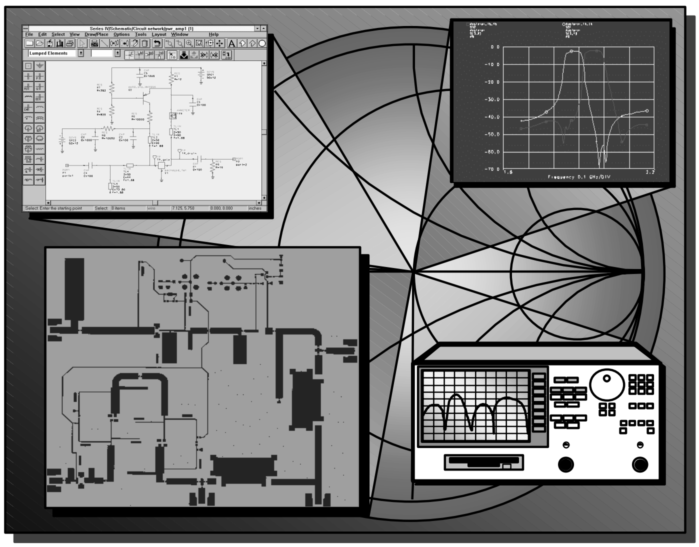
Slide #173

Appendix



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The appendix will briefly cover transmission lines, the Smith chart, measurement terminology for reflection and transmission measurements, S-parameters, and nonlinear measurements. For additional application notes visit our web site, http://www.hp.com or contact your local HP sales office.

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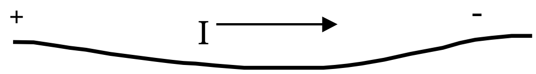
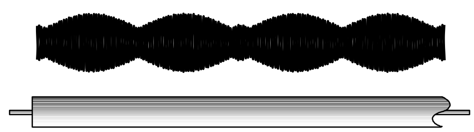
Slide #174

Power Transfer 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 (Z_0) is very important for low reflection and maximum power transfer
- measured envelope voltage dependent on position along line

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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. When the transmission line is terminated in its characteristic impedance Zo (which is generally a pure resistance such as 50 or 75 ohms), maximum power is transferred to the load. When the termination is not Zo, the portion of the signal which is not absorbed by the load is reflected back toward the source. This creates a condition where the voltage along the transmission line varies with position. We will examine the incident and reflected waves on transmission lines with different load conditions in following slides.

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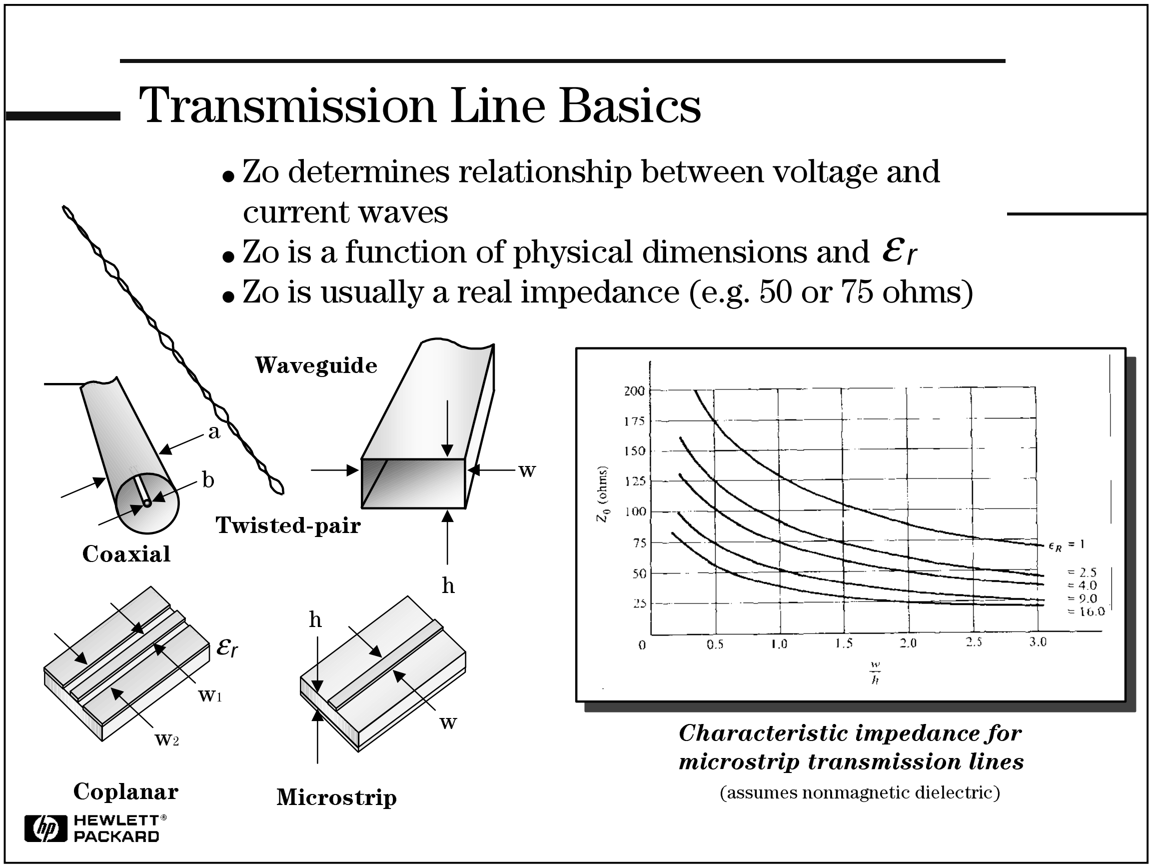

RF transmission lines can be made in a variety of transmission media. Common examples are coaxial, waveguide, twisted pair, coplanar, stripline and microstrip. RF circuit design on printed-circuit boards (PCB) often use coplanar or microstrip transmission lines. The fundamental parameter of a transmission line is its characteristic impedance Zo. Zo describes the relationship between the voltage and current traveling waves, and is a function of the various dimensions of the transmission line and the dielectric constant (er) of the non-conducting material in the transmission line. For most RF systems, Zo is either 50 or 75 ohms.

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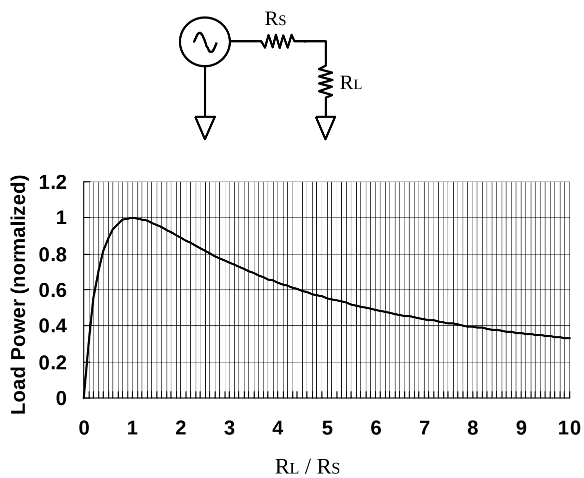
Slide #176

Power Transfer Efficiency



Maximum power is transferred when $R_L = R_S$

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Let's explore the criteria for efficient RF power transfer in more detail, starting with the condition for maximum power transfer into a load, given a source impedance of Rs. The graph above shows that the matched condition (RL = Rs) results in the maximum power dissipated in the load resistor. This condition is true regardless if the stimulus is a DC voltage source or an RF sinusoid.

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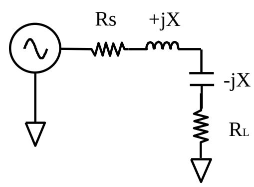
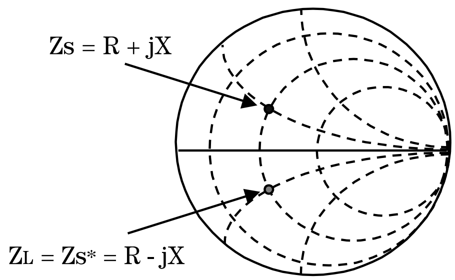
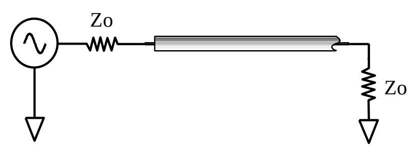
Power Transfer Efficiency

For complex impedances, maximum power transfer occurs when $Z_L = Z_S^*$ (conjugate match)





At high frequencies, maximum power transfer occurs when $R_S = R_L = Z_o$



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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 $R_s^* = 0.6 - j0.3$. This can be accomplished by mirroring the impedance about the horizontal axis of the Smith chart.

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 defined 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.

