Conversion to dB scale (I)

A convenient form to express large power or voltage ratios is to use "dB scale," based on logarithms.

$$\mathrm{dB} = 20\log\left(\frac{V_2}{V_1}\right) = 10\log\left(\frac{P_2}{P_1}\right)$$

$$= 20\,(\mathrm{exp}_V) = 10\,(\mathrm{exp}_P)$$

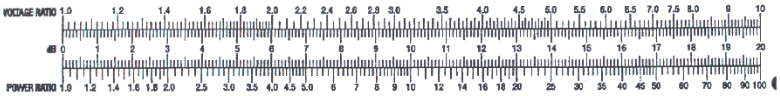
Where, V_1, V_2, P_1, and P_2 are the quantities to be compared, assuming equal terminations

The terms exp_V and exp_P are the 10-based exponents of the voltage and power ratios.

When dB is given, the conversions are:

$$\frac{V_2}{V_1} = 10^{\frac{\mathrm{dB}}{20}} \quad \text{and} \quad \frac{P_2}{P_1} = 10^{\frac{\mathrm{dB}}{20}}$$

VOLTAGE			POWER	
Ratio	Expon.	dB value	Expon.	Ratio
1	10^0	0	10^0	1
1.41	$10^{0.15}$	3	$10^{0.3}$	2
1.73	$10^{0.24}$	4.77	$10^{0.477}$	3
2	$10^{0.30}$	6	$10^{0.6}$	4
3.16	$10^{0.50}$	10	10^1	10
7.07	$10^{0.85}$	17	$10^{1.7}$	50
10	10^1	20	10^2	100
0.707	$10^{-0.15}$	-3	$10^{-0.3}$	0.5
0.5	$10^{-0.30}$	-6	$10^{-0.6}$	.25
0.316	$10^{-0.50}$	-10	10^{-1}	.1
0.1	10^{-1}	-20	10^{-2}	0.01



Summary

Expressing voltage and power ratios in dB simplifies the algebra since multiplication is reduced to addition, and division to subtraction.

To convert exponential power ratio to dB, multiply the exponent by 10.

To convert exponential voltage ratio to dB, multiply the exponent by 20.

Conversion to dB scale (II)

We use "**dB**" to simplify ratios.
The term "**dBm**" refers to an absolute power level.
"0 dBm" is used as a power reference level, defined to 1mW power dissipated in a resistive termination.

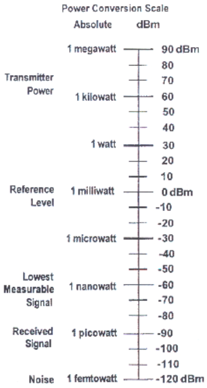
The "**dBW**" scale uses 1W as a reference instead of 1mW. Zero dBW is a power level of 1W.

To find dBm or dBW of an arbitrary power level:

$\text{dBm} = 10\log(P_{mW})$ and $\text{dBm} = 10\log(P_W)$

To get Watts or milliWatts from dBm or dBW:

$$P_{mW} = 10^{\frac{\text{dBm}}{10}} \quad \text{and} \quad P_W = 10^{\frac{\text{dBW}}{10}}$$



Summary

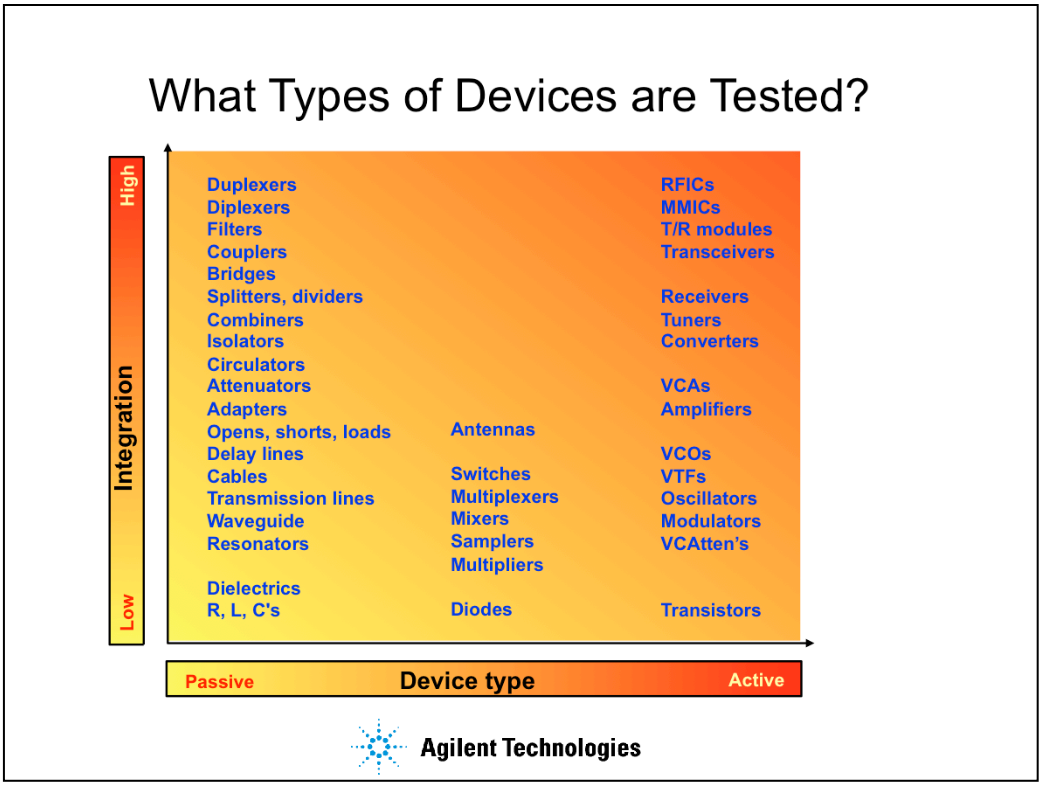
Power reference scale in dBm and dBW

Abs. Power:	**1nW**	**10nW**	**.1µW**	**1µW**	**10µW**	**.1mW**	**1mW**	**10mW**	**.1W**	**1W**	**10W**	**.1kW**	**1kW**
dBm:	-60	-50	-40	-30	-20	-10	0	+10	+20	+30	+40	+50	+60
dBW:	-90	-80	-70	-60	-50	-40	-30	-20	-10	0	+10	+20	+30

Changing units of power (fW → pW → nW → µW → mW → W → KW → MW) always refers to a factor of 1000, or 10^3. Therefore, on the db (or dBm) scale, the corresponding change is always 30 dB (or dBm).

Sometime it is more convenient to use voltage for reference. The "dBV," "dBmV," and "dBµV" scales are referenced to 1V, 1mV, or 1µV applied to a specific resistor. For example, 0dBmV/75Ω refers to 1mV across 75Ω. Or, +20dBµV / 50Ω means 10µV applied to a 50Ω resistor (remember to use 20log, not 10log in the computation).

Vector Network Analyzer basics



What Types of Devices are Tested?

High

Integration

Low

Duplexers
Diplexers
Filters
Couplers
Bridges
Splitters, dividers
Combiners
Isolators
Circulators
Attenuators
Adapters
Opens, shorts, loads
Delay lines
Cables
Transmission lines
Waveguide
Resonators

Dielectrics
R, L, C's

Antennas

Switches
Multiplexers
Mixers
Samplers
Multipliers

Diodes

RFICs
MMICs
T/R modules
Transceivers

Receivers
Tuners
Converters

VCAs
Amplifiers

VCOs
VTFs
Oscillators
Modulators
VCAtten's

Transistors

Passive Device type Active

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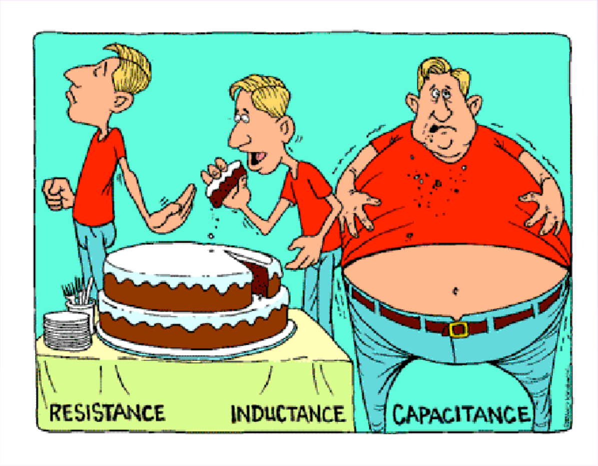
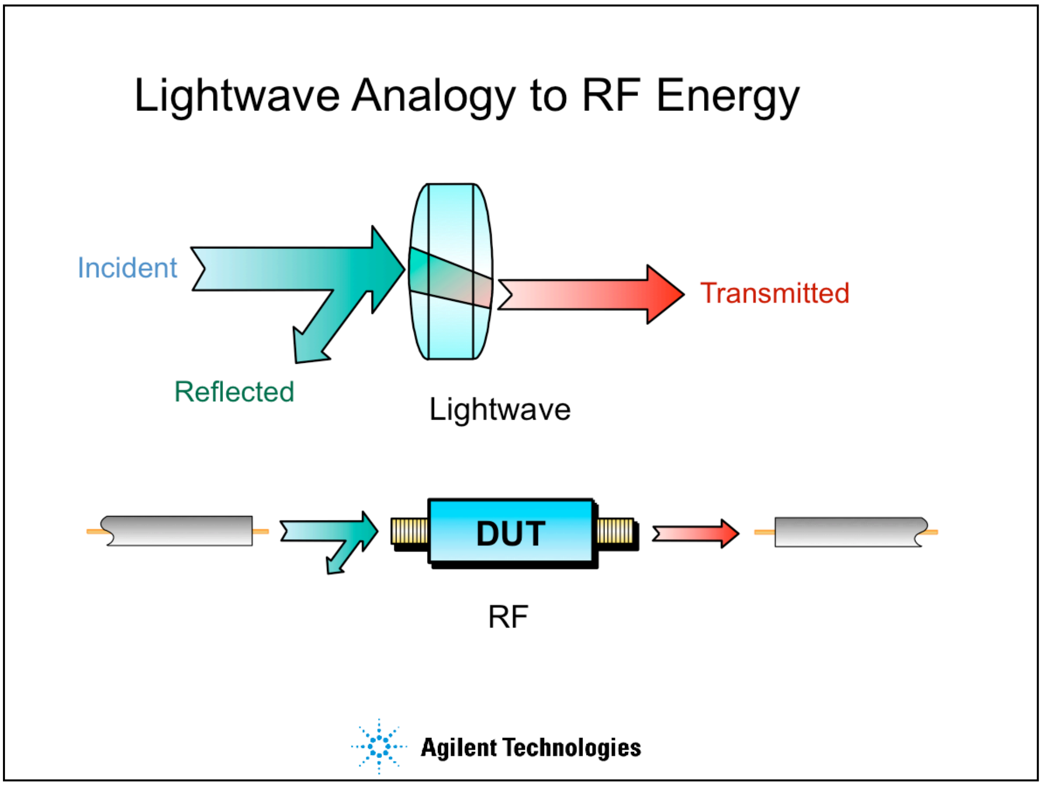
Slide 3

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

Trasmission lines





Lightwave Analogy to RF Energy

Incident

Reflected

Transmitted

Lightwave

DUT

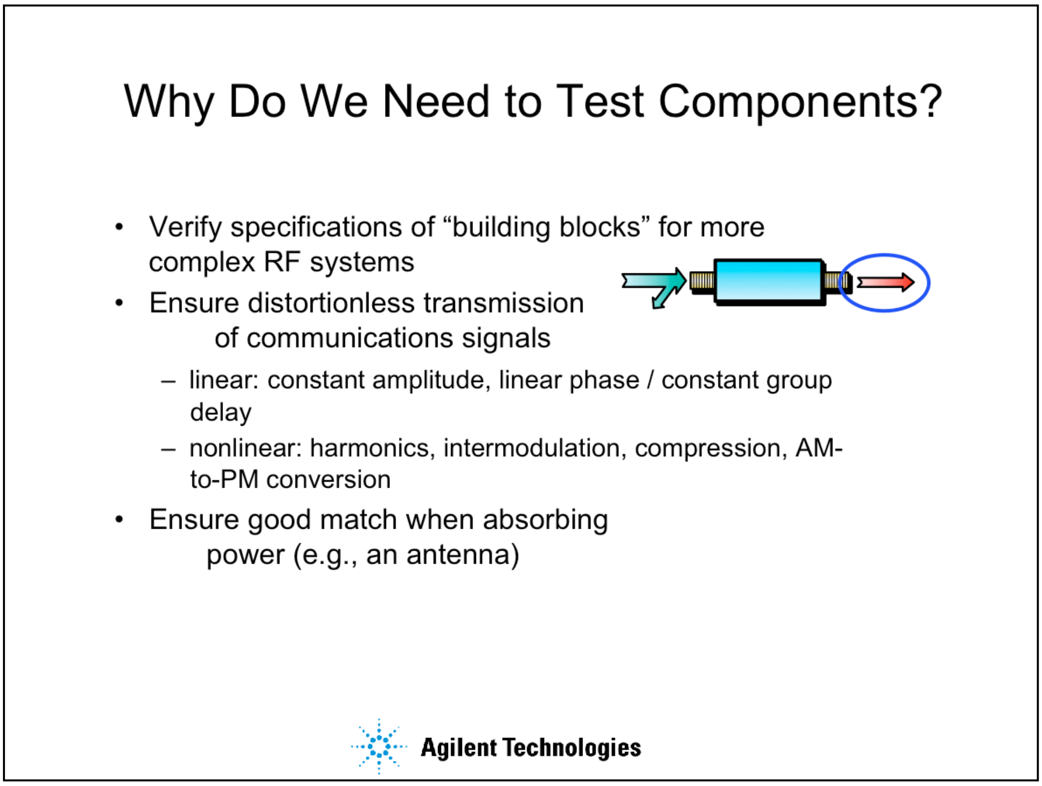
RF

Agilent Technologies

Slide 5

One of the most fundamental concepts 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 were made of some lossy material, then a portion 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 through the lens. 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.

