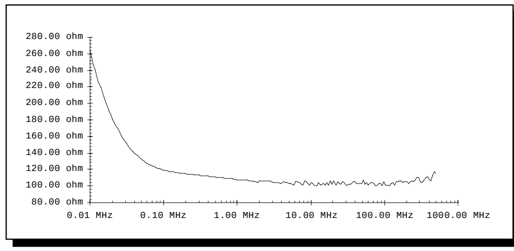
"*Crystal Resonators Measuring Functions of HP E5100A/B Network Analyzer*", Hewlett-Packard Product Note, (5965-4972E)

"*HP E5100A/B Network Analyzer*", Product Overview, (5963-3991E)

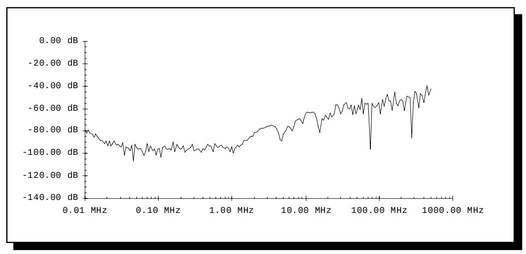
**Slide #93**

# RF Balanced-Cable Measurements

HP 4380S RF Balanced-Cable Test System




Example of characteristic impedance (Zc) measurement from 10 kHz to 500 MHz




Example of near-end crosstalk (NEXT) measurement

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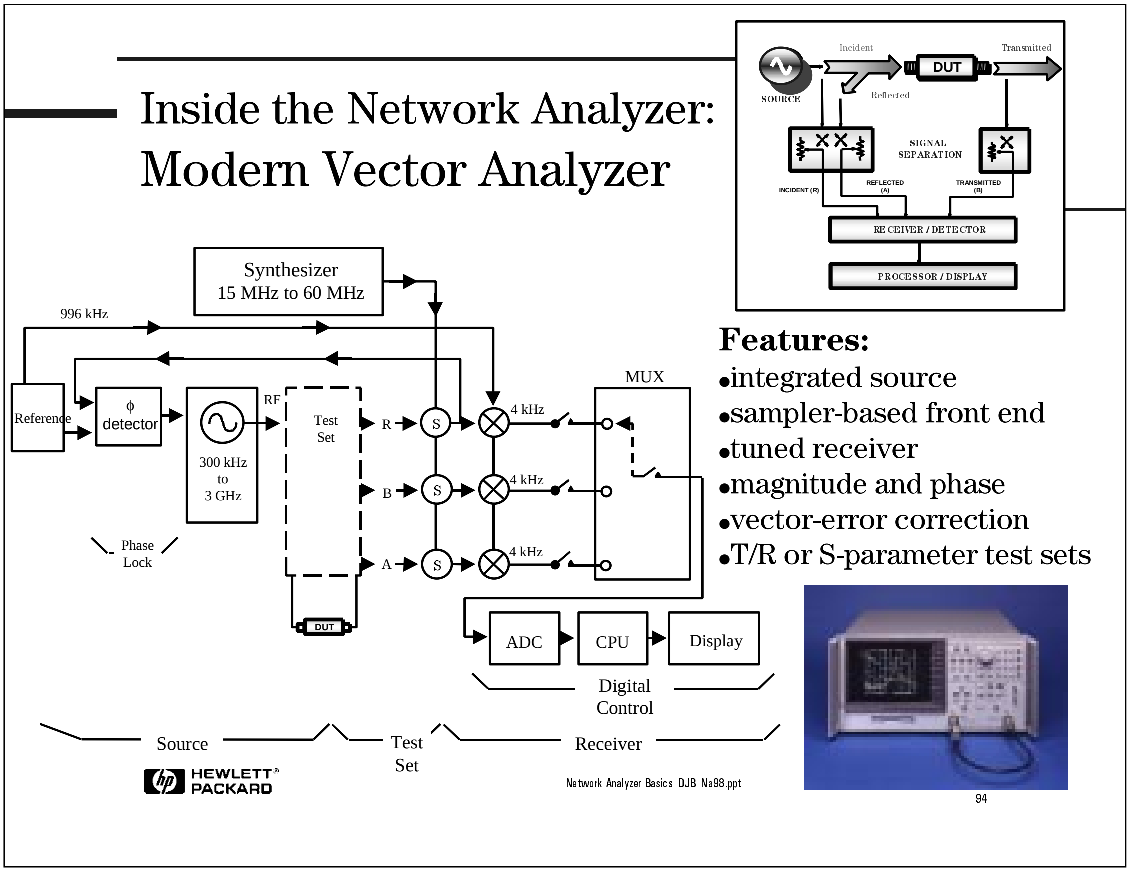
More information about measuring RF balanced cables can be obtained from the following source:

"*HP 4380S RF Balanced Cable Test System*", Product Overview, (5964-2391E)

**Slide #94**



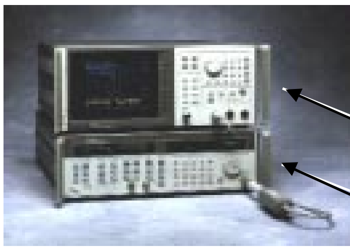

This section is provided to complement the Network Analyzer Hardware Section. It gives more detailed information about the internal hardware, specifically, the signal separation devices and receiver section.