Getting efficient transfer of RF power into a transmission line requires that Rs = Zo. For efficient absorption of power at the other end, RL must equal Zo.

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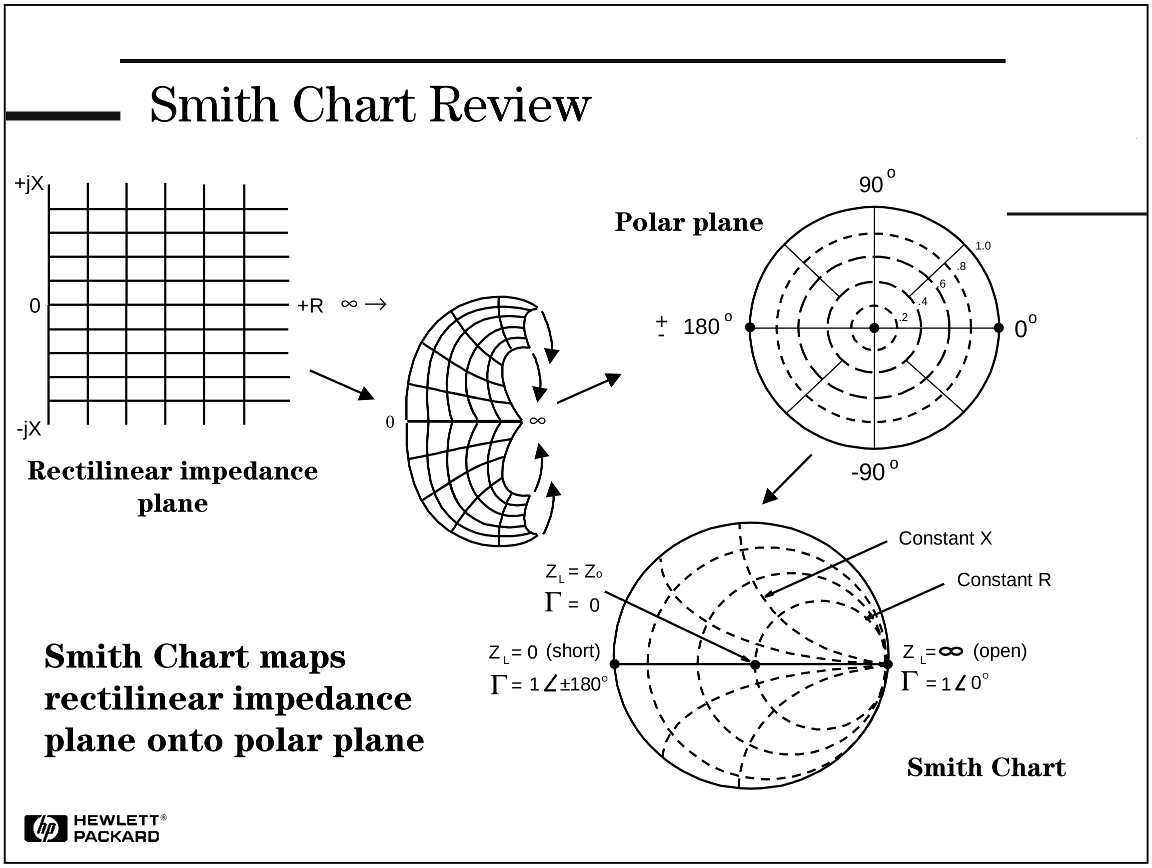

The Smith chart is the most common way to display complex impedance. Let's review how the Smith chart works. The amount of reflection that occurs when characterizing a device depends on the impedance the incident signal sees. Since any impedance can be represented as a real and imaginary part (R+jX or G+jB), we can easily see how these quantities can be plotted on a rectilinear grid (known as the complex impedance plane). Unfortunately, the open circuit (quite a common impedance value for RF measurements) appears at infinity on the x-axis.

The polar plot is very useful since the entire impedance plane is covered. But instead of actually plotting impedance, we display the reflection coefficient in vector form. The magnitude of the vector is the distance from the center of the display, and phase is displayed as the angle of vector referenced to a flat line from the center to the rightmost edge. The drawback of polar plots is that impedance values cannot be read directly from the display.

Since there is a one-to-one correspondence between complex impedance and reflection coefficient, we can map the positive real half of the complex impedance plane onto the polar display. The result is the Smith chart. All values of reactance and all positive values of resistance from 0 to ¥ fall within the outer circle of the Smith chart. Loci of constant resistance now appear as circles, and loci of constant reactance appear as arcs. Impedances on the Smith chart are always normalized to the characteristic impedance of the test system. A perfect termination (Zo) appears in the center of the chart.

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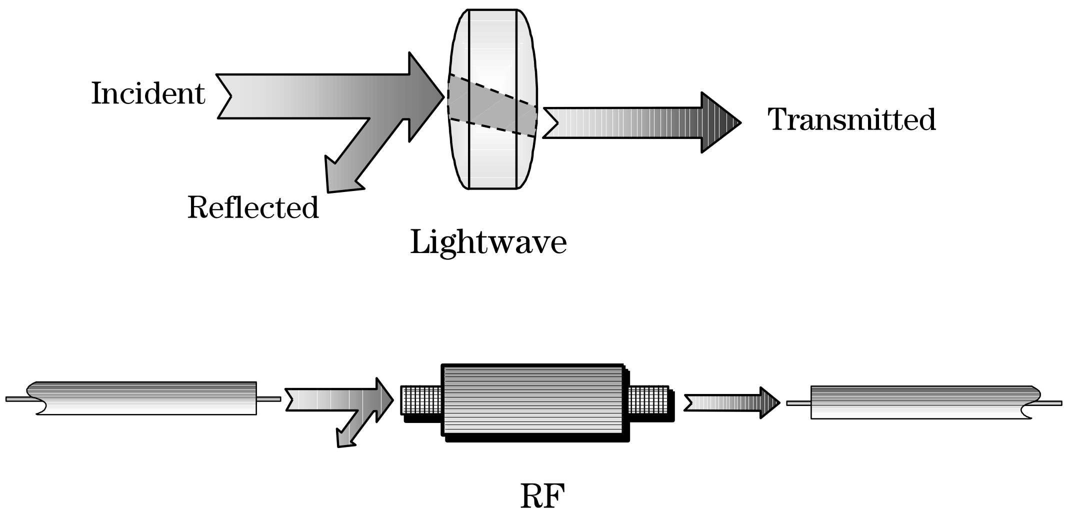
Slide #179

Lightwave Analogy to RF Energy



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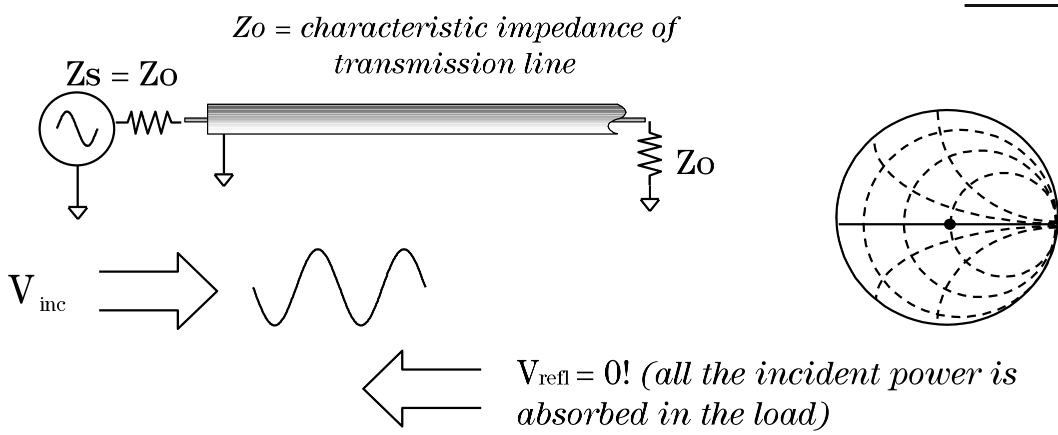
One of the fundamental concepts of RF power transmission involves incident, reflected and transmitted waves traveling along transmission lines. It is helpful to think of traveling waves along a transmission line in terms of a lightwave analogy. We can imagine incident light striking some optical component 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 were made of some lossy material, then some of the light could be absorbed within 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, except the electromagnetic energy is in the RF range instead of the optical range, and our components and circuits are electrical devices and networks instead of lenses and mirrors.