Network analysis is concerned with the accurate measurement of the *ratios* of the reflected signal to the incident signal, and the transmitted signal to the incident signal.



Why Do We Need to Test Components?

- Verify specifications of “building blocks” for more complex RF systems
- Ensure distortionless transmission of communications signals
  - linear: constant amplitude, linear phase / constant group delay
  - nonlinear: harmonics, intermodulation, compression, AM-to-PM conversion
- Ensure good match when absorbing power (e.g., an antenna)

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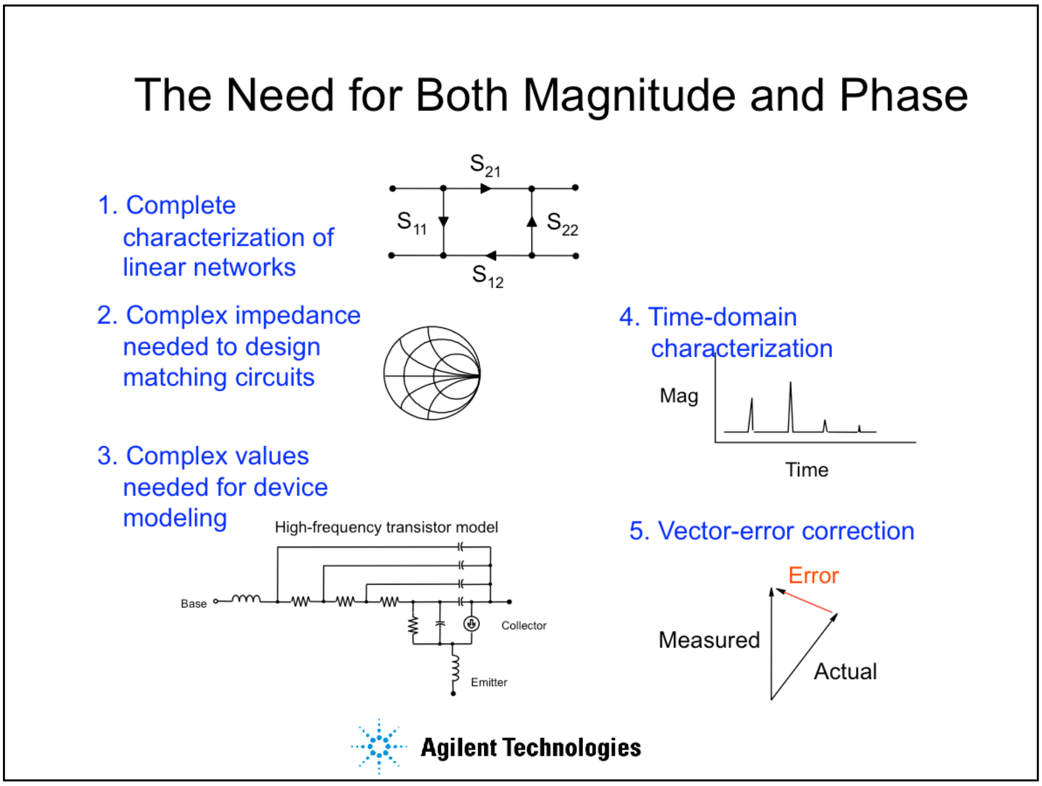
Slide 6

Components are tested for a variety of reasons. Many components are used as "building blocks" in more complicated RF systems. For example, in most transceivers there are amplifiers to boost LO power to mixers, and filters to remove signal harmonics. Often, R&D engineers need to measure these components to verify their simulation models and their actual hardware prototypes. For component production, a manufacturer must measure the performance of their products so they can provide accurate specifications. This is essential so prospective customers will know how a particular component will behave in their application.

When used in communications systems to pass signals, designers want to ensure the component or circuit is not causing excessive signal distortion. This can be in the form of linear distortion where flat magnitude and linear phase shift versus frequency is not maintained over the bandwidth of interest, or in the form of nonlinear effects like intermodulation distortion.

Often it is most important to measure how reflective a component is, to ensure that it absorbs energy efficiently. Measuring antenna match is a good example.



The Need for Both Magnitude and Phase

1. Complete characterization of linear networks
2. Complex impedance needed to design matching circuits
3. Complex values needed for device modeling
4. Time-domain characterization
5. Vector-error correction

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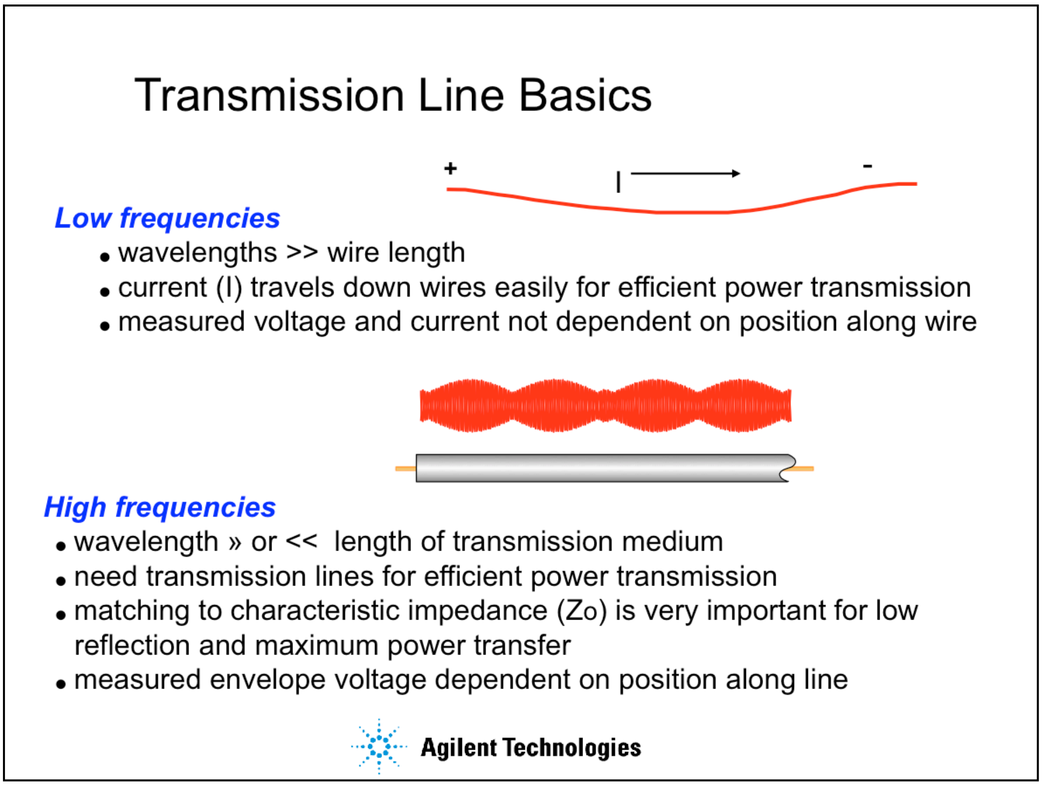
Slide 7

In many situations, magnitude-only data is sufficient for out needs. For example, we may only care about the gain of an amplifier or the stop-band rejection of a filter. However, as we will explore throughout this paper, measuring phase is a critical element of network analysis.

Complete characterization of devices and networks involves measurement of phase as well as magnitude. This is necessary for developing circuit models for simulation and to design matching circuits based on conjugate-matching techniques. Time-domain characterization requires magnitude and phase information to perform the inverse-Fourier transform. Finally, for best measurement accuracy, phase data is required to perform vector error correction.



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

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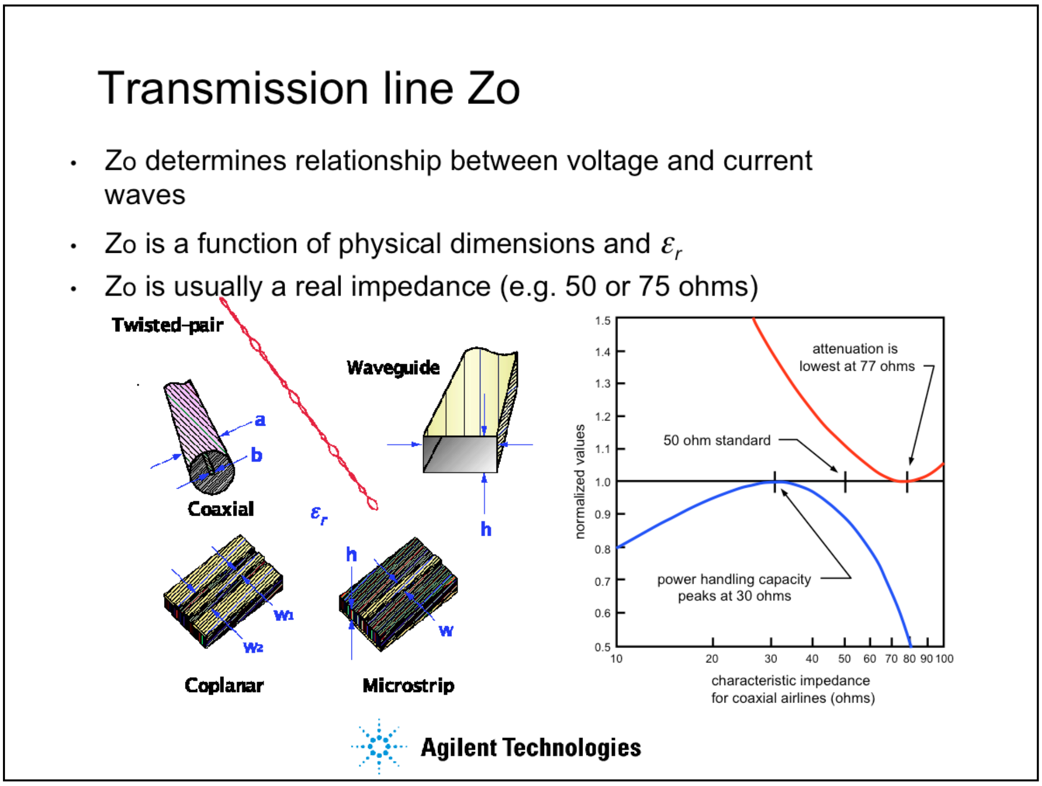
Slide 9

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 (Zo). In fact, an infinitely long transmission line appears to be a resistive load! When the transmission line is terminated in its characteristic impedance, 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 envelope 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



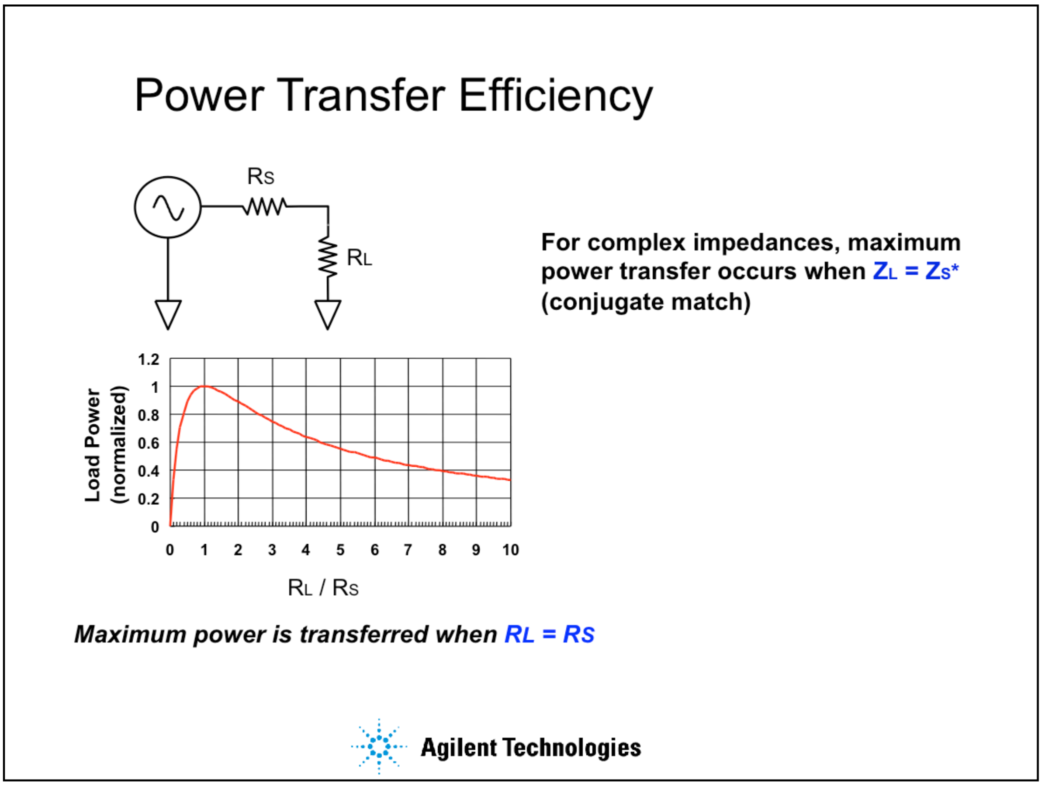
Transmission line Zo

- Zo determines relationship between voltage and current waves
- Zo is a function of physical dimensions and ε_r
- Zo is usually a real impedance (e.g. 50 or 75 ohms)

Slide 10

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 (ε_r) of the non-conducting material in the transmission line. For most RF systems, Zo is either 50 or 75 ohms.