Here is a block diagram of a modern vector network analyzer. It features an integrated source, a sampler-based front end, and a tuned receiver providing magnitude and phase data with vector-error correction. The test set (the portion of the instrument that contains the signal-separation devices and the switches for directing the RF power) can either be transmission/reflection (T/R) based or an S-parameter test set.

Modern scalar analyzers like the 8711C and 8713C look very similar to this block diagram, but they don't display phase information on the screen. Internally, however, they are essentially vector analyzers. This capability lets them make much more accurate measurements than traditional scalar analyzers.

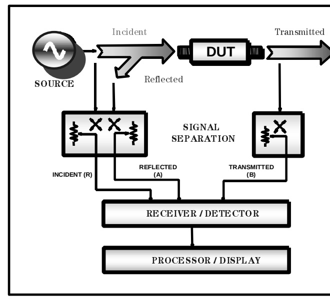
**Slide #95**

# Traditional Scalar Analyzer



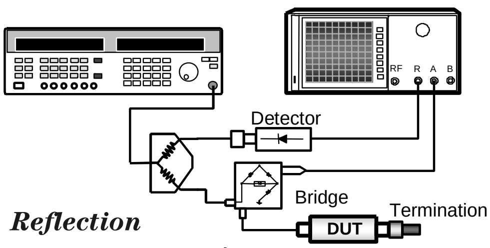
processor/display

source

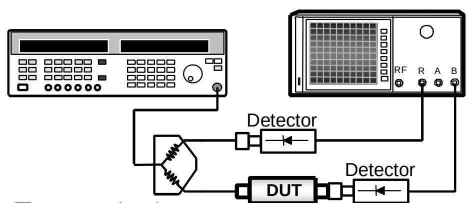



Example: **HP 8757D**

- requires external detectors, couplers, bridges, splitters
- good for low-cost microwave scalar applications




***Reflection***




***Transmission***

Network Analyzer Basics DJB Na98.ppt

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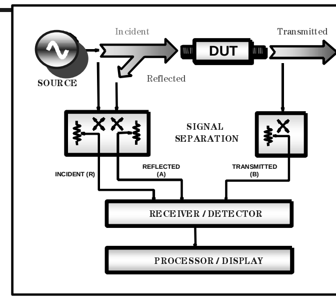
95

Here is a picture of a traditional scalar system consisting of a processor/display unit and a stand-alone source (HP 8757D and HP 8370B). This type of system requires external splitters, couplers, detectors, and bridges. While not as common as they used to be, scalar systems such as this are good for low-cost microwave scalar applications.

The configuration shown for reflection provides the best measurement accuracy (assuming the termination is a high-quality Zo load), especially for low-loss, bidirectional devices (i.e., devices that have low loss in both the forward and reverse directions). Alternately, the transmission detector can be connected to port two of the DUT, allowing both reflection and transmission measurements with a single setup. The drawback to this approach is that the detector match (which is considerably worse than a good load) will cause mismatch errors during reflection measurements.

