

# Keysight Technologies

## Techniques for Advanced Cable Testing

### Using FieldFox handheld analyzers

#### Application Note

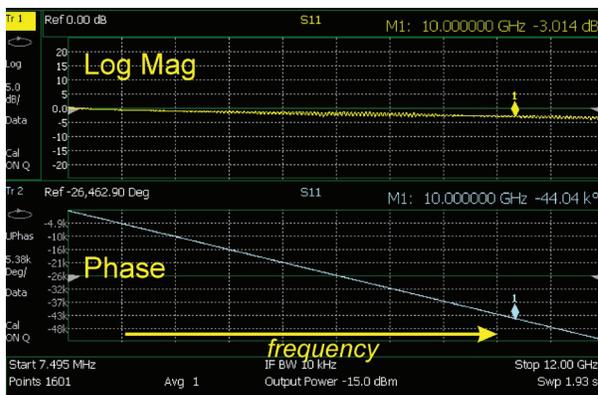


Transmission lines are used to guide the flow of energy from one point to another. Line types include coaxial cable, two-wire, waveguide and a variety of printed circuit configurations. Transmission lines, designed for a specific characteristic impedance, are used to interconnect antennas with communications hardware, distribute signals over large geographic areas or connect integrated circuits on a printed circuit board, to name just a few of the possible applications. Systems that use transmission lines include wireless and data communications, radar, medical and broadcast entertainment.

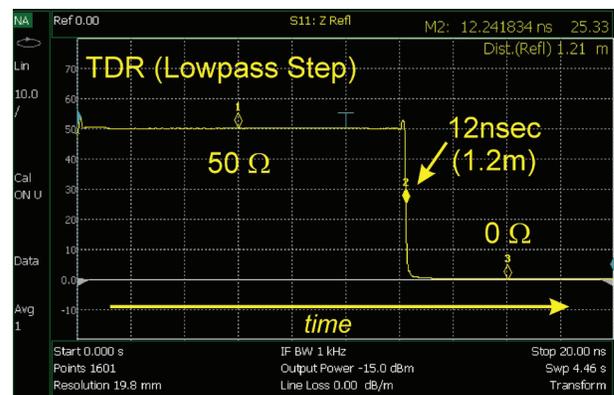
# Introduction

When the purpose of the transmission line is to interconnect devices and distribute signals, the physical geometries of the conductors and dielectric properties of the supporting material will be uniform across the length of the line. When a change occurs in these characteristics, such as point of contact between two different lines or damage along the line, these discontinuities may result in signal reflection and higher line attenuation. Characterizing and troubleshooting transmission lines and systems require measuring the performance in both the frequency domain and time domain. Frequency domain measurements are typically used to verify the RF performance over the specified frequency range of interest. Time domain and distance domain measurements are typically used to physically locate discontinuities along the line. Figure 1a and Figure 1b show typical frequency domain and time domain measurements using Keysight's FieldFox handheld analyzer. Figure 1a shows the measured amplitude and phase response for the reflected (S11) signal over a range of 12 GHz. In this case, the device under test (DUT) is a short length of coaxial cable terminated with a short. Figure 1b shows the measured input impedance of the same cable/short as a function of time. The time domain data is shown in a Time Domain Reflectometry (TDR) format which makes it easy to identify the 50 ohm cable and the location of the short. This application note will provide techniques and examples for measuring a variety of transmission lines including coaxial cable and waveguide.

Measurements made on a transmission line already installed into a system can be uniquely challenging as both ends of the line may be physically separated making it impossible to directly connect both ends of the line to the test equipment. Fortunately, there are measurement techniques available for characterizing long transmission lines when both ends are physically separated in the system. This application note will describe these techniques and show examples for measuring and troubleshooting transmission lines installed in a system.