For low-power situations (cable TV, for example) coaxial transmission lines are optimized for low loss, which works out to about 75 ohms (for coaxial transmission lines with air dielectric). For RF and microwave communication and radar applications, where high power is often encountered, coaxial transmission lines are designed to have a characteristic impedance of 50 ohms, a compromise between maximum power handling (occurring at 30 ohms) and minimum loss.



Power Transfer Efficiency

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

Maximum power is transferred when $R_L = R_S$

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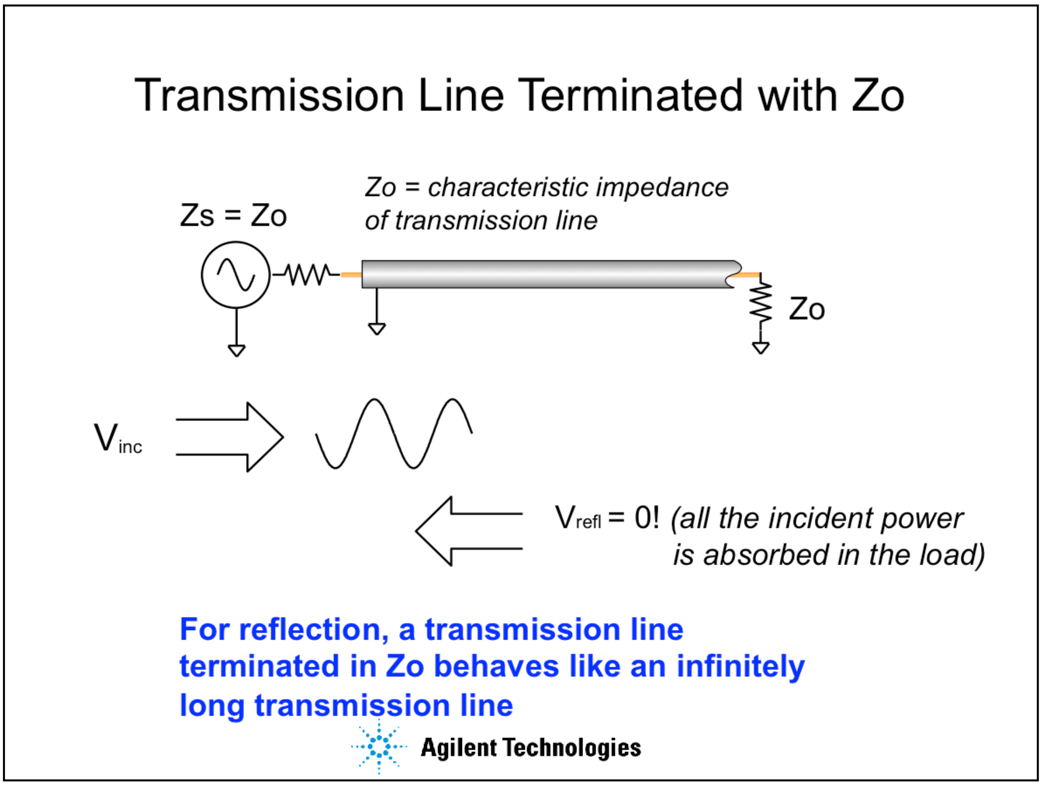
Slide 11

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 (RL = RS) 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 RS = 0.6 + j0.3, then the complex conjugate RS* = 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



Transmission Line Terminated with Zo

Zs = Zo

Zo = characteristic impedance of transmission line

Zo

V_{inc}

V_{refl} = 0! *(all the incident power is absorbed in the load)*

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

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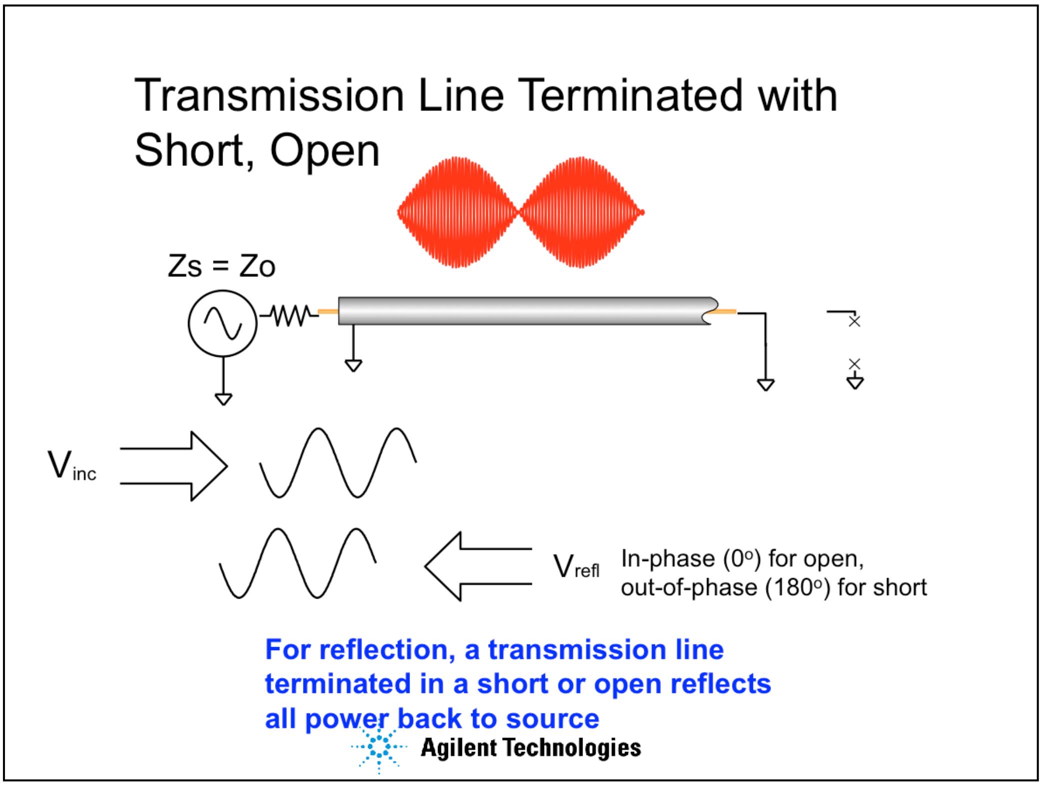
Slide 12

Let's review what happens when transmission lines are terminated in various 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.



Transmission Line Terminated with Short, Open

Zs = Zo

V_{inc}

V_{refl} In-phase (0°) for open,
out-of-phase (180°) for short

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

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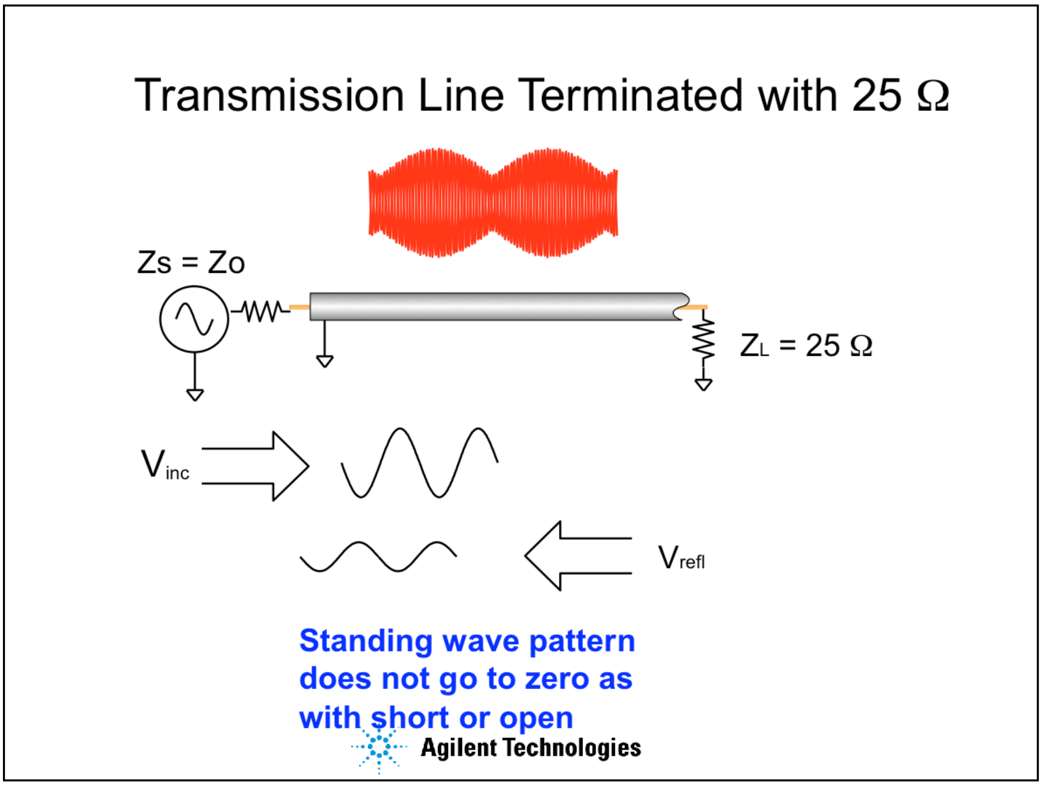
Slide 13

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.



Transmission Line Terminated with 25 Ω

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

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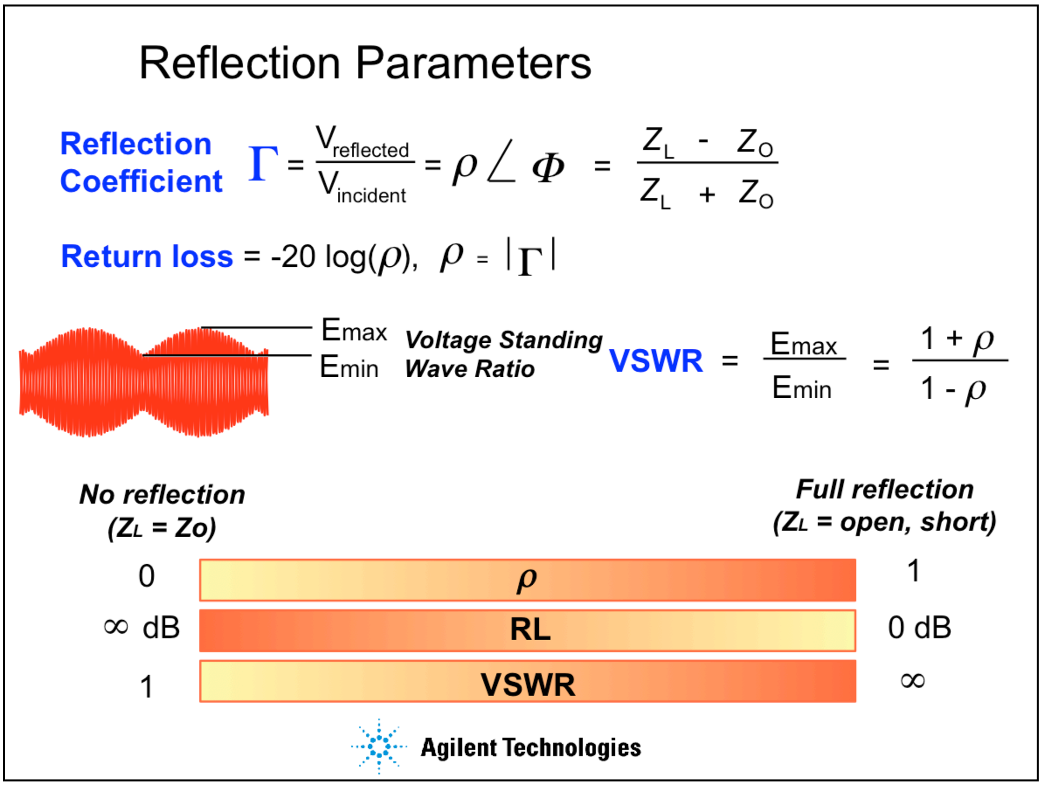
Slide 14

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 be 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° 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, as does the apparent impedance.