**Slide #96**

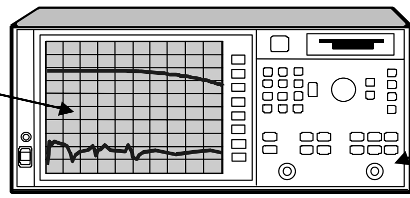
# Modern Scalar Analyzer




**Everything necessary for transmission and reflection measurements is *internal!***

One-port (reflection) and response (transmission) calibrations

Large display



Narrowband *and* broadband detectors

Synthesized source

Transmission/reflection test set

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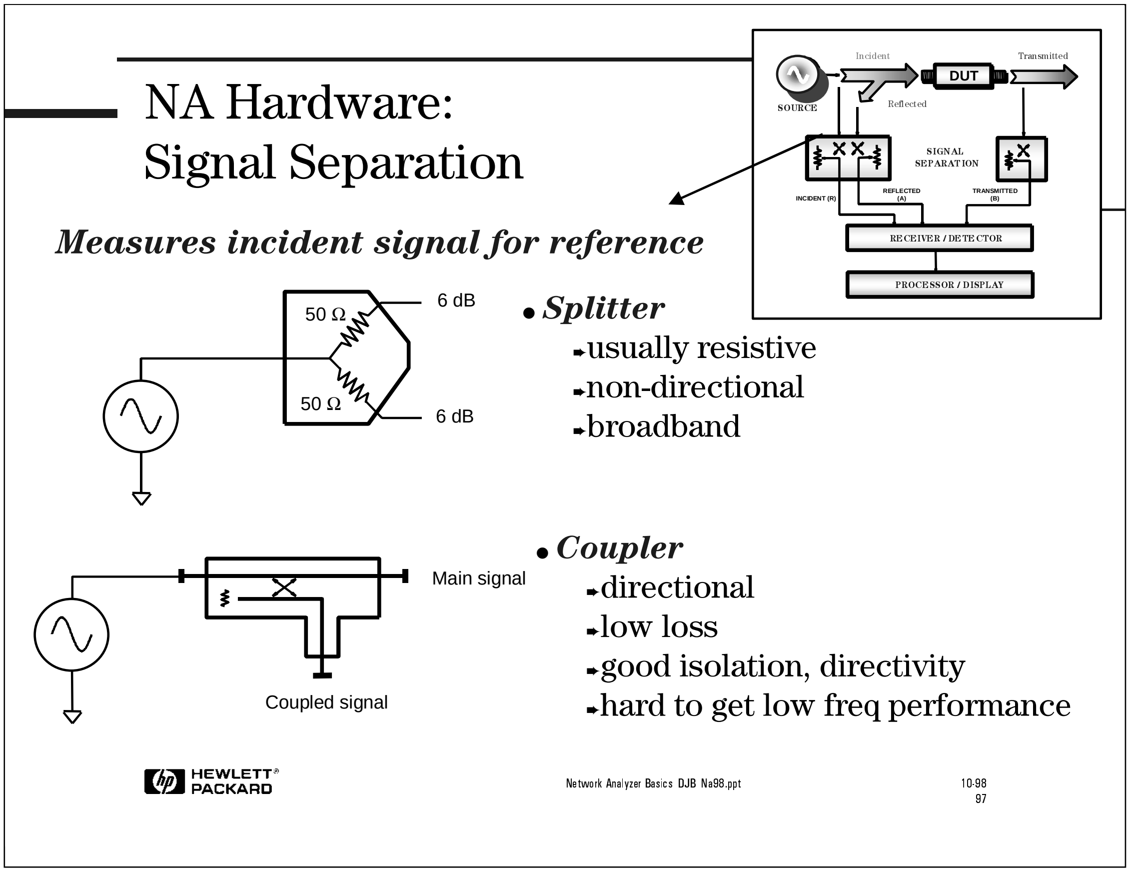
10-98
96

A modern scalar network analyzer has all of the components needed for reflection and transmission measurements built-in to the instrument, such as a synthesized source, a test set, and a large display. Some scalar network analyzers even have attributes of vector analyzers such as narrowband detection and vector error correction (one-port and enhanced-response calibration). These instruments exhibit very high dynamic range and good measurement accuracy.

**Slide #97**




In this section, we will cover the signal separation block. The hardware used for this function is generally called the "test set". The test set can be a separate box or integrated within the network analyzer.

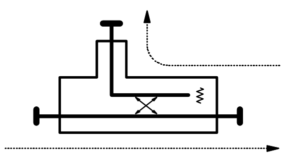
There are two functions that our signal-separation hardware must provide. The first is to measure a portion of the incident signal to provide a reference for ratioing. This can be done with splitters or directional couplers. Splitters are usually resistive. They are non-directional devices (more on directionality later) and can be very broadband. The trade-off is that they usually have 6 dB or more of loss in each arm.

Directional couplers can be built with very low loss (through the main arm) and good isolation and directivity. However, it is hard to make them operate at low frequencies. This can be a problem in RF network analyzers, where low frequency coverage is important.

**Slide #98**

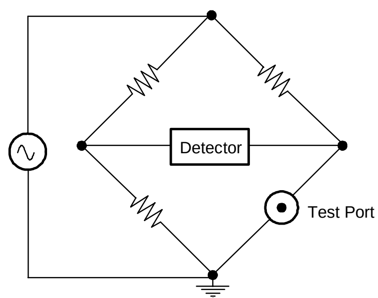
## Signal Separation

***Separating incident and reflected signals***



- ***Coupler***
  - directional
  - low loss
  - good isolation, directivity
  - hard to get low freq performance

- ***Bridge***
  - used to measure reflected signals only
  - broadband
  - higher loss

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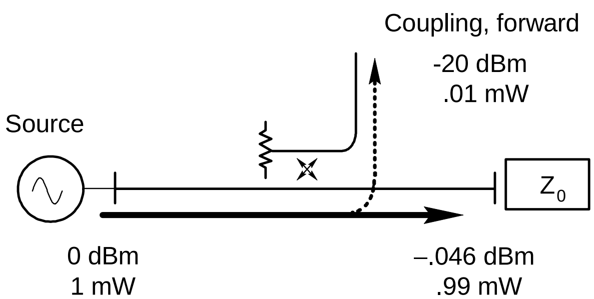
The second function of the signal-splitting hardware is to separate the incident (forward) and reflected (reverse) traveling waves at the input of our DUT. Again, couplers are ideal in that they are directional, have low loss, and high reverse isolation. However, due to the difficulty of making truly broadband couplers, bridges are often used.