(a) Frequency Domain (S11)



(b) Time Domain (TDR)

Figure 1. Frequency response and TDR response of a short coaxial cable terminated in a short

## Transmission Line Configurations and Properties

A transmission line can have a variety of physical configurations which support propagation of energy. Figure 2 shows two popular types, namely the coaxial cable and two-wire line. Both of these transmission lines were investigated in the early days of telephone, television and telegraph [1,2] as suitable for signal transmission over large distances. The characteristic impedance,  $Z_0$ , of the lines is related to the cross sectional dimensions of the two conductors and dielectric constant of the insulating material. For example, Figure 2a shows the geometry of a coaxial line with center conductor diameter defined as  $d$  and outer conductor diameter, measured from the inner surface of the outer conductor, as  $D$ . Typically an insulator, having dielectric constant  $\epsilon_r$ , is used to hold the inner conductor symmetrically within the outer conductor. The characteristic impedance is related to the ratio of the diameters of outer to inner conductors. The characteristic impedance of the line is inversely related to the square root of the dielectric constant. Typical impedance values for coaxial cable are 50 ohm and 75 ohm. Figure 2b shows the cross-sectional geometry of a two-wire transmission line with wire diameter defined as  $d$  and wire spacing  $D$ . In this figure, the wires are surrounded by air but typically the wires are coated in an insulator to maintain the correct spacing between the wires for a specific characteristic impedance. The two-wire configuration has been used in 300 ohm twin-lead, 100 ohm twinaxial, and 100 ohm twisted pair. Table 1 shows the equation for calculating the characteristic impedance of the coaxial cable and two-wire.

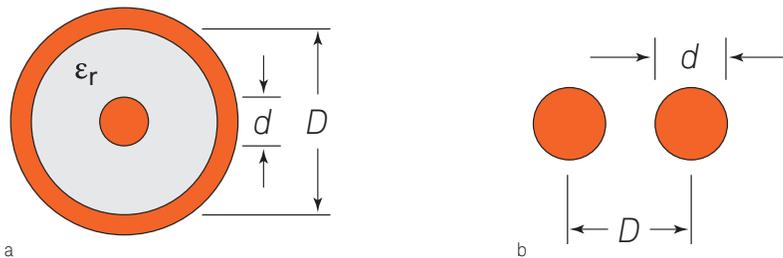


Figure 2. Cross sectional geometry for coaxial cable and two-wire transmission lines

Table 1. Equation for Characteristic Impedance [3]

Coaxial Cable	Two-Wire Line
$Z_0 = \frac{60}{\sqrt{\epsilon_r}} \ln\left(\frac{D}{d}\right)$	$Z_0 = 120 \cosh^{-1}\left(\frac{D}{d}\right) \approx 120 \ln\left(\frac{2D}{d}\right)$

Selection of the characteristic impedance is often related to propagation characteristics of the transmission line. For example, Figure 3 shows the curves for the attenuation, power handling and voltage breakdown on a coaxial cable as a function of ratio  $D/d$ . The values shown in the figure have all been normalized to their respective minimum or maximum values for comparison purposes. These curves are based on the equations [4] listed in Table 2. From Figure 3, coaxial cables have a minimum attenuation when the  $D/d$  ratio is 3.6. Using the equation in Table 1, this leads to an impedance of 76.9 ohms. As the attenuation curve is fairly broad in the range around this value, the broadcast cable industry settled on 75 ohms for the transmission of signals over very long distances. It should be noted that the equation for attenuation in Table 2 is listed for conductor loss, using copper, but also includes a term for dielectric loss. For the relative attenuation curve shown here, it is assumed that the operating frequency is fixed. In this case, the attenuation changes as a function of conductor geometry and associated impedance. When considering that the operating frequency could change, the dielectric losses in coaxial line are linearly proportional to frequency [4] and the conductor losses are proportional to the square root of the frequency so that at the higher frequencies, the dielectric loss of the medium is more important.

When examining the power handling capability of coaxial cable, the peak power capability occurs when the  $D/d$  ratio is 1.65 resulting in an impedance of 30 ohms. The  $D/d$  ratio with the highest voltage breakdown occurs at 2.7 or equivalently 59.6 ohms. As the voltage breakdown curve is also relatively flat in the area of 59.6 ohms, a compromise was reached between power and voltage breakdown to settle on 50 ohms as the impedance for most other systems. In this case, the voltage breakdown is 98% of the peak and the power handling is 86% of the peak. Other research [5] includes examining the performance of braided, polyethylene filled coaxial cable and has shown that cables for general-purpose use are optimized at 50 ohms and for those applications where low attenuation is of prime importance, 75 ohms is ideal.