Reflection Parameters

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

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

Voltage Standing Wave Ratio **VSWR** $= \frac{E_{max}}{E_{min}} = \frac{1 + \rho}{1 - \rho}$

No reflection* *(Z_L = Zo)		***Full reflection* *(Z_L = open, short)***
0	ρ	1
∞ dB	**RL**	0 dB
1	**VSWR**	∞

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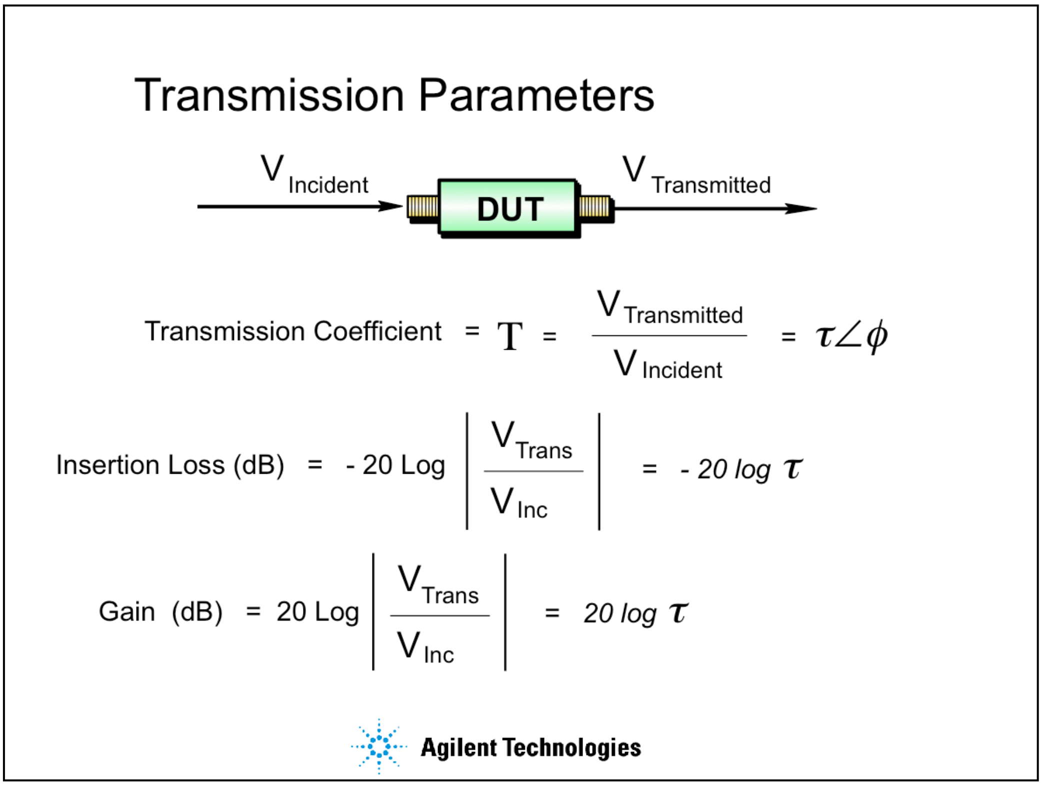
Slide 16

Let's now examine reflection measurements. The first term for reflected waves is reflection coefficient gamma (Γ). Reflection coefficient is the ratio of the reflected signal voltage to the incident signal voltage. It can be calculated as shown above by knowing the impedances of the transmission line and the load. The magnitude portion of gamma is called rho (ρ). A transmission line terminated in Zo will have all energy transferred to the load; hence $V_{refl} = 0$ and $\rho = 0$. When Z_L is not equal to Zo , some energy is reflected and ρ is greater than zero. When Z_L is a short or open circuit, all energy is reflected and $\rho = 1$. The range of possible values for ρ is therefore 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 transmission line cause a "standing wave". This condition can be measured in terms of the voltage-standing-wave ratio (VSWR or SWR for short). VSWR 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



Transmission Parameters

$V_{Incident}$ DUT $V_{Transmitted}$

$$\text{Transmission Coefficient} = \mathrm{T} = \frac{V_{Transmitted}}{V_{Incident}} = \tau\angle\phi$$

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

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

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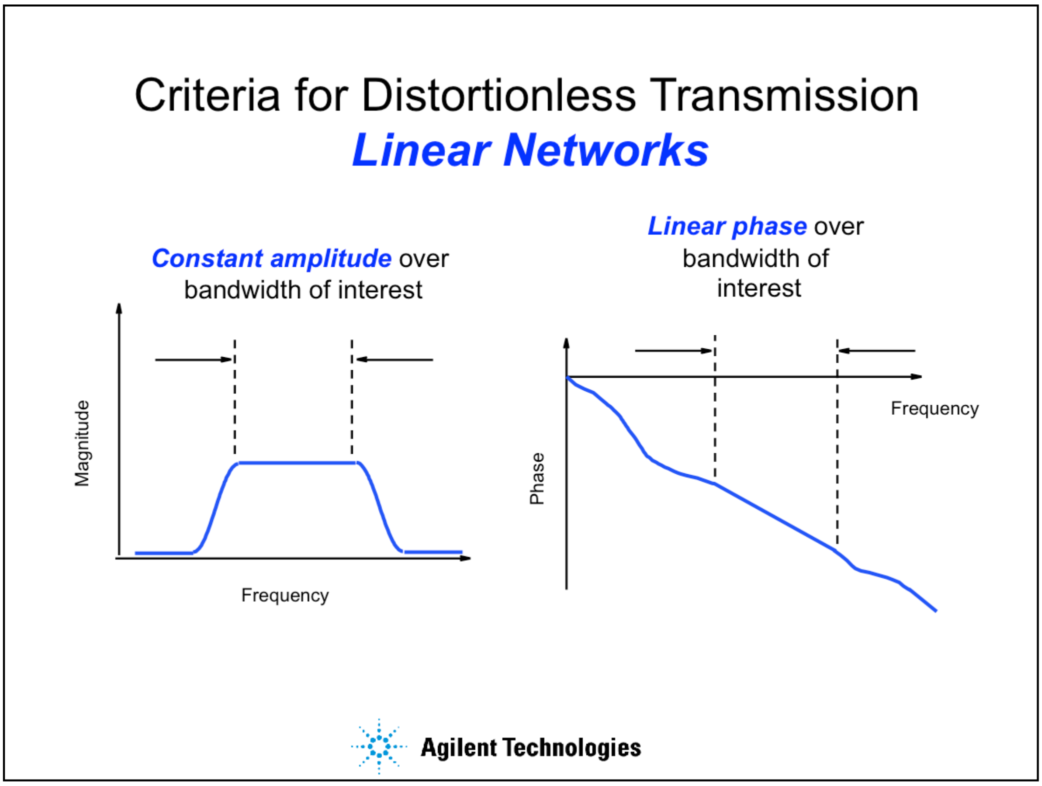
Slide 18

Transmission coefficient T is defined as the transmitted voltage divided by the incident voltage. If $|V_{trans}| > |V_{inc}|$, the DUT has gain, and if $|V_{trans}| < |V_{inc}|$, the DUT exhibits 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. In communications systems, signals are time varying -- they occupy a given bandwidth and are made up of multiple frequency components. It is important then to know to what extent the DUT alters the makeup of the signal, thereby causing signal distortion. While we often think of distortion as only the result of nonlinear networks, we will see shortly that linear networks can also cause signal distortion.



Criteria for Distortionless Transmission
Linear Networks

Constant amplitude over bandwidth of interest

Linear phase over bandwidth of interest

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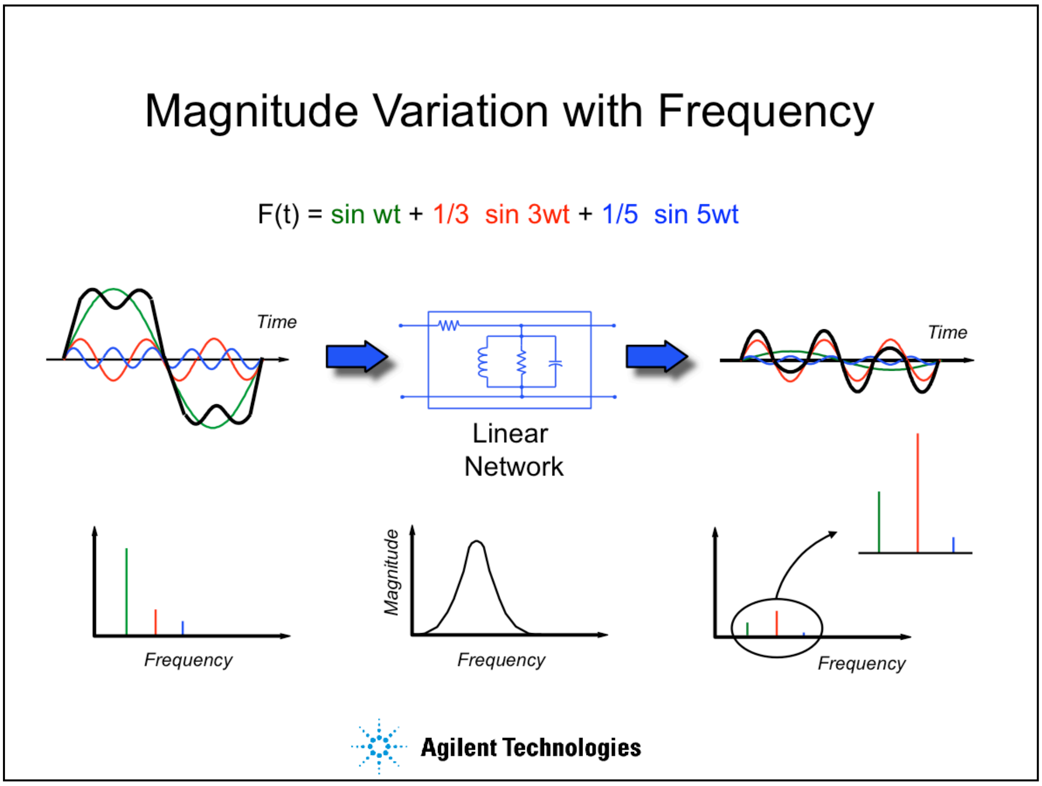
Slide 20

Now lets examine how linear networks can cause signal distortion. There are three 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. And last, the device must exhibit a "minimum-phase response", which means that at 0 Hz (DC), there is 0° phase shift
(0° ± n*180° is okay if we don't mind an inverted signal).

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

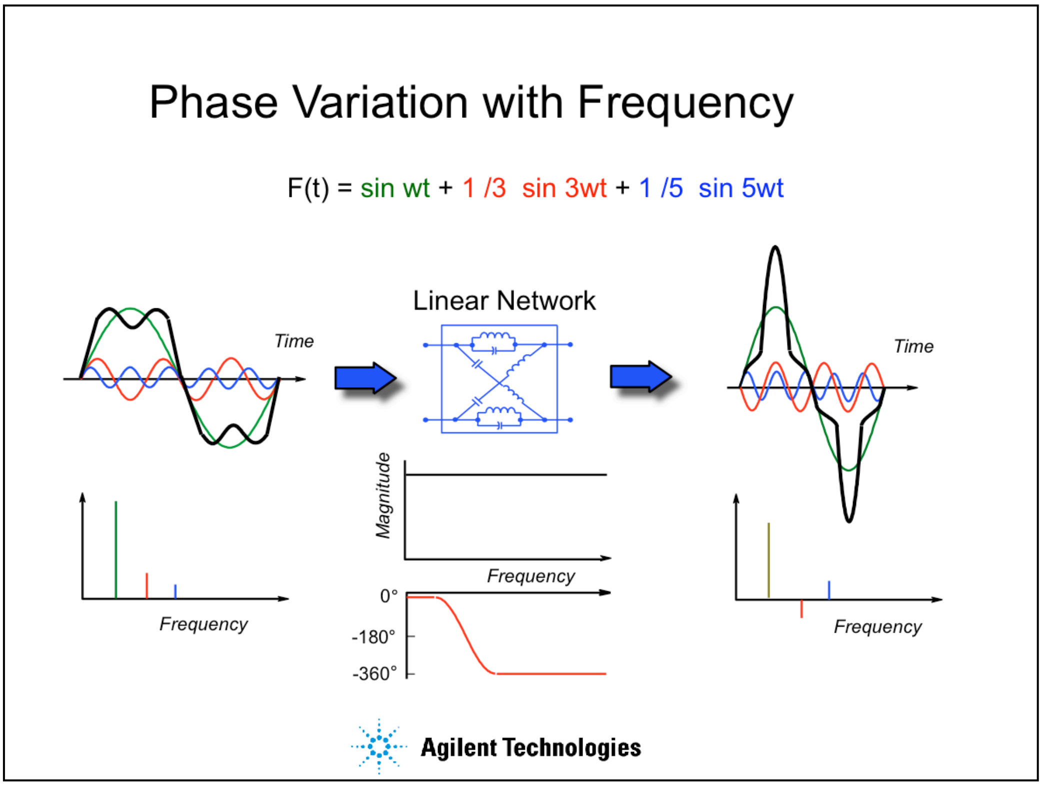


Magnitude Variation with Frequency

$F(t) = \sin wt + 1/3 \sin 3wt + 1/5 \sin 5wt$

Slide 21

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.