Bridges are very useful for measuring reflection because they can be made to work over a very wide range of frequencies. The main trade-off is that they exhibit more loss to the transmitted signal, resulting in less power delivered to the DUT for a given source power.

**Slide #99**

## Forward Coupling Factor




***Example of 20 dB Coupler***

$$\text{Coupling Factor (dB)} = -10 \log \frac{P_{\textit{coupling forward}}}{P_{\textit{incident}}}$$

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The directional coupler measures (couples) a portion of the signal traveling in one direction only. The signal flowing through the main arm is shown as a solid line, and the coupled signal is shown as a dotted line.

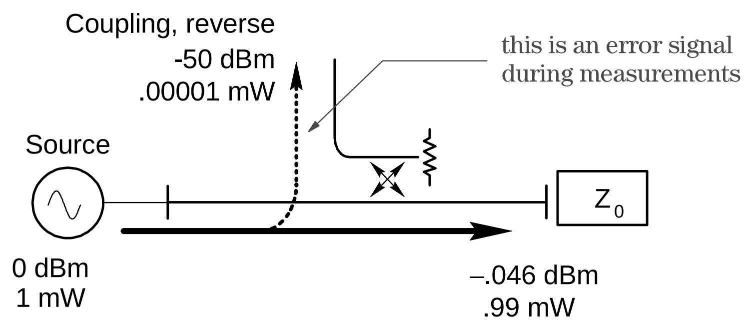
The signal appearing at the coupled port is reduced by an amount known as the coupling factor. This is measured by placing the coupler in the forward direction and measuring the power at the coupled port, relative to the incident power:

Coupling Factor (dB) = -10 log ($P_{fwd-cpl}/P_{in}$)

In this example of a 20 dB directional coupler, the level of a signal at the coupled port is 20 dB below that of the input port. The loss through the main arm is only .046 dB. There are also frequency response terms associated with the main-arm response and the coupling factor, expressed in terms of ± dB.

**Slide #100**

# Directional Coupler Isolation (Reverse Coupling Factor)




***Example of 20 dB Coupler "turned around"***

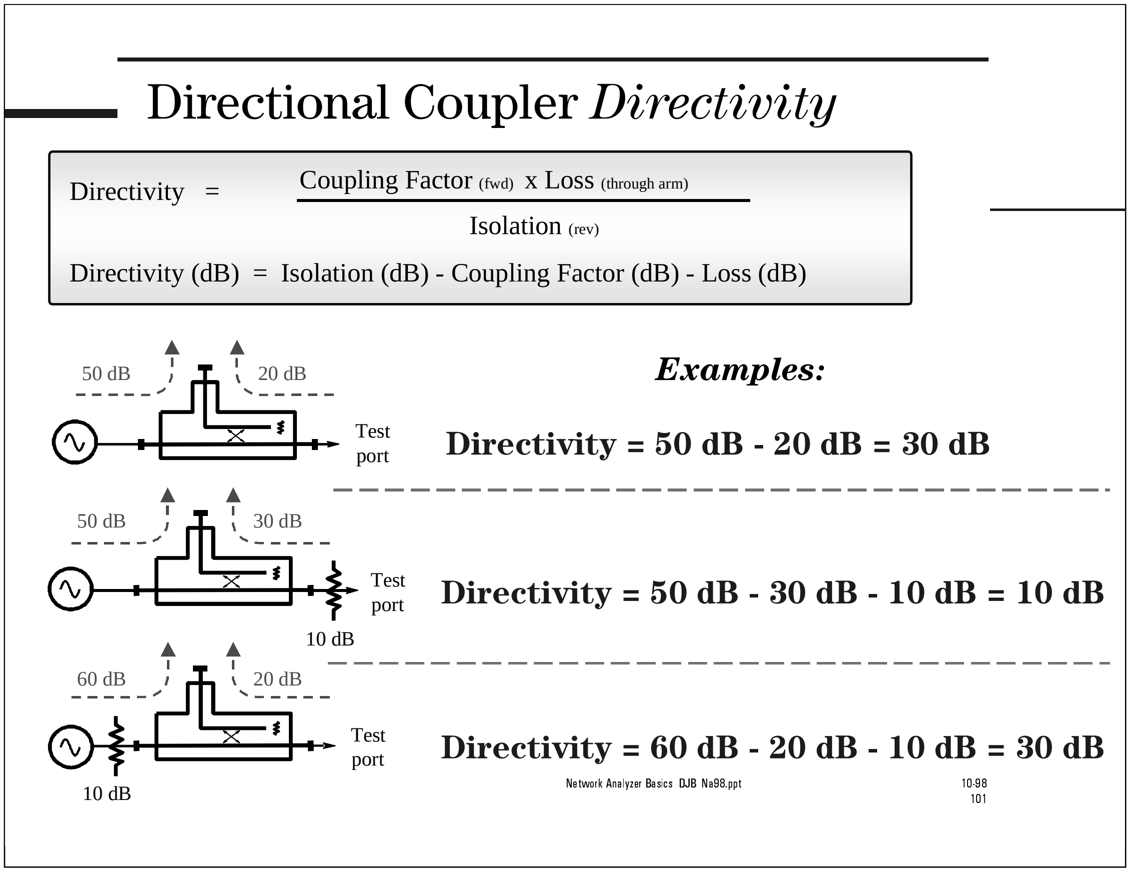
$$\text{Isolation Factor (dB)} = -10 \log \frac{P_{\textit{coupled reverse}}}{P_{\textit{incident}}}$$