Slide #180

Transmission Line Terminated with Zo



For reflection, a transmission line terminated in Zo behaves like an infinitely long transmission line

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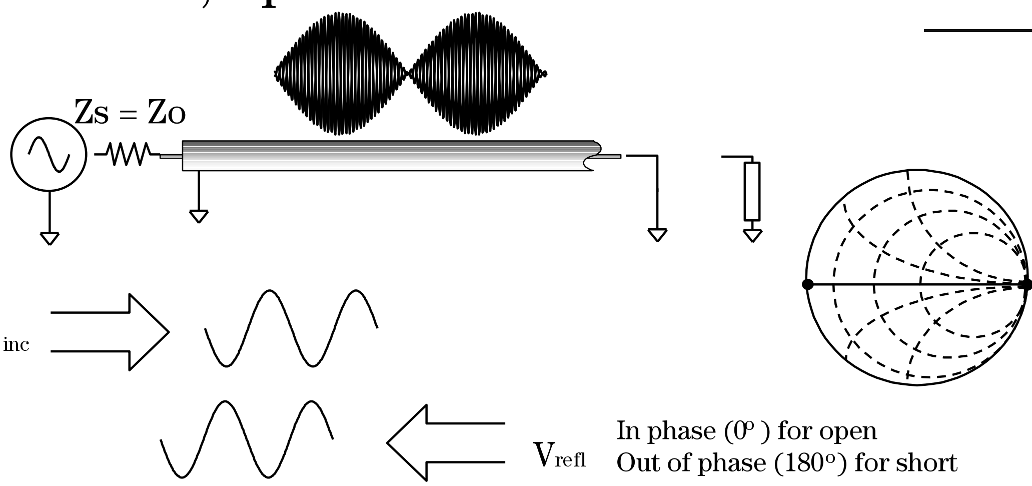
Let's review what happens when transmission lines are terminated in different impedances, starting with a Zo load. Since a transmission line terminated in its characteristic impedance results in maximum transfer of power to the load, there is no reflected signal. This result is the same as if the transmission 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 (forward).

Slide #181

Transmission Line Terminated with Short, Open



For reflection, a transmission line terminated in a short or open reflects all power back to source

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 180o out of phase with it (at the plane of the short). 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's 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 180o 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 conditions will be shifted in position along the transmission line with respect to each other, in order to satisfy Ohm's law as described above.

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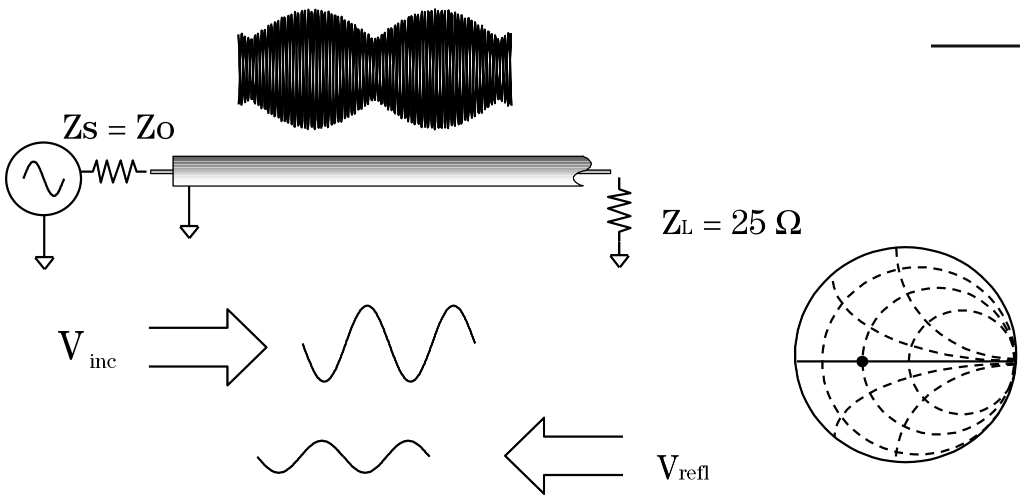
Slide #182

Transmission Line Terminated with 25 Ω



Standing wave pattern does not go to zero as with short or open

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Finally, let's terminate our line with a 25-ohm resistor (an impedance between the full reflection of an open or short circuit and the perfect termination of a 50-ohm 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 180o 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 to zero, and the peak will be less than that of the short/open case.

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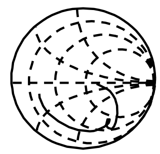
Slide #183

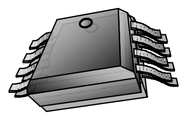
Device Characteristics

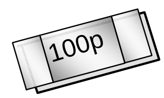
Devices have many distinctive characteristics such as:

- ***electrical behavior***
  - DC power consumption
  - linear (e.g. S-parameters, noise figure)
  - nonlinear (e.g. distortion, compression)

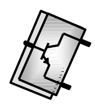


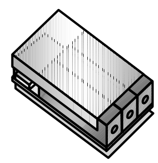
- ***physical specifications***
  - package type
  - package size
  - thermal resistance





- ***other things...***
  - cost
  - availability





When selecting parts for design, characteristics are traded-off

Let's look at important electrical characteristics for RF design ...

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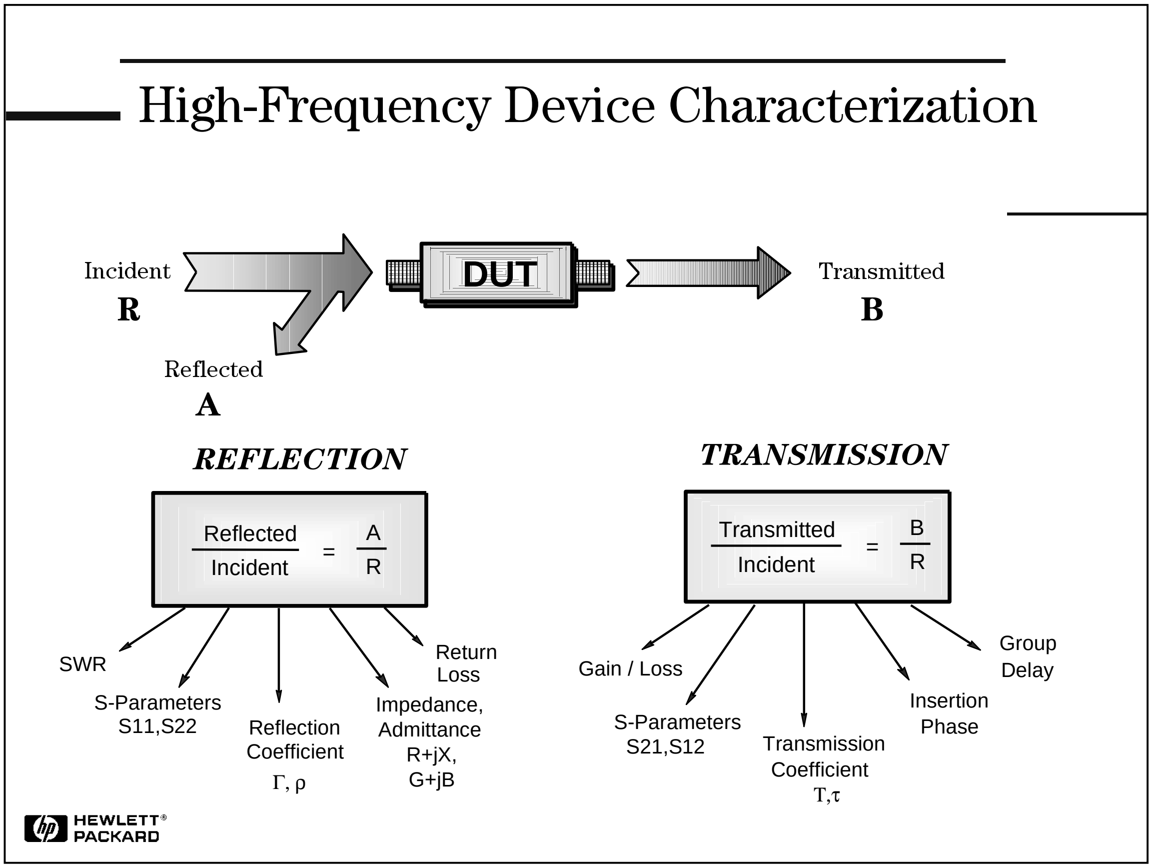
Let's shift our discussion now from transmission lines to characterizing devices. A device or component has many different characteristics ranging from physical and electrical parameters to less tangible (but nevertheless important) considerations like cost and availability. During the course of selecting parts to be used in our design, we have to trade-off these various parameters to arrive at an optimum balance. For example, cost and performance are often trade-offs, or package size and maximum-power dissipation.

In the next few slides, we will review the terms for electrical characteristics of devices that are important for RF design.

Slide #184



Now that we fully understand the nature of electromagnetic waves, we must also recognize the terms used to describe them. Common network analyzer terminology has the incident wave measured with the R (for reference) channel, while 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 (complex) 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.