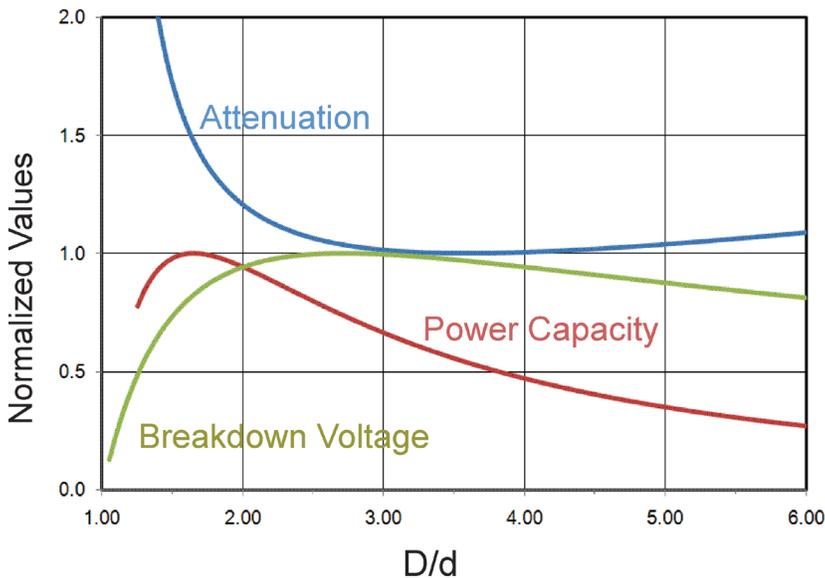


Figure 3. Propagation characteristics of coaxial cable as a function of  $D/d$

Table 2. Table of Propagation Characteristics for Coaxial Cable

Attenuation (dB/cm)	$\alpha = 2.98 \times 10^{-9} \sqrt{f} \frac{1}{D} \left(1 + \frac{D}{d}\right) \frac{\sqrt{\epsilon_r}}{\ln\left(\frac{D}{d}\right)} + \alpha_D; f \text{ in Hz}$ <p>(<math>\alpha_D = \text{dielectric loss}</math>)</p>
Power Capacity (watts)	$P = \frac{(E_m)^2 D^2}{60} \frac{\ln\left(\frac{D}{d}\right)}{\left(\frac{D}{d}\right)^2}; E_m \text{ in volts/cm}$
Breakdown Voltage (volts)	$V = E_m D \frac{\ln\left(\frac{D}{d}\right)}{\left(\frac{D}{d}\right)}$
TE <sub>11</sub> Mode (GHz) [6]	$f_{c,TE11} \approx \frac{7.514}{(D + d)\sqrt{\epsilon_r}}; D, d \text{ inches}$
Propagation Velocity (m/s)	$v = \frac{c}{\sqrt{\epsilon_r}}; c \text{ is speed of light}$
Velocity Factor (VF) (%)	$VF = \frac{1}{\sqrt{\epsilon_r}} * 100\%$

From the equations listed in Table 2, at fixed characteristic impedance, it is noted that the attenuation is inversely proportional to the outer diameter of the coaxial cable. For example, a 50 ohm coaxial cable filled with a dielectric constant of 2.3 would require a  $D/d$  ratio equal to 3.54. Not including dielectric loss, the line attenuation would be  $0.05/D$  dB per meter at 1 GHz. Therefore, a larger diameter cable will result in the lowest attenuation per unit length. Also, for the same fixed line impedance, the power handling capacity and breakdown voltage are directly proportional to this outer diameter. Once again, the largest diameter will produce a cable with the highest performance. While it appears that large diameter cables are always preferred, there is a limit when the cable diameter becomes too large and an undesired mode of propagation can result. Table 2 includes the equation [6] for the frequency when the TE<sub>11</sub> waveguide mode can be produced within the cable. In this case, a large diameter will limit the upper frequency range for the cable. For example, by increasing the outer diameter of a cable by a factor of two, the breakdown capability will be doubled while the maximum operating frequency range will be cut in half. The relationship to operating frequency and waveguide modes will be discussed later in this application note.