Phase Variation with Frequency

$F(t) = \sin wt + 1/3 \sin 3wt + 1/5 \sin 5wt$

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Slide 22

Let's apply the same square wave to another filter. Here, the third harmonic undergoes a 180º phase shift, but the other components are not phase shifted. 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.

Scattering parameters

Characterizing Unknown Devices

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

- gives 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
- 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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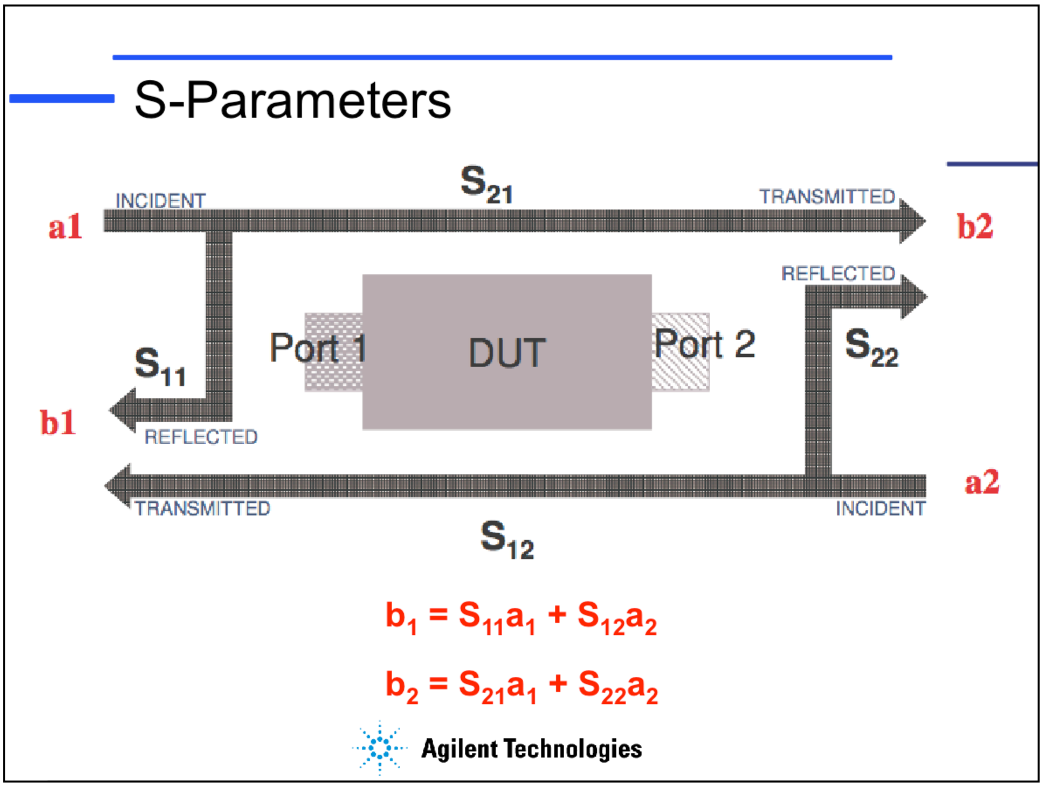
Slide 26

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.



S-Parameters

$$b_1 = S_{11}a_1 + S_{12}a_2$$

$$b_2 = S_{21}a_1 + S_{22}a_2$$

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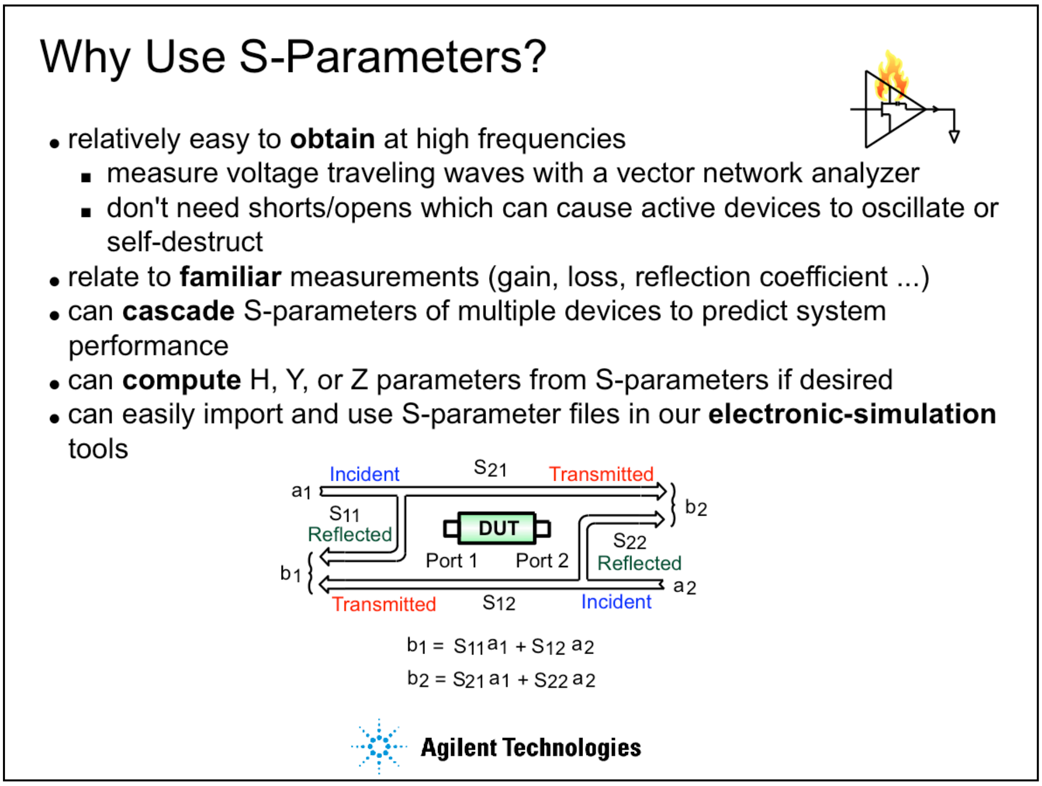
For the two port device there are two independent equations may be written, expressing the independent variable b in terms of the dependant variable a. In the above diagram b_1 comprises the sum of a quantity reflected from port 1 and a quantity that is the result of transmission through the device in the reverse direction. The quantities are scaled to be proportional to the voltage wave amplitude and phase such that $|b_n|^2$ = power emerging from the n’th port and $|a_n|^2$ is the power incident on the n’th port.



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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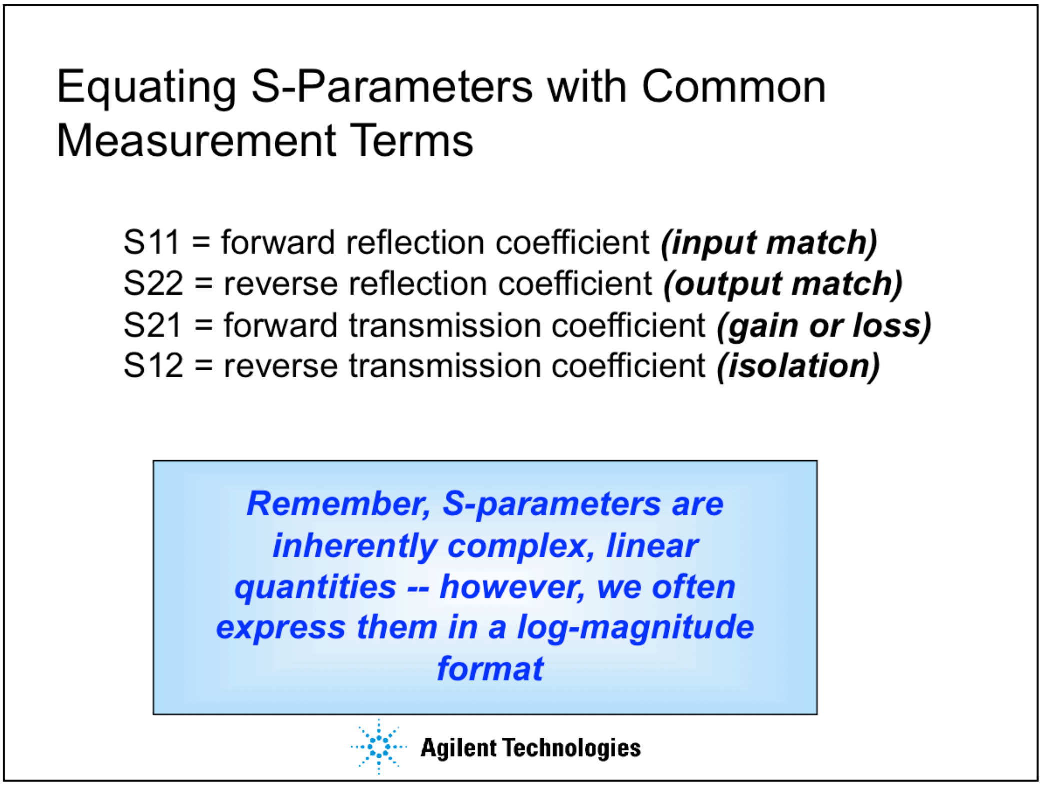
At high frequencies, it is very hard to measure total voltage and current at the device ports. One cannot simply connect a voltmeter or current probe and get accurate measurements due to the impedance of the probes themselves and the difficulty of placing the probes at the desired positions. In addition, 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. 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 for circuit simulations in electronic-design automation (EDA) tools like Agilent's Advanced Design System (ADS). S-parameters are the shared language between simulation and measurement.

An N-port device 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, S21 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., S11), it indicates a reflection measurement, as the input and output ports are the same. The incident terms (a1, a2) and output terms (b1, b2) represent voltage traveling waves.



Equating S-Parameters with Common Measurement Terms

S11 = forward reflection coefficient ***(input match)***
S22 = reverse reflection coefficient ***(output match)***
S21 = forward transmission coefficient ***(gain or loss)***
S12 = reverse transmission coefficient ***(isolation)***

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

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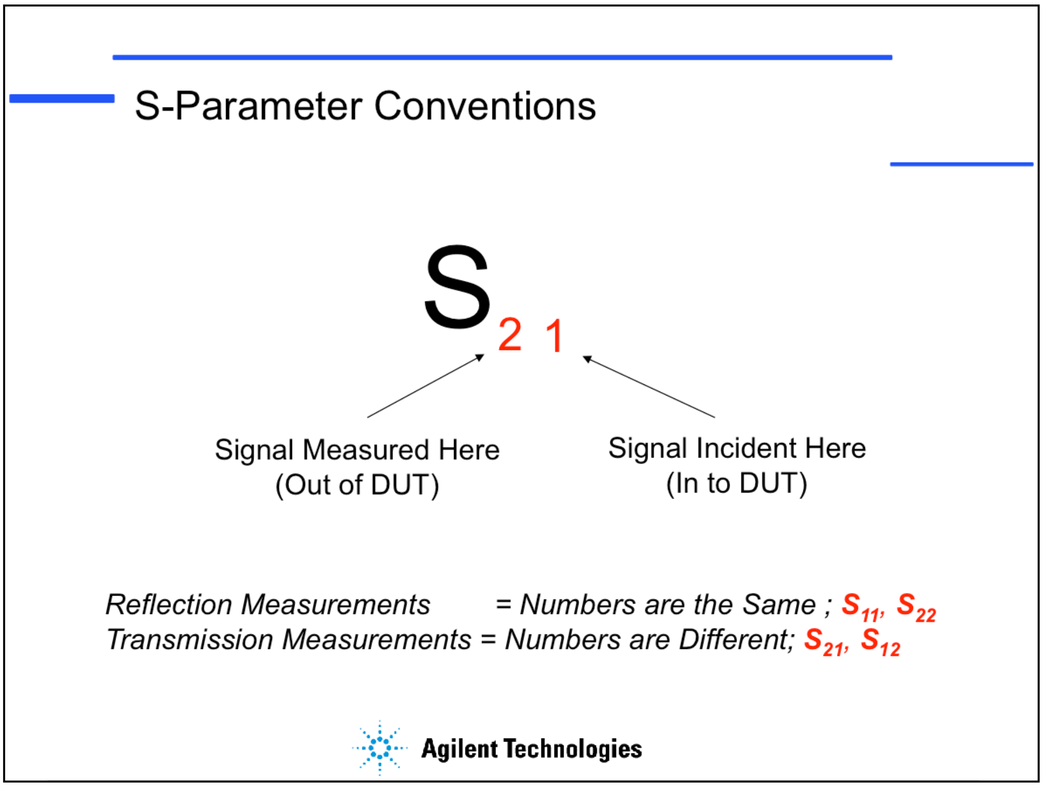
S-parameters are essentially the same parameters as some of the terms we have mentioned before, such as input match and insertion loss. It is important to separate the fundamental definition of S-parameters and the format in which they are often displayed. S-parameters are inherently complex, linear quantities. They are expressed as real-and-imaginary or magnitude-and-phase pairs. However, it isn't always very useful to view them as linear pairs. Often we want to look only at the magnitude of the S-parameter (for example, when looking at insertion loss or input match), and often, a logarithmic display is most useful. A log-magnitude format lets us see far more dynamic range than a linear format.