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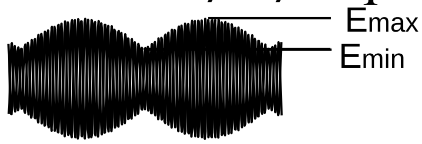
Slide #185

Reflection Parameters

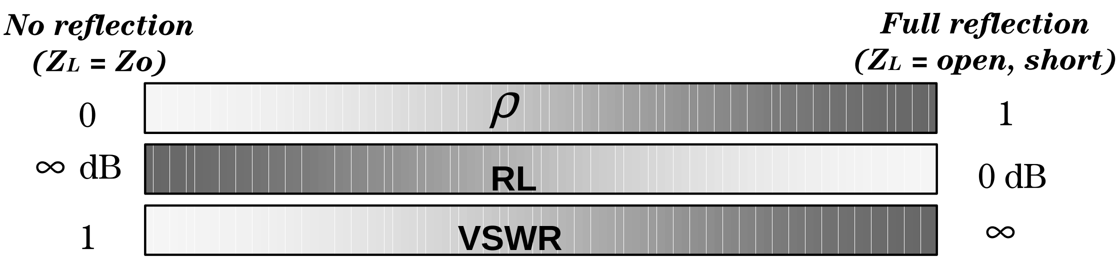
Reflection Coefficient $\Gamma = \dfrac{V_{reflected}}{V_{incident}} = \rho \angle \Phi = \dfrac{Z_L - Z_O}{Z_L + Z_O}$

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



Voltage Standing Wave Ratio

$$\textbf{VSWR} = \frac{E_{max}}{E_{min}} = \frac{1+\rho}{1-\rho}$$



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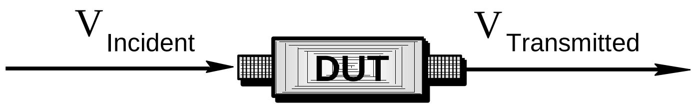
The principle term for reflected waves is reflection coefficient gamma (G). The magnitude portion of gamma is called rho (r). Reflection coefficient is the ratio of the reflected signal voltage to the incident signal voltage. For example, a transmission line terminated in Zo will have all energy transferred to the load; hence Vrefl = 0 and r = 0. When ZL is not equal to Zo , some energy is reflected and r is greater than zero. When ZL = a short or open circuit, all energy is reflected and r = 1. The range of possible values for r 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 Zo 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+r)/(1-r). VSWR can take on values between one and infinity.

Slide #186

Transmission Parameters



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

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

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

Insertion Phase (deg) = $\angle \dfrac{V_{Trans}}{V_{Inc}} = \phi$

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Transmission coefficient T is defined as the transmitted voltage divided by the incident voltage. If |Vtrans| > |Vinc|, we have gain, and if |Vtrans| < |Vinc|, 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.

Looking at insertion phase directly is usually not very useful. This is because the phase has a large negative slope with respect to frequency due to the electrical length of the device (the longer the device, the greater the slope). Often, the electrical delay feature of the network analyzer is used to remove the linear portion of the phase response. This has the effect of removing the electrical length of the DUT, resulting in a high-resolution display showing any deviations from linear phase.

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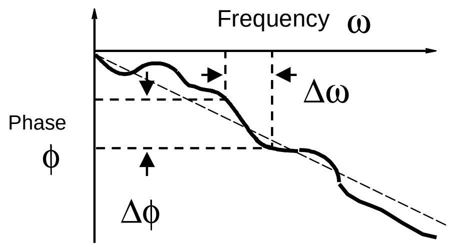
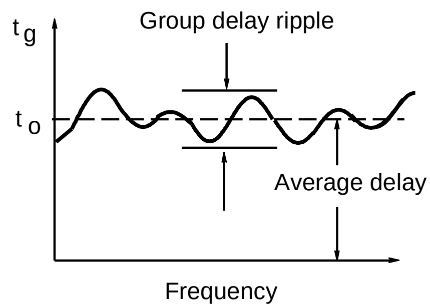
Slide #187

Group Delay (GD)





Group Delay (t_g) =

$$\frac{-d\,\phi}{d\,\omega} = \frac{-1}{360^{o}} * \frac{d\,\phi}{d\,f}$$

ϕ in radians

ω in radians/sec

ϕ in degrees

f in Hertz ($\omega = 2\,\pi\,f$)

- average delay indicates electrical length
- GD ripple indicates distortion
- aperture of measurement is very important
  - aperture is frequency-delta used to calculate GD
  - wider aperture: lower noise / less resolution
  - narrower aperture: more resolution / higher noise

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Looking at insertion phase directly is usually not very useful. This is because the phase has a large 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 that we are generally concerned with, it is desirable to remove the linear portion of the phase response. This can be accomplished by displaying the phase response of the DUT in terms of group delay. Group delay is a measure of the transit time of a signal through a DUT versus frequency. Group delay is calculated by differentiating the insertion-phase response of the DUT versus frequency. 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. 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.

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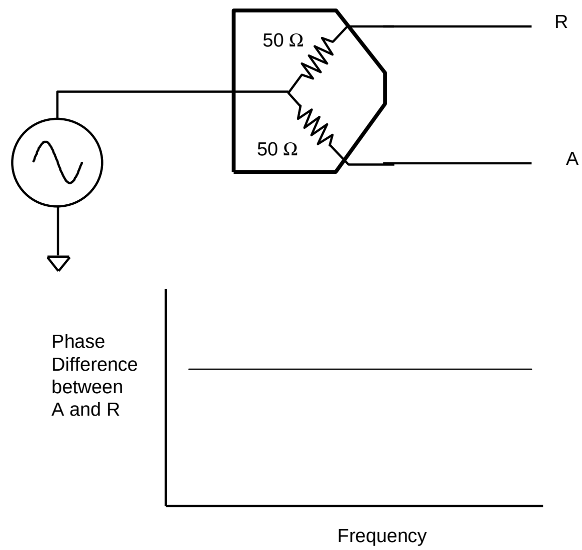
Slide #188

Phase versus Frequency



Here are additional slides to clarify deviation from linear phase. First consider phase difference between A and R channel. When distance is equal, phase difference is zero. Phase change verses frequency is flat, and scale per division can be reduced to view any ripple or noise

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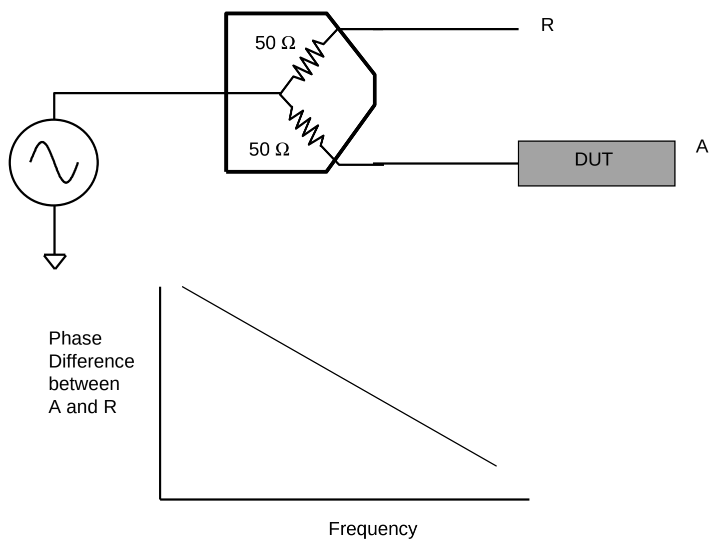
Slide #189

Phase versus Frequency



Now add device under test (DUT). Phase at the output will lag with respect to the R channel (also with respect to DUT input. Phase difference will increase (negative) with frequency as wavelengths become shorter.