Ideally, a signal traveling in the coupler's reverse direction will not appear at all at the coupled port, since its energy is either absorbed in the coupler's internal load or the external termination at the end of the main arm. In reality, however, some energy does leak through the coupled arm, as a result of finite isolation.

To measure isolation, we turn the coupler around and send power in the reverse direction. Isolation is defined as the leakage power at the coupled port relative to the incident power:

**Slide #101**




One of the most important measured parameter for couplers is their directivity. Directivity is a measure of a coupler's ability to separate signals flowing in opposite directions within the coupler. It can be thought of as the dynamic range available for reflection measurements. Directivity can be defined as:

Directivity (dB) = Isolation (dB) - Forward Coupling Factor (dB) - Loss (through-arm) (dB)

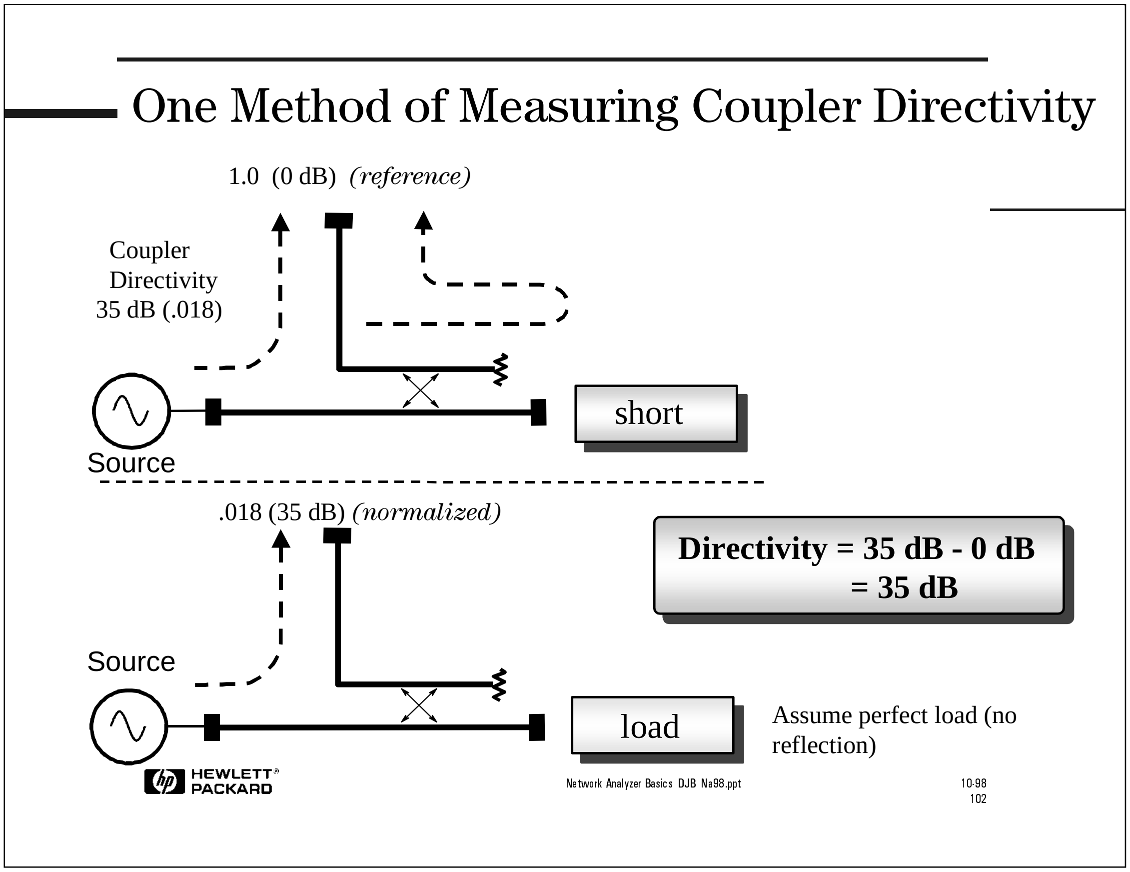
In the upper example in the above slide, our coupler exhibits a directivity of 30 dB. This means that during a reflection measurement, the directivity error signal is at best 30 dB below the desired signal (when measuring a device with full reflection or $\rho = 1$). The better the match of the device under test, the more measurement error the directivity error term will cause.