S-Parameter Conventions

$S_{2\,1}$

Signal Measured Here
(Out of DUT)

Signal Incident Here
(In to DUT)

Reflection Measurements = Numbers are the Same ; S_{11}, S_{22}
Transmission Measurements = Numbers are Different; S_{21}, S_{12}

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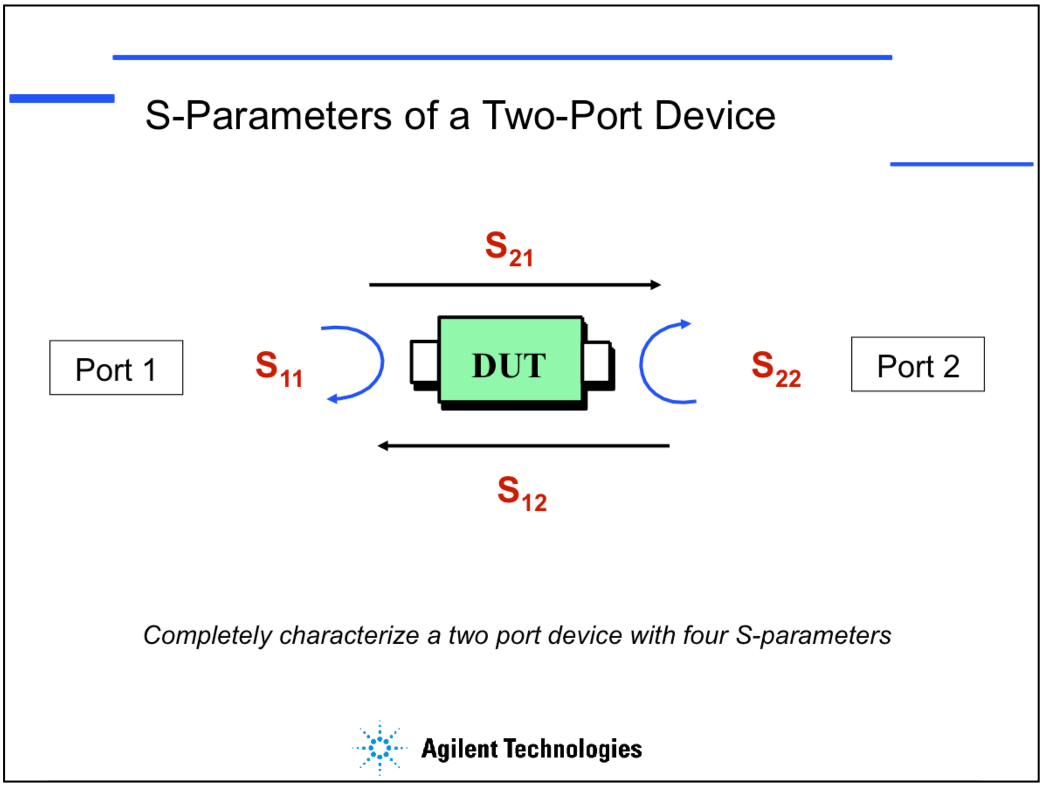
To quantify and express these reflected and transmitted coefficients, the scattering or S-parameter were defined. S-parameters give a simple way to quantify and organize these ratios of voltages exiting or scattering from a network relative to the voltages incident upon the network.

S-parameters are always a ratio of two complex (magnitude and phase) quantities. S-parameter notation identifies these quantities using the numbering scheme shown above. The first number refers to the test-device port where the signal is emerging, or another way to look at it, which network analyzer port is the signal being measured. The second number refers to which test-device port the signal is incident or which network analyzer port the signal is coming from. For example, the S-parameter, S11, identifies the measurement as the complex ratio of the signal emerging from port 1 of the device to the signal applied to port 1 of the device (a reflection measurement).



S-Parameters of a Two-Port Device

S_{21}

Port 1

S_{11}

DUT

S_{22}

Port 2

S_{12}

Completely characterize a two port device with four S-parameters

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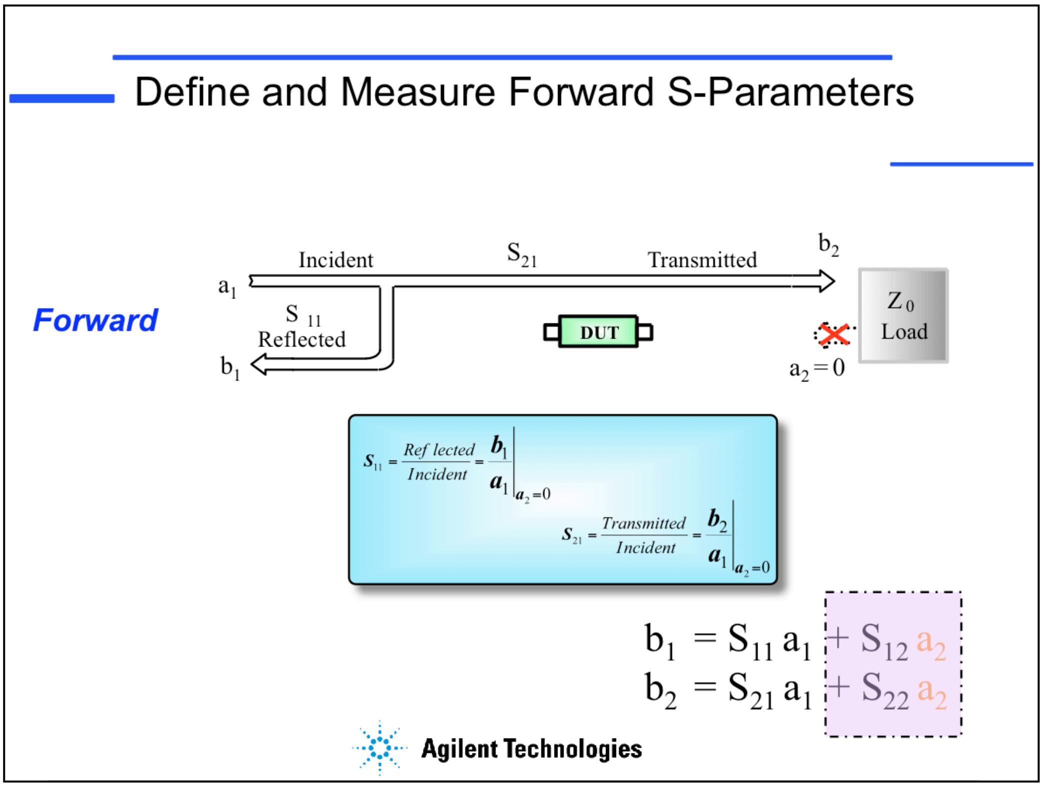
A two-port device or network has four S-parameters. Two of the terms are related to the reflection from the input and output ports of the DUT. The other two terms are related to the transmission through the DUT in the forward and reverse directions. These concepts can be expanded to multi-port devices and the number of S-parameters is a function of 2^n, where n = the number of ports. For example, a four port device would have 16 S-parameters.



Define and Measure Forward S-Parameters

Forward

$$S_{11} = \frac{Reflected}{Incident} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$$

$$S_{21} = \frac{Transmitted}{Incident} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$$

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

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To define S-parameters, it is assumed that a sinewave is incident upon one port of the device while the other port is terminated by a Z_0 load.

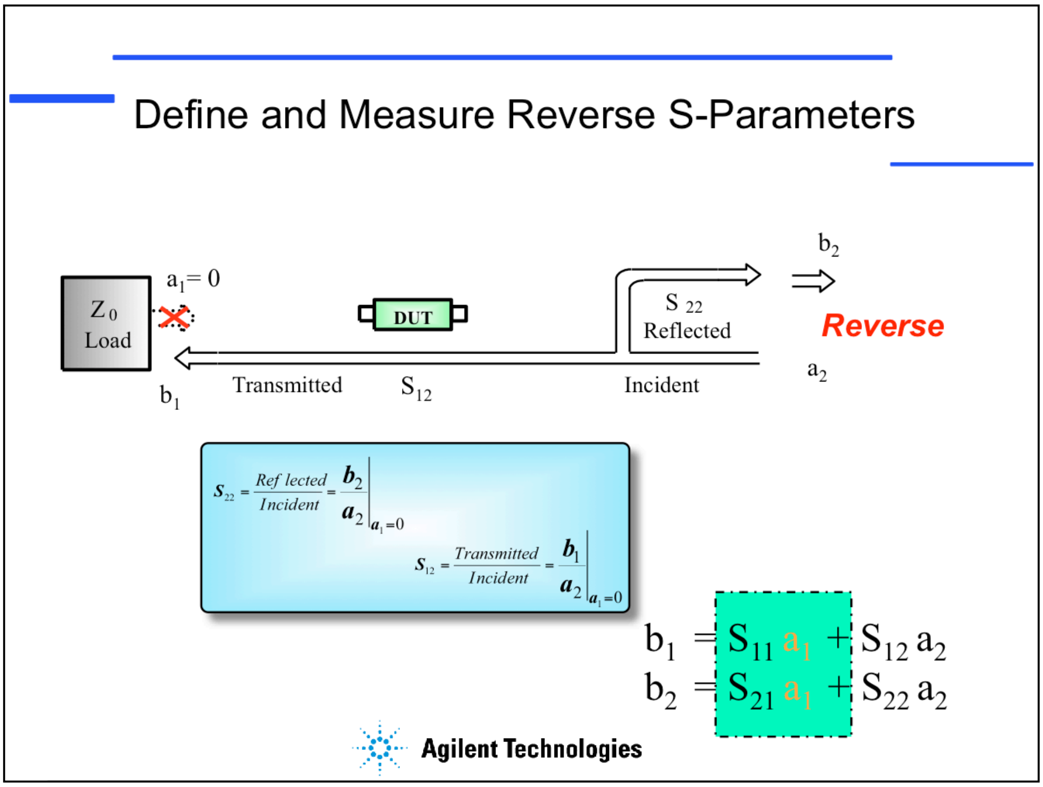
$$s_{21} = \text{Transmission Coefficient} = \frac{\text{Transmitted}}{\text{Incident}} = \left.\frac{b_2}{a_1}\right|_{a_2=0}$$

$$s_{11} = \text{Reflection Coefficient} = \frac{\text{Reflected}}{\text{Incident}} = \left.\frac{b_1}{a_1}\right|_{a_2=0}$$

For a 2-port device, there will be 2 scenarios:

- 1) The stimulus is incident upon port 1 of the device. This will be call the *forward* condition, defined above.
- 2) The stimulus is incident upon port 2 *(reverse)*.

The test system must be able to sample the incident signal seen as quantity a_1 in the forward direction and a_2 in the reverse direction. The setup must also measure the signal reflected and transmitted (b). Lastly, to yield S-parameters, the system must be able to take the complex (magnitude and phase) ratio of the reflected or transmitted signal versus the incident signal (b/a). The measurements and ratios must be complex to produce true S-parameters.