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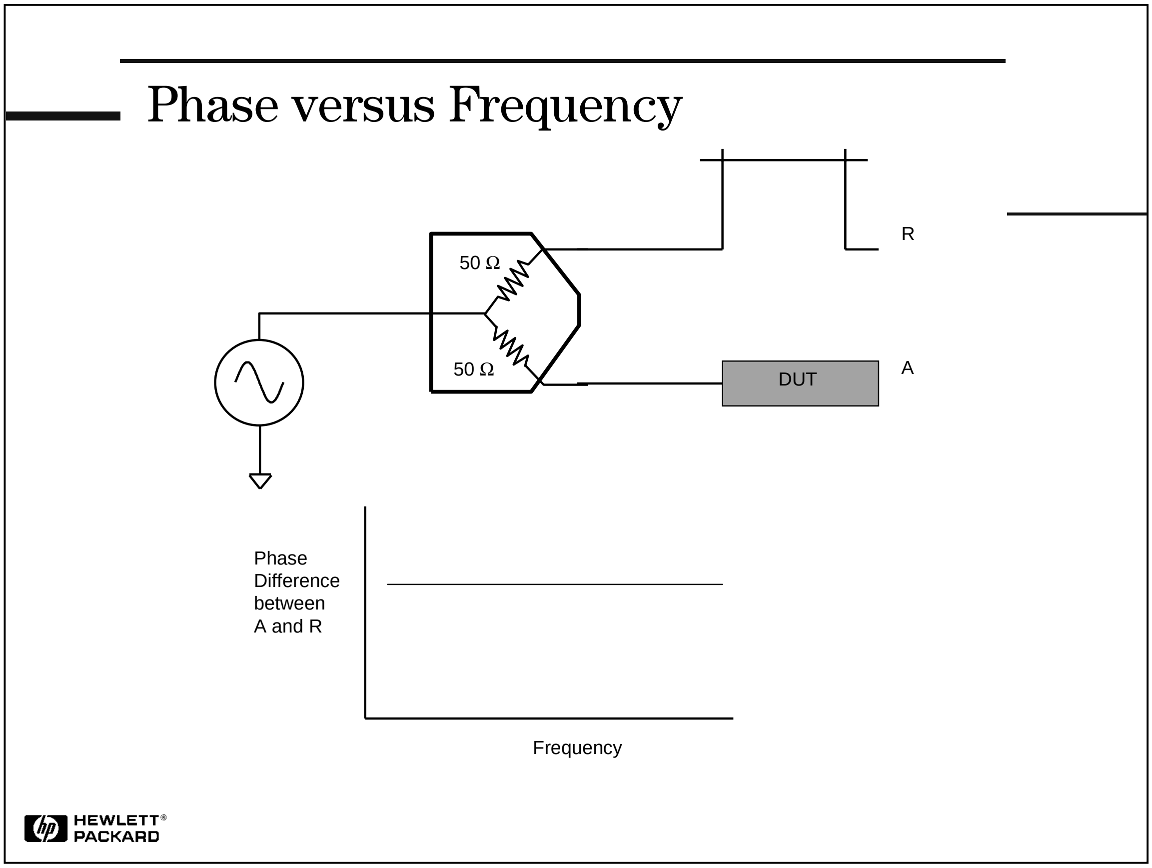
Slide #190

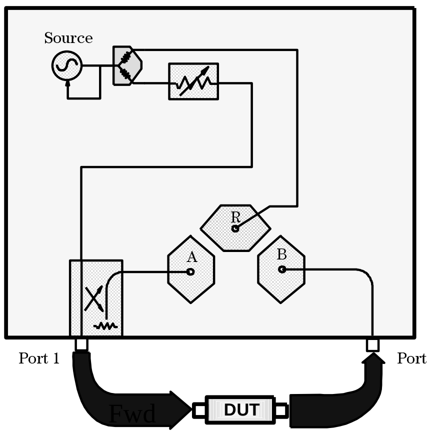


Electrical delay is added to the R channel to correct for non equal lengths. Adding delay “flattens” the phase versus frequency curve. Now the scale per division can be reduced again to see any ripple or noise.

Slide #191

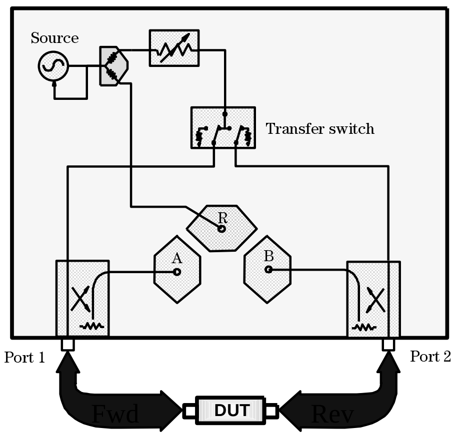
T/R Versus S-Parameter Test Sets

Transmission/Reflection Test Set



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

S-Parameter Test Set



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

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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.

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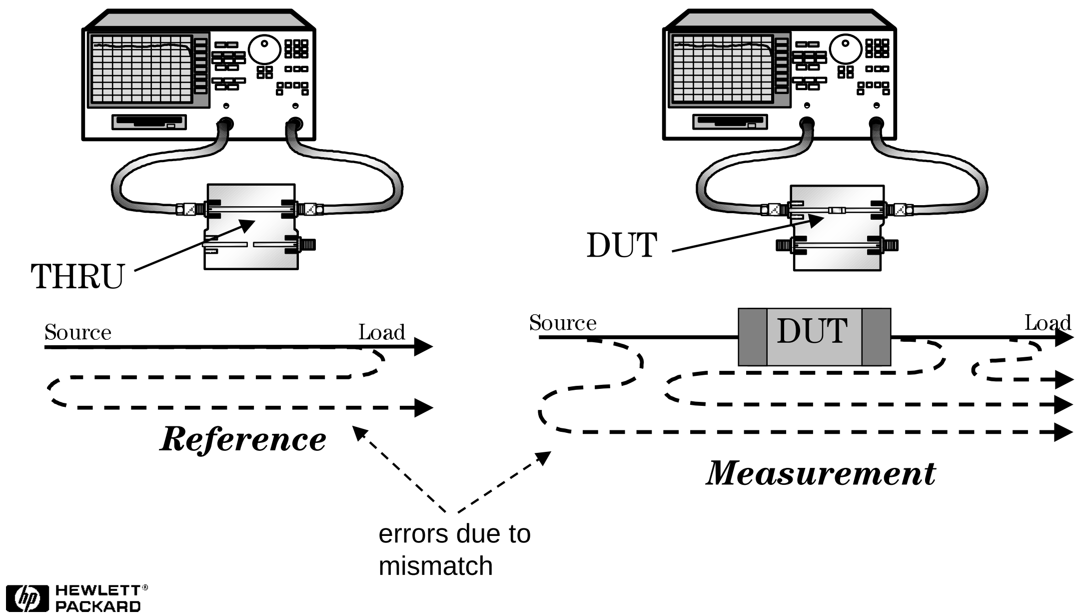
Slide #192

Response Calibration



Response calibration has a serious inherent weakness because no correction can be done for errors due to source and load match. This is especially a problem for low-loss transmission measurements (such as measuring a filter passband or a cable) and for reflection measurements. Using response calibration for transmission measurements on low-loss devices can result in considerable measurement uncertainty in the form of ripple. Measurement accuracy will depend on the relative mismatch of the test fixture and network analyzer compared to the DUT.

When response calibration is used for transmission measurements with fixtures, considerable measurement improvement can be made by first performing a two-port correction at the ends of the test cables. This will improve the effective source and load match of the network analyzer, thus helping to reduce the measurement ripple due to reflections from the fixture and the analyzer's test ports.

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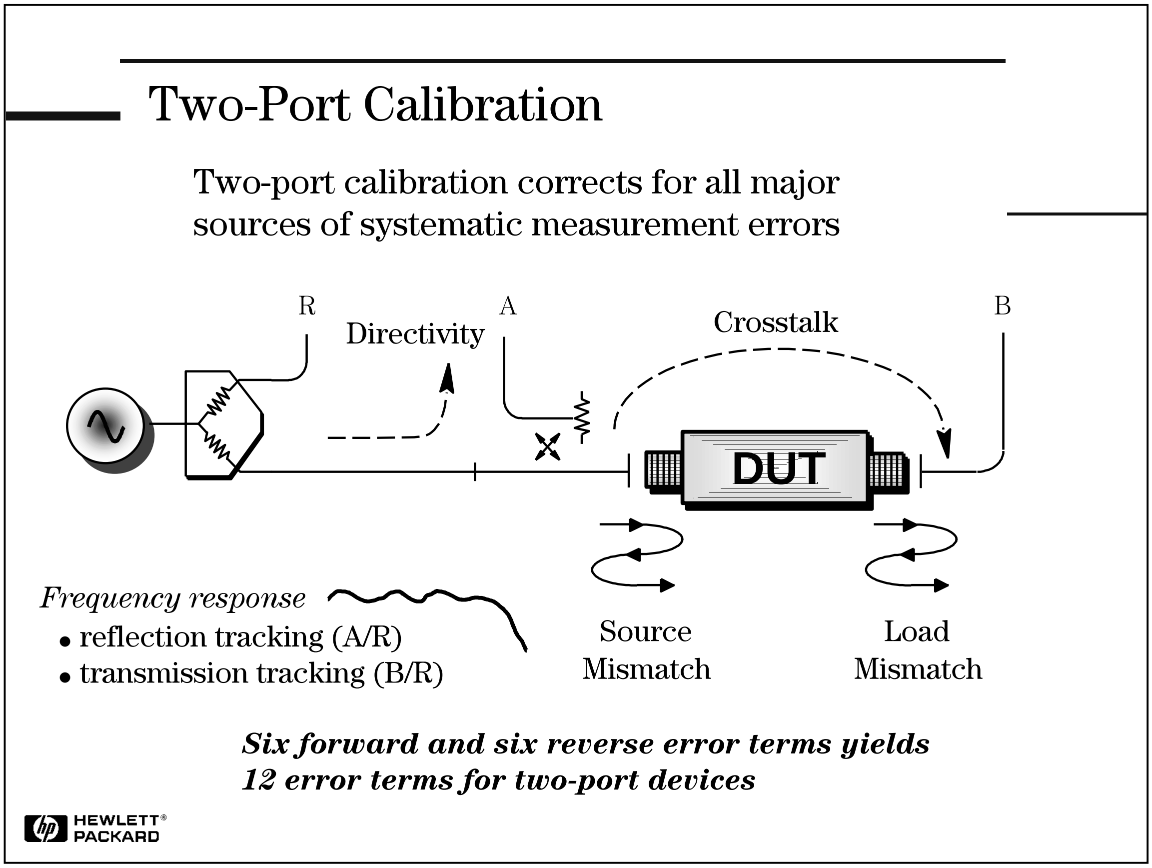
Slide #193