The slide also shows the effect of adding attenuators to the various ports of the coupler. The middle example shows that adding attenuation to the test port of a network analyzer reduces the raw (uncorrected) directivity by twice the value of the attenuator. While vector-error correction can correct for this, the stability of the calibration will be greatly reduced due to the degraded raw performance.

The lower example shows that adding an attenuator to the source side of the coupler has no effect on directivity. This makes sense since directivity is not a function of input-power level.

Adding an attenuator to the coupled port (not shown) affects both the isolation and forward-coupling factor by the same amount, so directivity is unaffected.

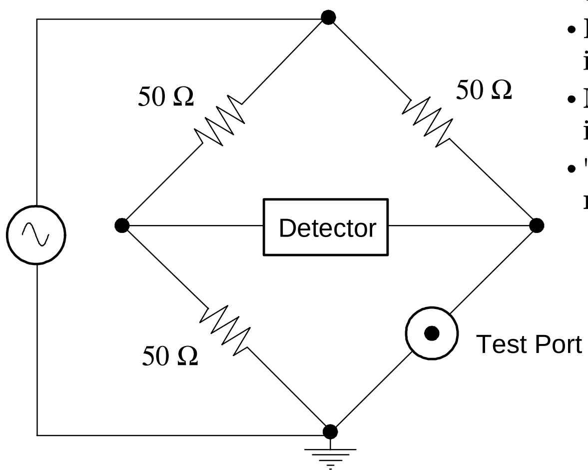
**Slide #102**




This is one method of measuring directivity in couplers (or in a network analyzer) that doesn't require forward and reverse measurements. First we place a short at the output port of the main arm (the coupler is in the reverse direction). We normalize our power measurement to this value, giving a 0 dB reference. This step accounts for the coupling factor and loss. Next, we place a (perfect) termination at the coupler's main port. Now, the only signal we measure at the coupled port is due to leakage. Since we have already normalized the measurement, the measured value is the coupler's directivity.

**Slide #103**

# Directional Bridge




- 50 ohm load at test port balances the bridge - detector reads zero
- Extent of bridge imbalance indicates impedance
- Measuring magnitude and phase of imbalance gives complex impedance
- "Directivity" is difference between maximum and minimum balance



Another device used for measuring reflected signals is the directional bridge. Its operation is similar to the simple Wheatstone bridge. If all four arms are equal in resistance (50 Ω connected to the test port) a voltage null is measured (the bridge is balanced). If the test-port load is not 50 Ω, then the voltage across the bridge is proportional to the mismatch presented by the DUT's input. The bridge is imbalanced in this case. If we measure both magnitude and phase across the bridge, we can measure the complex impedance at the test port.

A bridge's equivalent directivity is the ratio (or difference in dB) between maximum balance (measuring a perfect $Z_0$ load) and minimum balance (measuring a short or open). The effect of bridge directivity on measurement uncertainty is exactly the same as we discussed for couplers.

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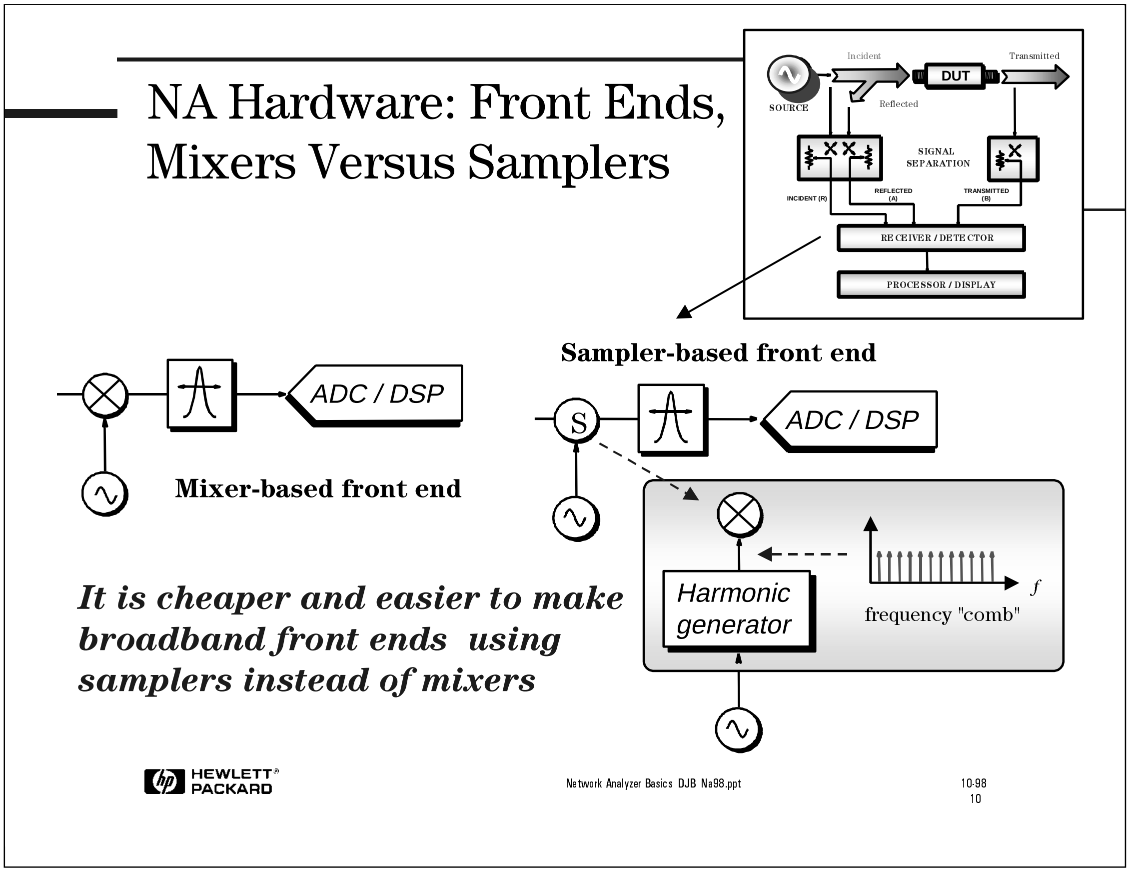
**Slide #104**