Define and Measure Reverse S-Parameters

$$S_{22} = \frac{Reflected}{Incident} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$$

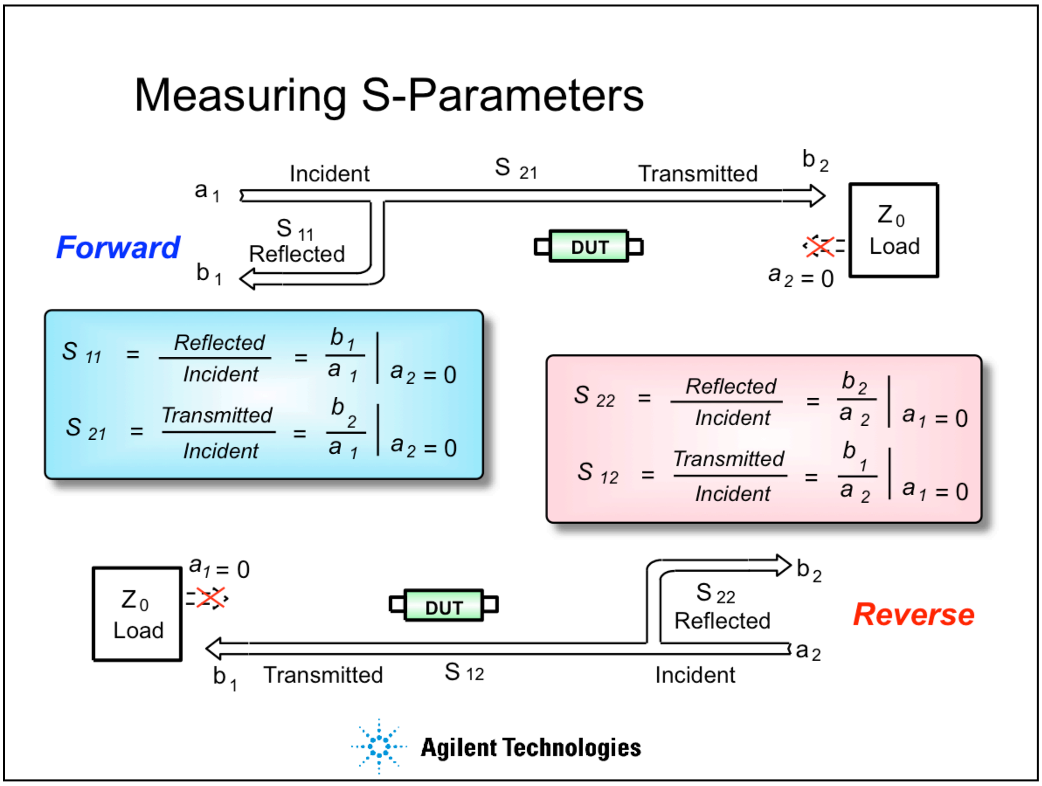
$$S_{12} = \frac{Transmitted}{Incident} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$$

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

With the stimulus at port 2, (reverse condition):

$$s_{12} = \text{Transmission Coefficient} = \frac{\text{Transmitted}}{\text{Incident}} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$$

$$s_{22} = \text{Reflection Coefficient} = \frac{\text{Reflected}}{\text{Incident}} = \left.\frac{b_2}{a_2}\right|_{a_1=0}$$



Measuring S-Parameters

$$S_{11} = \frac{\text{Reflected}}{\text{Incident}} = \left.\frac{b_1}{a_1}\right|_{a_2 = 0}$$

$$S_{21} = \frac{\text{Transmitted}}{\text{Incident}} = \left.\frac{b_2}{a_1}\right|_{a_2 = 0}$$

$$S_{22} = \frac{\text{Reflected}}{\text{Incident}} = \left.\frac{b_2}{a_2}\right|_{a_1 = 0}$$

$$S_{12} = \frac{\text{Transmitted}}{\text{Incident}} = \left.\frac{b_1}{a_2}\right|_{a_1 = 0}$$

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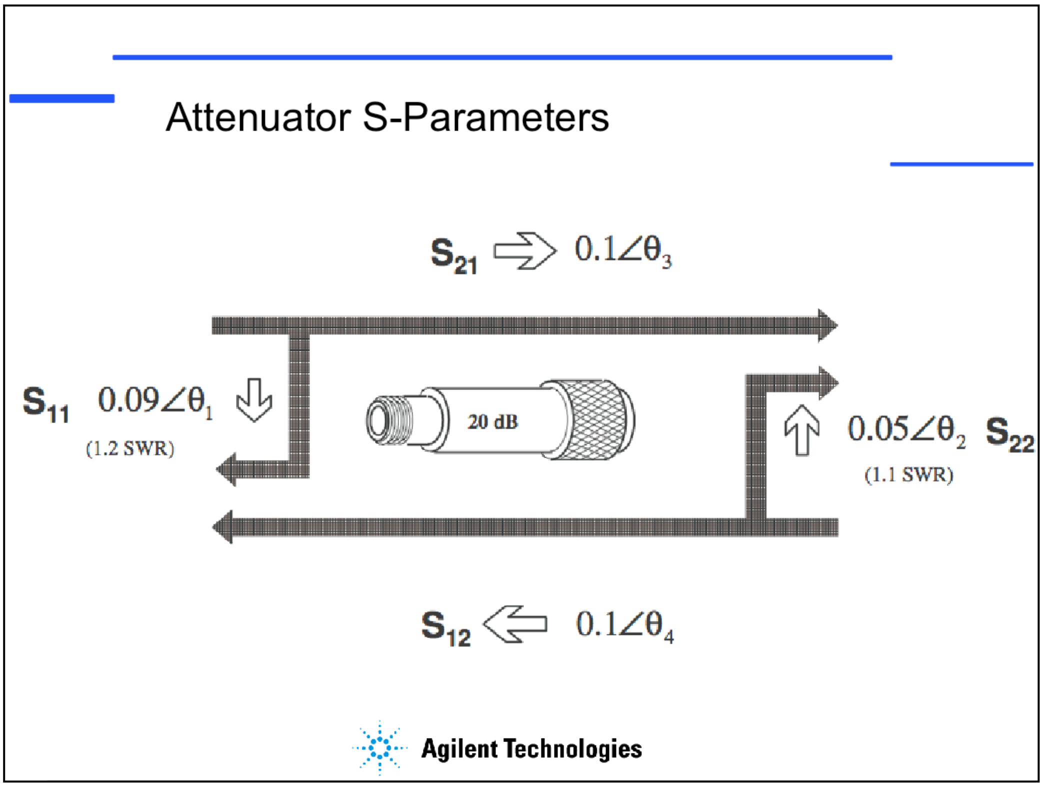
S11 and S21 are determined by measuring the magnitude and phase of the incident, reflected and transmitted voltage signals when the output is terminated in a perfect Zo (a load that equals the characteristic impedance of the test system). This condition guarantees that a_2 is zero, since there is no 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 load port (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).



Attenuator S-Parameters

$S_{21} \Rightarrow 0.1\angle\theta_3$

S_{11} $0.09\angle\theta_1$ (1.2 SWR)

$0.05\angle\theta_2$ S_{22} (1.1 SWR)

$S_{12} \Leftarrow 0.1\angle\theta_4$

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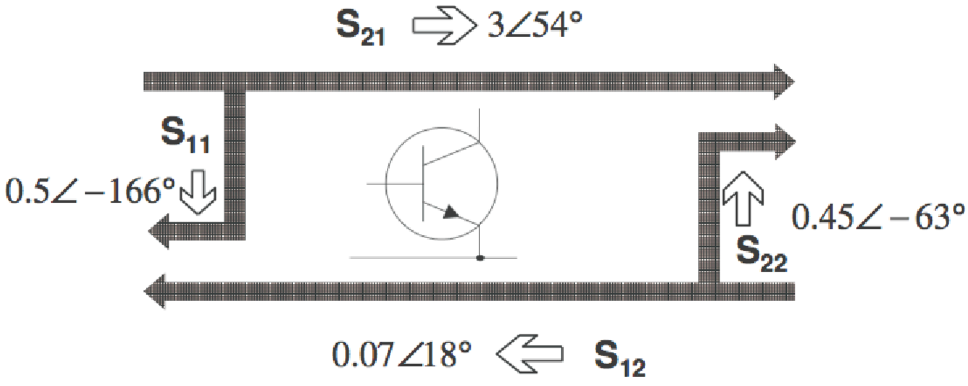
Let’s look at some examples of S-parameters for typical devices. Shown here is a 20 dB attenuator. Remembering that S-parameters are voltage ratios, it makes sense that the forward transmission S-parameter, S_{21}, would be .1 with some phase shift from incident to output. An attenuator is just a voltage divider, so its reverse transmission characteristics, $S_{12,}$ will be similar. One of the attenuator’s design goals is to be well matched to Z_0. Therefore, the reflection S-parameters, $S_{11,}$ and S_{22} will be small (0.09 and 0.05 respectively).

Once these linear ratios have been measured, they can easily be converted to other common formats such as standing wave ratio (SWR), or expressed as gain or loss parameters in dB.

Transistor S-parameters



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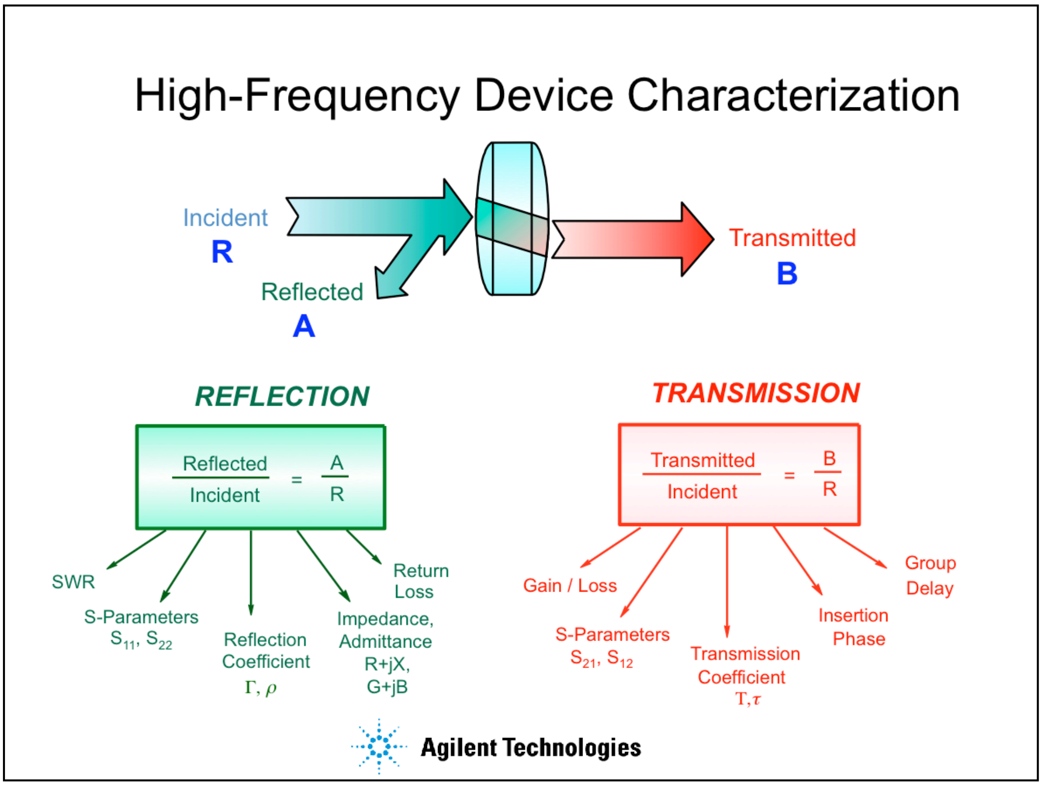
A transistor will have a very different set of S-parameters from an attenuator. Notice, in the forward direction, S_{21} is a positive whole number denoting device gain, where in the reverse direction the device is lossy. The reflection S-parameters indicate a large percent of the incident signal is reflected. This is typical of transistors to have a poor match and require a matching networks to couple signals into and out of the device.

We can start to see the value of making these measurements. From here the designer can create matching networks for the transistor to maximize signal transfer, gain and stability of the combined or cascaded networks.