Two-port calibration is the most accurate form of error correction since it accounts for all of the major sources of systematic error associated with network measurements. These errors are shown above. The errors relating to signal leakage are directivity and crosstalk. The 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 #194

Thru-Reflect-Line (TRL) Calibration

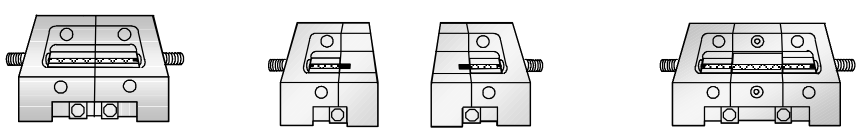
TRL calibration was developed for non-coaxial microwave measurements

Advantages

- microwave cal standards **easy** to make (no open or load)
- based on **transmission line** of known length and impedance
- do not need to know characteristics of **reflect** standard

Disadvantages

- impractical **length** of RF transmission lines
- fixtures usually more **complicated** (and expensive)
- 8:1 BW **limitation** per transmission line



As was mentioned previously, the two basic types of two-port calibrations are TRL and SOLT. TRL calibration was originally developed for non-coaxial microwave measurements (such as wafer-probing, fixtured, and waveguide applications). The main advantage of TRL is that the calibration standards are relatively easy to make and define at microwave frequencies. This is a big benefit for microwave applications, where it is difficult to build good open and load standards that are needed for an SOLT calibration. TRL uses a transmission line of known length and impedance as one standard. The only restriction is that the line needs to be significantly longer in electrical length than the thru line, which typically is of zero length. The general rule-of-thumb for the line standard is that it should be between 20 and 160 degree in length. TRL calibration also uses a high-reflection standard (usually a short or open) whose impedance does not have to be well characterized.

For RF applications, the lengths of the transmission lines needed to cover down to low frequencies become impractical (too long). It is also difficult to make good TRL standards on printed-circuit boards, due to dielectric and line-dimension variations. And, the typical TRL fixture tends to be more complicated and expensive.

For additional reference material on TRL calibration, contact your local field engineer and request the following Product notes:

Applying the HP 8510 TRL calibration for non-coaxial measurements Product Note 8510-8A (HP 5091-3645E)

In-fixture microstrip device measurements using TRL* calibration Product Note 8720-2 (HP 5091-1943E)

Slide #195

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

$V_1 = h_{11}I_1 + h_{12}V_2$

$I_2 = h_{21}I_1 + h_{22}V_2$

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

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

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.

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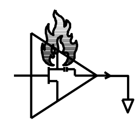
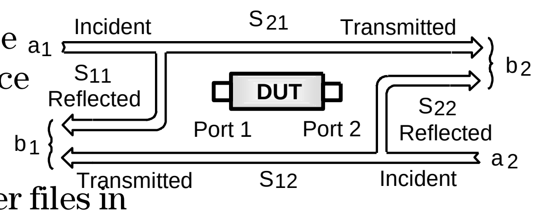
Slide #196

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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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.

For these reasons, 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 relatively easy to measure, and don't require the 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 if desired. 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 device has N2 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 energy emerges, and the second number is the port where energy enters. So, S21 is a measure of power coming out port two as a result of applying an RF stimulus to port one. When the numbers are the same (e.g. S11), it indicates a reflection measurement.

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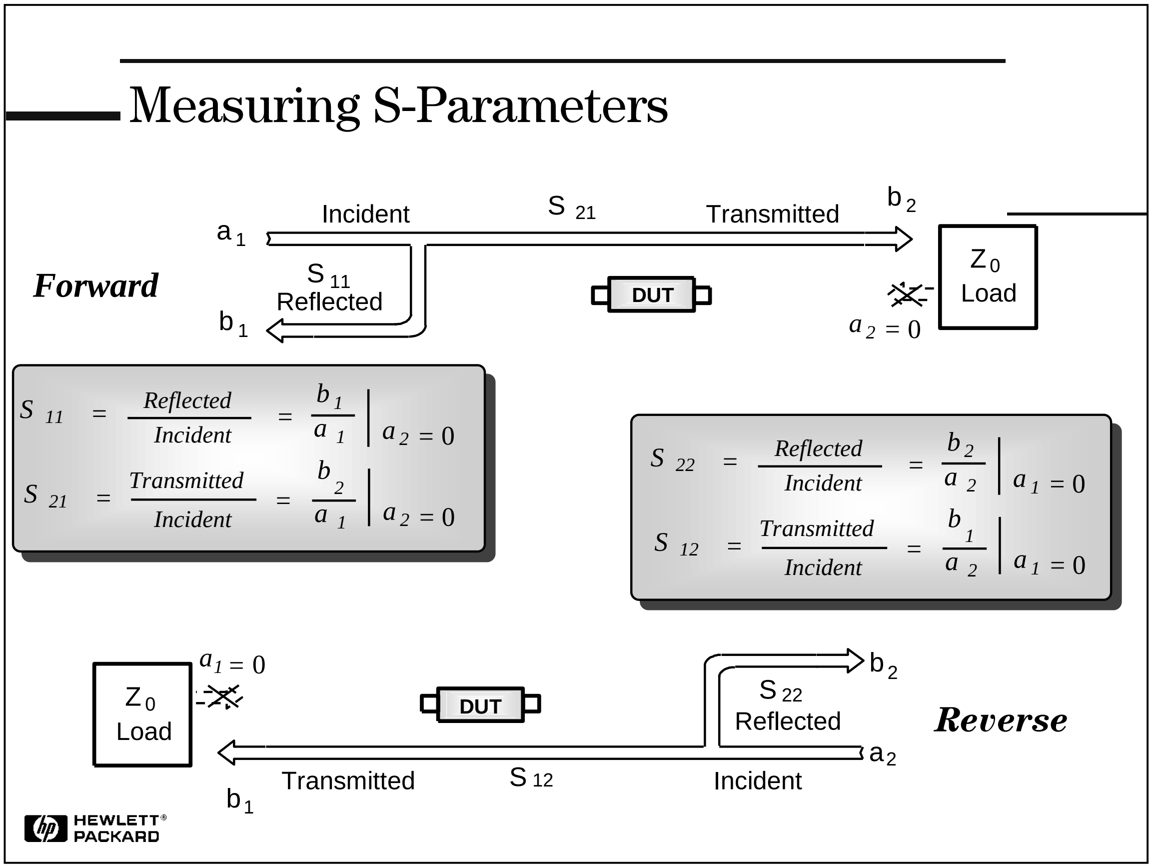
Slide #197



S11 and S21 are determined by measuring the magnitude and phase of the incident, reflected and transmitted signals when the output is terminated in a perfect Zo load. This condition guarantees that a2 is zero. 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 a1 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 a1 or a2 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.

Slide #198

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.

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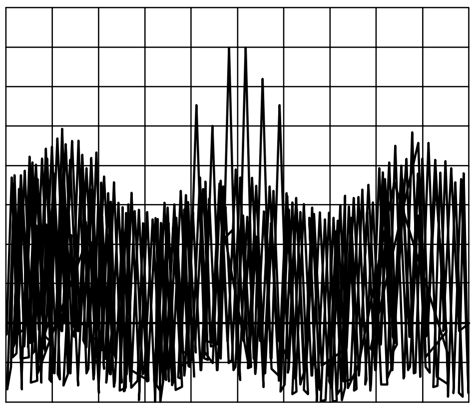
Slide #199

Going Beyond Linear Swept-Frequency Characterization

So far, we've only talked about linear swept-frequency characterization (used for passive and active devices).
Two other important characterizations for active devices are:

- nonlinear behavior
- noise figure



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So far, we've only talked about linear swept-frequency characterization (used for passive and active devices). Two other important characterizations for active devices are nonlinear behavior such as intermodulation distortion and gain compression, and noise figure.