In this section, additional information is provided about the receiver section of the network analyzer.

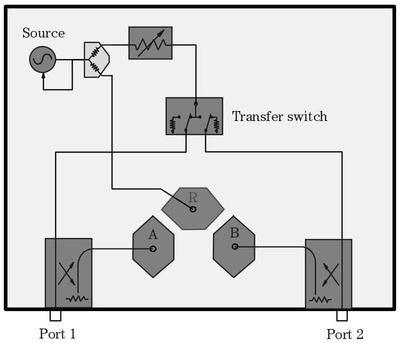
Tuned receivers can be implemented with mixer- or sampler-based front ends. It is often cheaper and easier to make wideband front ends using samplers instead of mixers, especially for microwave frequency coverage. Samplers are used with many of HP's network analyzers, such as the HP 8753D RF family and the HP 8720D microwave family of analyzers.

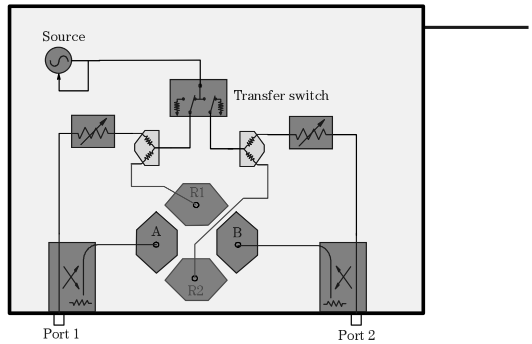
The sampler uses diodes to sample very short time slices of the incoming RF signal. Conceptually, the sampler can be thought of as a mixer with an internal pulse generator. The pulse generator creates a broadband frequency spectrum (often referred to as a "comb") composed of harmonics of the LO. The RF signal mixes with one of the spectral lines (or "comb tooth") to produce the desired IF. Compared to a mixer-based network analyzer, the LO in a sampler-based front end covers a much smaller frequency range, and a broadband mixer is no longer needed. The tradeoff is that the phase-lock algorithms for locking to the various comb teeth is more complex and time consuming.

Sampler-based front ends also have somewhat less dynamic range than those based on mixers and fundamental LOs. This is due to the fact that additional noise is converted into the IF from all of the comb teeth. Network analyzers with narrowband detection based on samplers still have far greater dynamic range than analyzers that use diode detection.

**Slide #105**

## Three Versus Four-Receiver Analyzers




***3 receivers***
- more economical
- TRL*, LRM* cal only
- includes:
  - HP 8753D
  - HP 8720D (std.)

***4 receivers***
- more expensive
- true TRL, LRM cal
- includes:
  - HP 8720D (opt. 400)
  - HP 8510C

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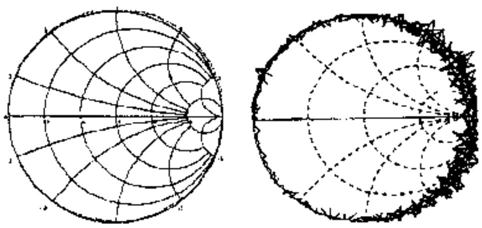
As already discussed there are two main types of test sets, transmission/reflection and s-parameter test sets. The s-parameter test set has two basic test set architectures: one employing three receivers (samplers or mixers) and one employing four receivers. The three-receiver architecture is simpler and less expensive, but the calibration choices are not as good. This type of network analyzer can do TRL* and LRM* calibrations (more on this later), but not true TRL or LRM.