High-Frequency Device Characterization

Incident
R

Reflected
A

Transmitted
B

REFLECTION

$$\frac{\text{Reflected}}{\text{Incident}} = \frac{A}{R}$$

SWR

S-Parameters
S_{11}, S_{22}

Reflection
Coefficient
Γ, ρ

Impedance,
Admittance
R+jX,
G+jB

Return
Loss

TRANSMISSION

$$\frac{\text{Transmitted}}{\text{Incident}} = \frac{B}{R}$$

Gain / Loss

S-Parameters
S_{21}, S_{12}

Transmission
Coefficient
T, τ

Insertion
Phase

Group
Delay

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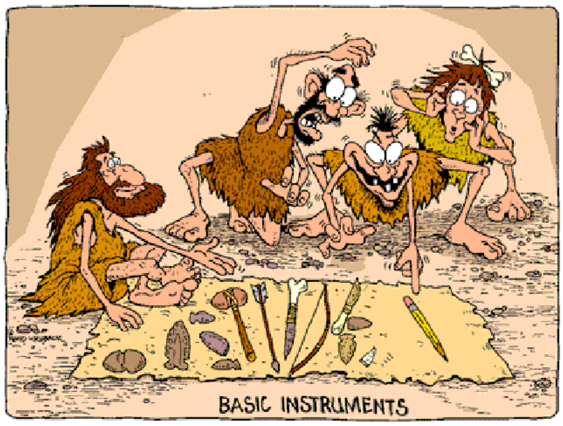
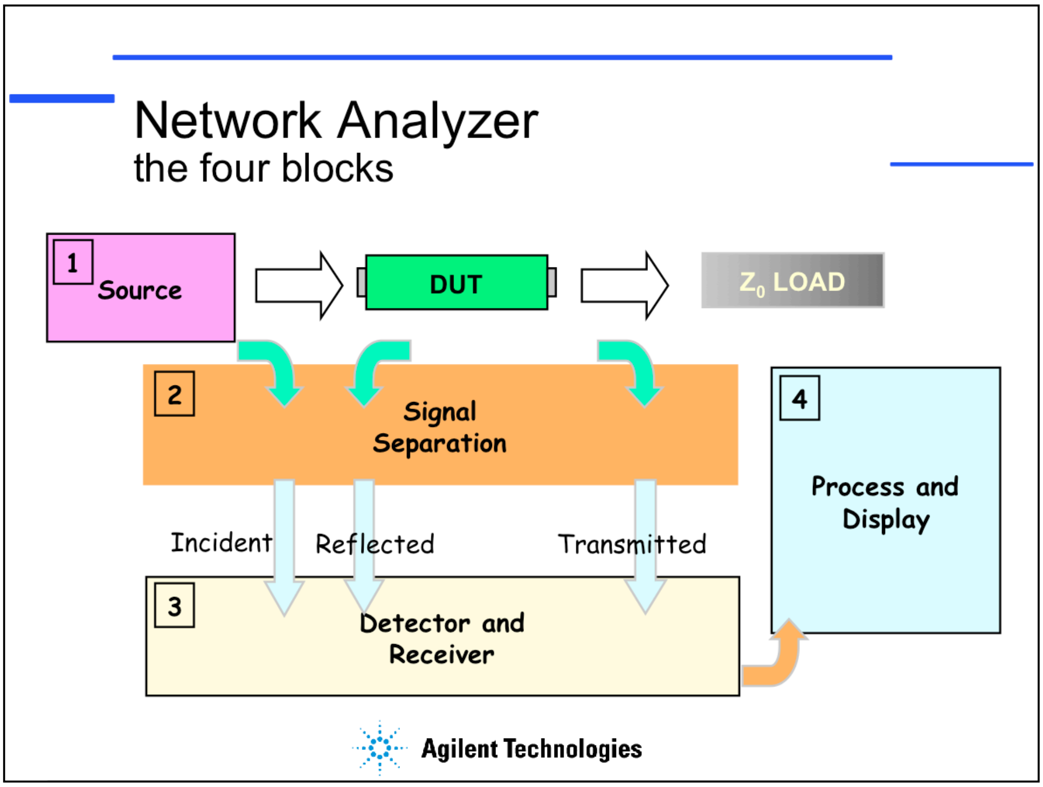
Now that we fully understand the relationship of electromagnetic waves, we must also recognize the terms used to describe them. Common network analyzer terminology has the incident wave measured with the R (for reference) receiver. The reflected wave is measured with the A receiver and the transmitted wave is measured with the B receiver. 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 results from a vector reflection measurement. Some, like group delay, are purely phase-related measurements.

Ratioed reflection is often shown as A/R and ratioed transmission is often shown as B/R, relating to the measurement receivers used in the network analyzer

Modern VNA





Network Analyzer

the four blocks

1 Source

DUT

Z_0 LOAD

2 Signal Separation

Incident

Reflected

Transmitted

3 Detector and Receiver

4 Process and Display

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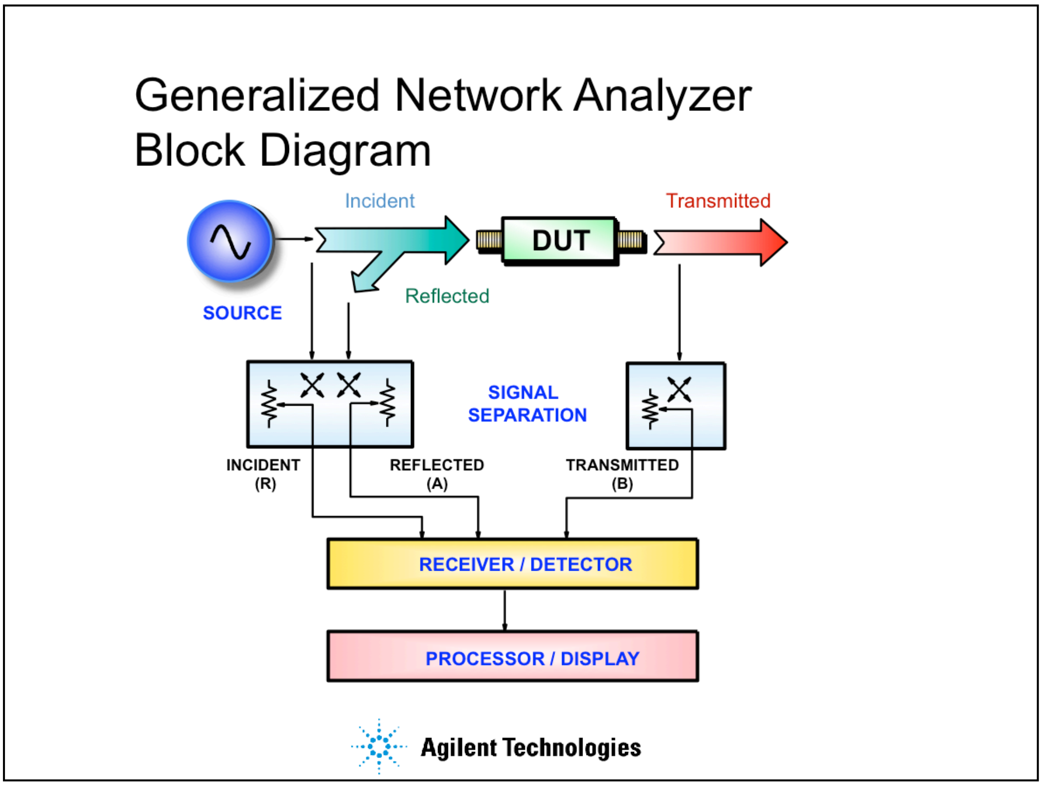

Generalized Network Analyzer Block Diagram

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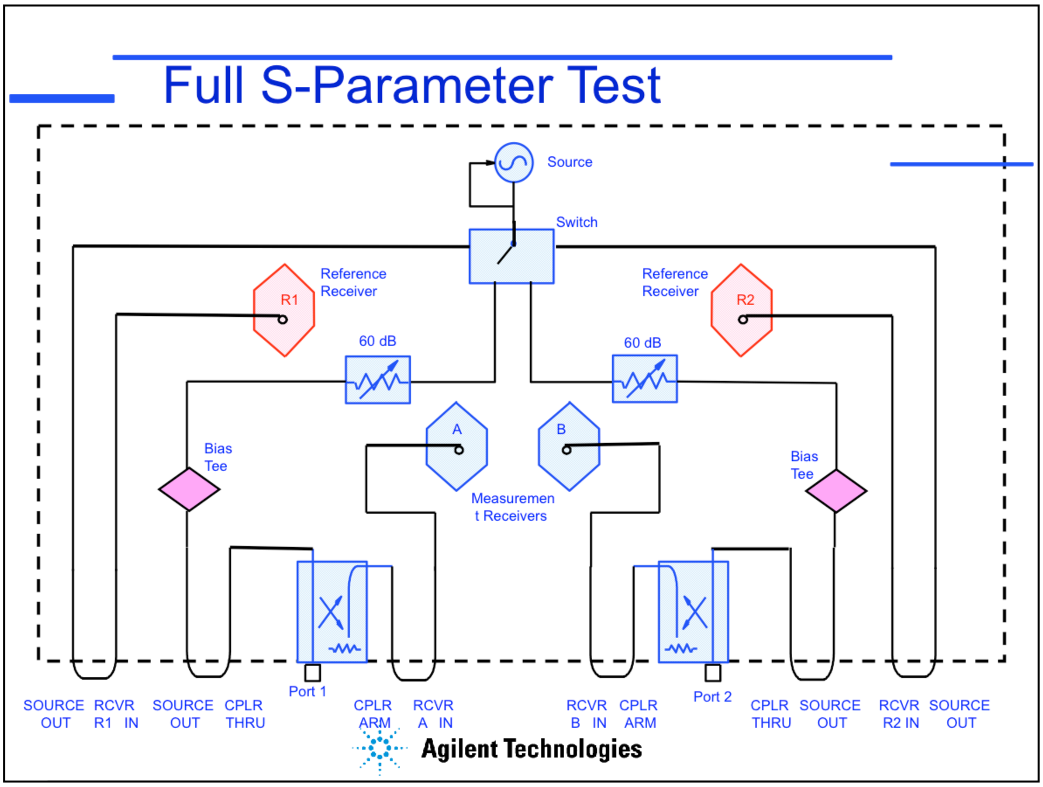
Here is a generalized block diagram of a network analyzer, showing the major signal-processing sections. In order to measure the incident, reflected and transmitted signal, four sections are required:

- Source for stimulus
- Signal-separation devices
- Receivers that downconvert and detect the signals
- Processor/display for calculating and reviewing the results

We will briefly examine each of these sections. More detailed information about the signal separation devices and receiver section are in the appendix.



Full S-Parameter Test

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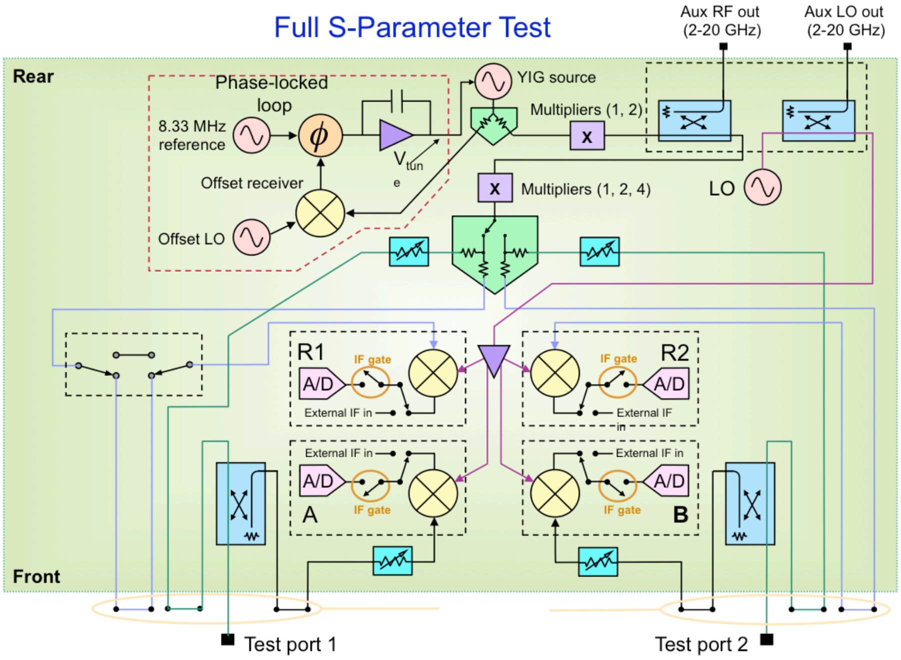
Full S-Parameter Test



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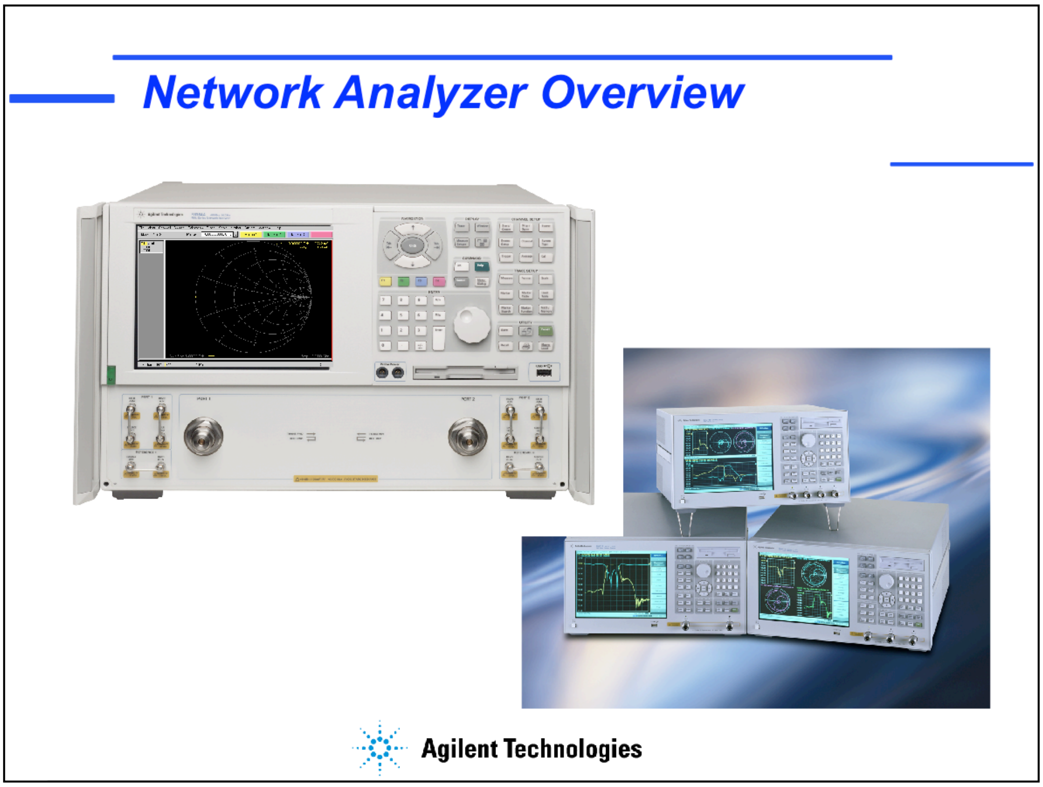

Network Analyzer Overview

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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. Error modeling, accuracy enhancement, and various calibration techniques will then be presented. Finally, some typical impedance matching examples will be covered.