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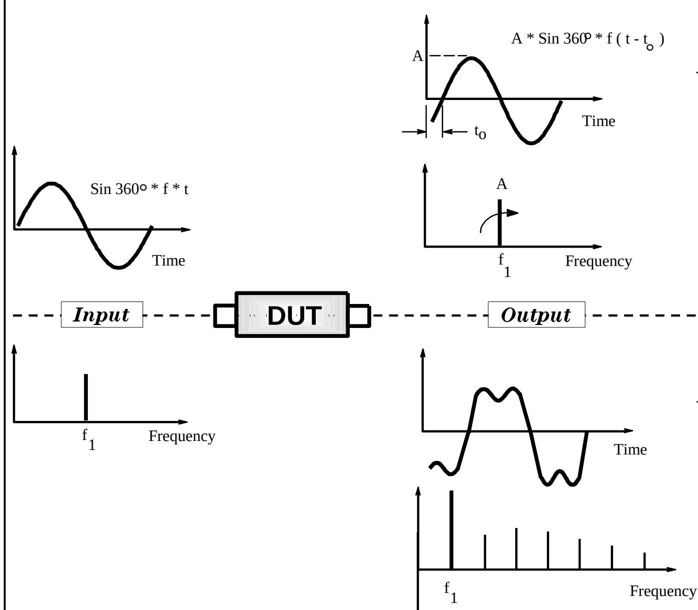
Slide #200

Linear Versus Nonlinear Behavior



Linear behavior:
- input and output frequencies are the same (no additional frequencies created)
- output frequency only undergoes magnitude and phase change

Nonlinear behavior:
- output frequency may undergo frequency shift (e.g. with mixers)
- additional frequencies created (harmonics, intermodulation)

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In order to understand nonlinear measurements, 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.

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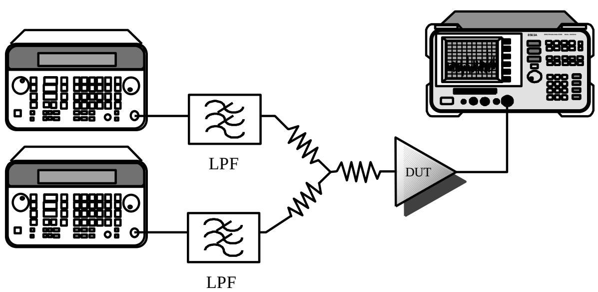
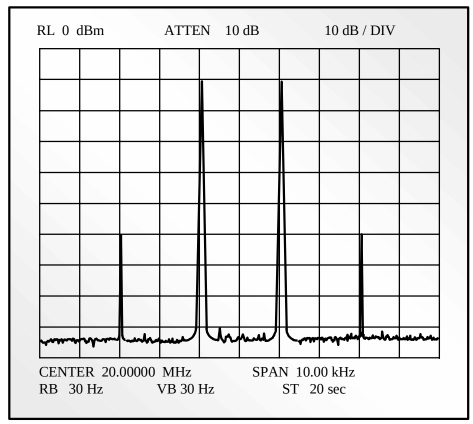
Slide #201

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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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 power-amplifier section of this presentation.

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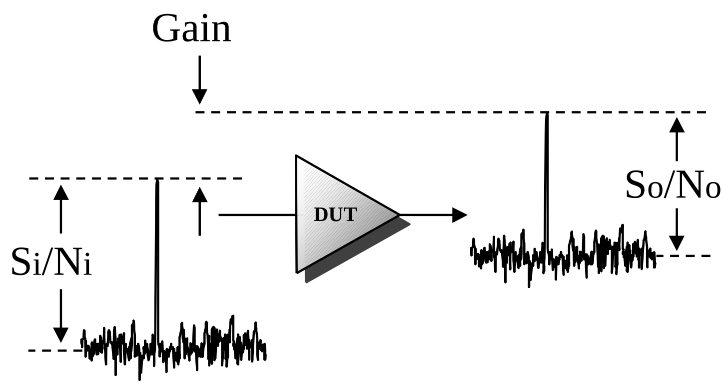
Slide #202

Noise Figure (NF)



- Measure of noise added by amplifier
- NF = 10 log [(S_i/N_i) / (S_o/N_o)]
- Perfect amp would have 0 dB NF

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Another important characterization of active devices or circuits is noise figure. Noise figure is a measure of the noise added by an amplifier (or a more complicated circuit such as a receiver). It is generally expressed as 10 log of the ratio of the input signal-to-noise ratio to the output signal-to-noise ratio. A perfect (non-realizable) amplifier would have a noise figure of 0 dB.

Noise figure is usually measured with a noise-figure meter or a spectrum analyzer with a special measurement personality. We will cover the most commonly used technique for measuring noise figure in more detail during the low-noise-amplifier section of this presentation.

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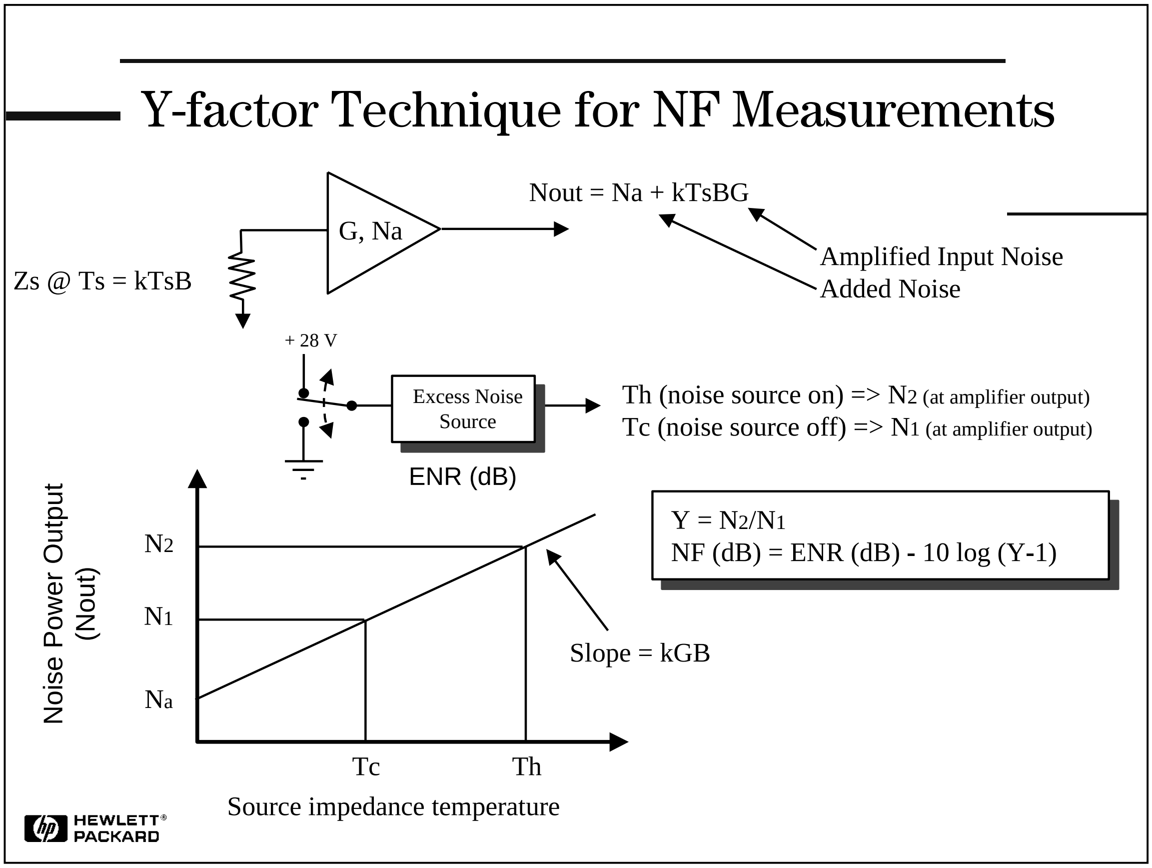
Slide #203



Perhaps the most common and easiest way to measure the noise figure of an amplifier is by using the Y-factor technique. The Y-factor technique makes use of two measurements at the output of an amplifier: one with a terminated input, and one with the excess-noise source applied to the input. The excess-noise source is usually an avalanche diode, biased to give a repeatable amount of noise above that of a simple resistor. The amount of excess noise is expressed as the excess-noise ratio (ENR). The noise at the output of the amplifier is a combination of the amplified input noise and the noise added by the amplifier itself.