Four-receiver analyzers are more expensive but provide better accuracy for noncoaxial measurements. We will cover this in more detail in the discussion about TRL in the next section.

**Slide #106**

## Why Are Four Receivers Better Than Three?

TRL



TRL*

***HP 8720D Opt. 400 adds fourth sampler, allowing full TRL calibration***

- **TRL***
  - assumes the source and load match of a test port are equal (port symmetry between forward and reverse measurements)
  - this is only a fair assumption for three-receiver (sampler) network analyzers
  - TRL* requires ten measurements to quantify eight unknowns
- **TRL**
  - Four receivers are necessary for all the measurements required for a full TRL cal (fourteen measurements to quantify ten unknowns)
  - TRL and TRL* use identical calibration standards
- **In noncoaxial applications:**
  - TRL achieves better source match and load match correction than TRL*
- **What about coaxial applications?**
  - SOLT is usually the preferred calibration method
  - Coaxial TRL can be more accurate than SOLT but not commonly used

10-98
10

Just what is the difference between TRL and TRL*?  TRL* assumes the source and load match of a test port are equal (i.e., there is port-impedance symmetry between forward and reverse measurements). This is only a fair assumption for a three-receiver network analyzer. TRL* requires ten measurements to quantify eight unknowns. True TRL calibration requires four receivers (two reference receivers plus one each for reflection and transmission) and fourteen measurements to quantify ten unknowns. TRL and TRL* use identical calibration standards.  The isolation portion of a TRL calibration is the same as for SOLT.

In noncoaxial applications, TRL achieves better source match and load match correction than TRL*, resulting in less measurement error. For coaxial applications, SOLT calibration is almost always the preferred method.  HP can provide coaxial calibration kits all the way up to 110 GHz, with a variety of connector types. While not commonly done, coaxial TRL calibration can be more accurate than SOLT calibration, but only if very-high quality coaxial transmission lines (such as beadless airlines ) are used.

Option 400 for the HP 8720D family adds a fourth sampler, allowing these analyzers to do a full TRL calibration.

## Slide #107

### Challenge Quiz

**1. Can filters cause distortion in communications systems?**
- A. Yes, due to impairment of phase and magnitude response
- B. Yes, due to nonlinear components such as ferrite inductors
- C. No, only active devices can cause distortion
- D. No, filters only cause linear phase shifts
- E. Both A and B above

**2. Which statement about transmission lines is false?**
- A. Useful for efficient transmission of RF power
- B. Requires termination in characteristic impedance for low VSWR
- C. Voltage is independent of position along line
- D. Used when wavelength of signal is small compared to length of line
- E. Can be realized in a variety of forms such as coaxial, waveguide, microstrip

**3. Which statement about narrowband detection is false?**
- A. Is only available in vector network analyzers
- B. Provides much greater dynamic range than diode detection
- C. Uses variable-bandwidth IF filters to set analyzer noise floor
- D. Provides rejection of harmonic and spurious signals
- E. Uses mixers or samplers as downconverters

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## Slide #108

### Challenge Quiz (continued)

**4. Maximum dynamic range with narrowband detection is defined as:**
- A. Maximum receiver input power minus the stopband of the device under test
- B. Maximum receiver input power minus the receiver's noise floor
- C. Detector 1-dB-compression point minus the harmonic level of the source
- D. Receiver damage level plus the maximum source output power
- E. Maximum source output power minus the receiver's noise floor

**5. With a T/R analyzer, the following error terms can be corrected:**
- A. Source match, load match, transmission tracking
- B. Load match, reflection tracking, transmission tracking
- C. Source match, reflection tracking, transmission tracking
- D. Directivity, source match, load match
- E. Directivity, reflection tracking, load match

**6. Calibration(s) can remove which of the following types of measurement error?**
- A. Systematic and drift
- B. Systematic and random
- C. Random and drift
- D. Repeatability and systematic
- E. Repeatability and drift

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## Slide #109

### Challenge Quiz (continued)

**7. Which statement about TRL calibration is false?**
- A. Is a type of two-port error correction
- B. Uses easily fabricated and characterized standards
- C. Most commonly used in noncoaxial environments
- D. Is not available on the HP 8720D family of microwave network analyzers
- E. Has a special version for three-sampler network analyzers

**8. For which component is it hardest to get accurate transmission and reflection measurements when using an HP 8711C scalar network analyzer?**
- A. Amplifiers because output power causes receiver compression
- B. Cables because load match cannot be corrected
- C. Filter stopbands because of lack of dynamic range
- D. Mixers because of lack of broadband detectors
- E. Attenuators because source match cannot be corrected

**9. Power sweeps are good for which measurements?**
- A. Gain compression
- B. AM to PM conversion
- C. Saturated output power
- D. Power linearity
- E. All of the above

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The correct answers to the challenge quiz are:

1. E
2. C
3. A
4. B
5. C
6. A
7. D
8. B
9. E

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