Since noise can be represented as an equivalent noise temperature (the Ts in kTsB), we can plot output noise power versus input noise temperature. The terminated input is equivalent to the cold temperature Tc , and the excess-noise source is equivalent to the hot temperature Th. The graph above shows the two measured output powers N1 and N2 , and the two effective input temperatures. The y-intercept of the line drawn between the two measured points represents the noise added by the amplifier, which can be expressed in terms of noise figure. The slope of the line is proportional to the gain of the amplifier.

To actually compute noise figure, the Y-factor is defined as N2 divided by N1 (in linear terms). Then, noise figure is calculated as follows:

$$NF\ (dB) = ENR\ (dB) - 10 * \log (Y-1)$$

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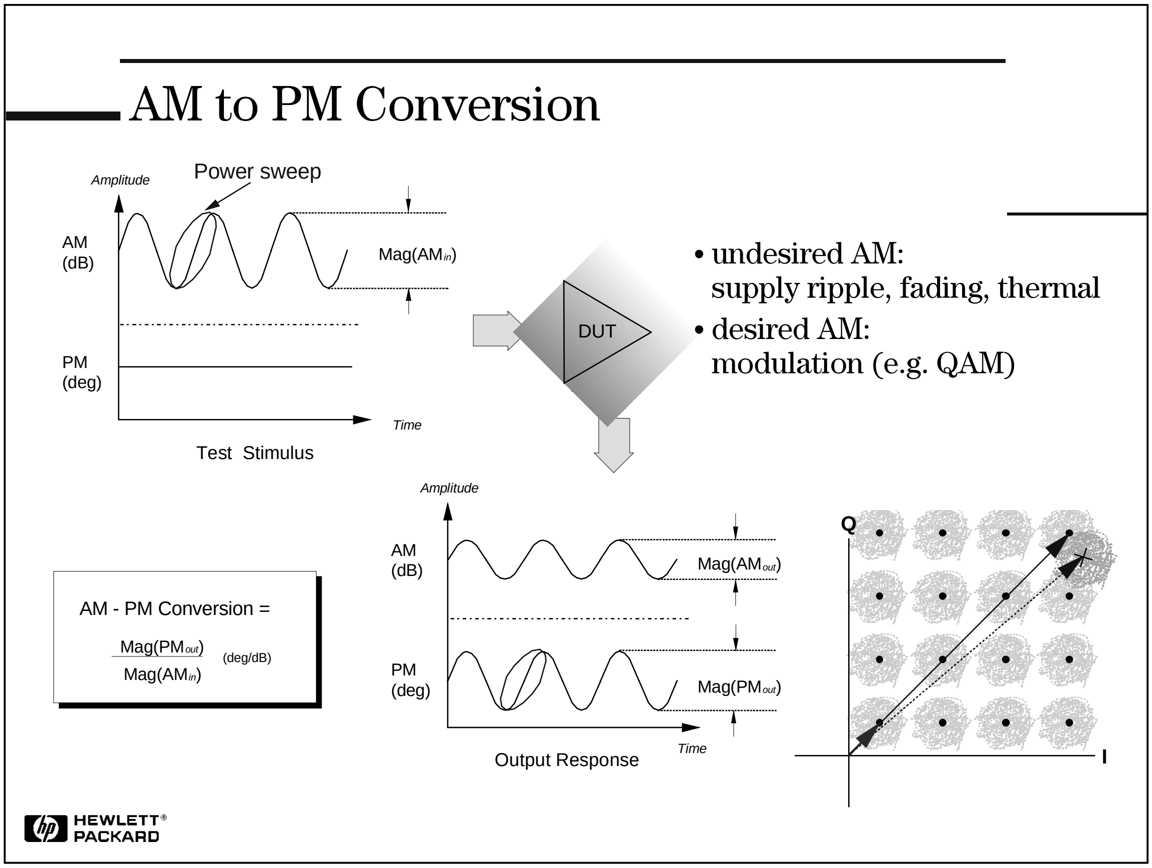
Slide #204



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.

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

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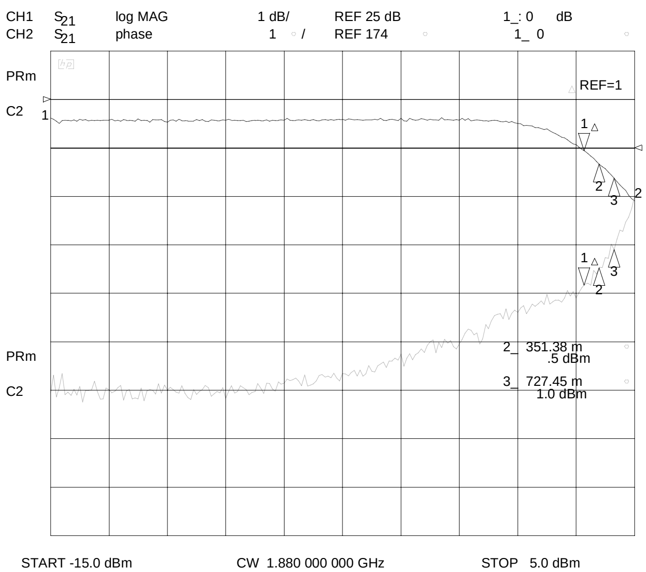
Slide #205

Measuring AM to PM Conversion



- use transmission setup with a power sweep
- display phase of S21
- AM to PM = 0.727deg/dB

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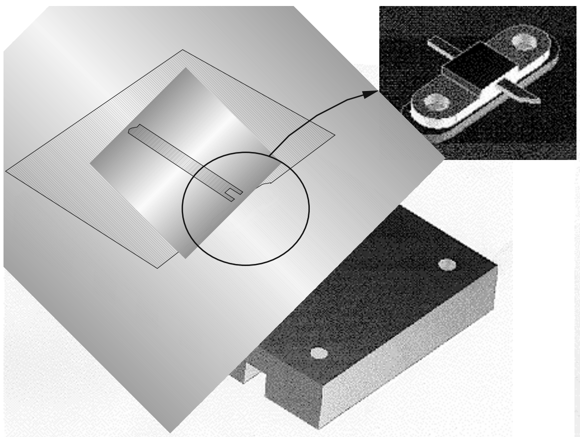
AM-to-PM conversion can be measured by performing a power sweep with a vector network analyzer, using the same transmission setup that we used for measuring gain compression. The displayed data is formatted as the phase of S21 versus power. AM-to-PM conversion can be computed by choosing a small amplitude 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-to-PM conversion. The plot above shows AM-to-PM conversion of 0.727 o/dB, centered at an input power of 3.7 dBm and an output power of 27.633 dBm.

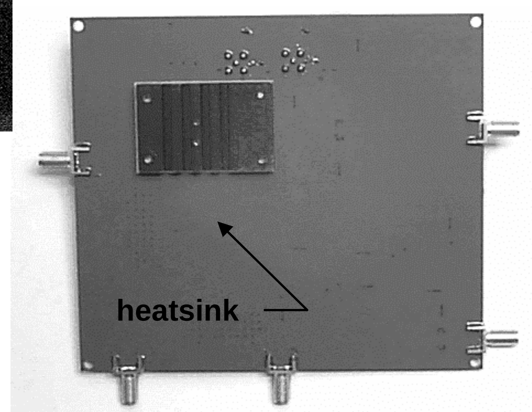
RF Design and Measurement Seminar

Slide #206

Heat Sinking

- for power devices, a heatsink is essential to keep $T_{junction}$ low
- heatsink size depends on material, power dissipation, air flow, and $T_{ambient}$
- ridges or fins increase surface area and help dissipate heat
- usually device attaches directly to heatsink (flange mounts help)
- bolt device in place first, then solder





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For the power amplifier output stage, a heatsink was necessary since the FET was dissipating 2.25 W of DC power (9 V x 250 mA). The heatsink ensures that the junction temperature of the FET does not exceed its maximum value. In general, the size of the heatsink depends on the material used (e.g. copper, aluminum), the power dissipation in the transistor, the air flow around the heat sink, and the ambient temperature during operation. The more air flow available and the lower the operating ambient temperature, the smaller the heatsink can be. Ridges or fins on the heatsink increase surface area and help dissipate heat.

For maximum heat transfer, the active device usually attaches directly to the heatsink. Flange-mount device packages are very useful for this purpose. The FET is soldered to the top side of the PCB and the heatsink is bolted to the bottom side. After the heatsink has been attached to the PCB, it is important to bolt the device in place before soldering the leads. This ensures that no undue stress is placed on the leads which could cause them to detach from the body of the package prematurely. Our heatsink required a small pedestal underneath the FET, since the thickness of the PCB was greater than the distance between the bottom of the leads and the bottom of the FET package.