Table 2 also lists propagation velocity and velocity factor for a coaxial cable. For a coaxial cable, the electric field is completely contained between the two conductors in a uniform dielectric. In this case, the line is considered “non-dispersive” as the velocity is not a function of frequency and only depends on the dielectric constant of the insulator. In comparison, unshielded two-wire transmission lines will have a part of the electric field propagating in air and part in the surrounding wire insulation. In this case, the line will be dispersive as the propagation velocity will be a function of frequency which may limit the high frequency range of operation.

Other key parameters when selecting transmission lines are shielding requirements for cable-to-cable crosstalk and immunity to external interference. Coaxial lines abandon electrical balance and depend entirely on metallic shielding resulting in a higher cost when compared to balanced two-wire lines. Two-wire transmission lines often operate in a balanced feed where the two wires are of the same gauge and material. The crosstalk between two adjacent coaxial cables falls off very rapidly as the frequency is increased [1]. This effect is quite different for unshielded two-wire systems that rely upon balance to limit the coupling. For both types of lines, more shielding is often required to limit external interference than is required to limit cross talk [1]. This is one reason that some two-wire systems introduce a shielded conductor surrounding the two-wires. Figure 4a shows the cross section of a two-wire transmission line including an outer shield. This line, often called twinaxial, has controlled impedance with improved propagation velocity over a much larger range of frequencies. Twinaxial is often designed for 100 ohm characteristic impedance.

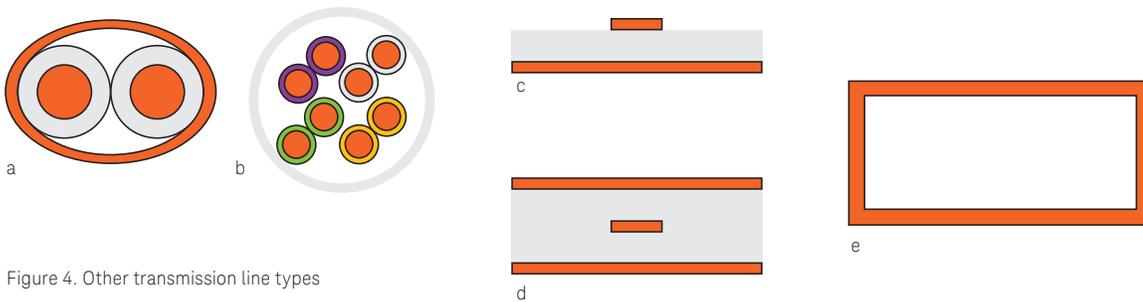


Figure 4. Other transmission line types

Twisted-pair is another two-wire transmission line that is extremely popular in data networking applications including the CAT-5 and CAT-6 cable series. These cables combine 4 pairs of wires bundled within a common outer sheath. Figure 4b shows the cross section of a twisted pair cable. This configuration allows 4 independent signals to travel in parallel along the same cable to improve the total data capacity of the cable. Inexpensive twisted-pair cables are typically unshielded. Higher operating frequency can be achieved when a common shield is placed around the 4 pairs of wire. Twisted-pair lines are designed for 100 ohm impedance.

Microstrip and stripline are transmission lines that are primarily used in printed circuit board and integrated circuit applications. Typical configurations are shown in Figure 4c and 4d respectively. They are ideal for connecting transmission lines with surface mount components including integrated circuits in small packages. Here again the geometry of the conductors and dielectric properties of the supporting material determine the impedance and propagation characteristics for the line. While the coaxial and other two-wire cables are appropriate for use over very long distances, microstrip and stripline are used within small devices including radio and radar components and systems.

The last transmission line type to mention is waveguide. This is the one transmission line structure that does not require two separate conductors. This hollow tube can be constructed using a rectangular cross section, as shown in Figure 4e, or using circular or elliptical cross sections. Electromagnetic field theory is required to understand how the signal travels along this rigid transmission line but key features for using waveguide are the relatively low insertion loss and high power capability. Several examples of waveguide measurements will be provided in this application note.

One key point to mention regarding operating frequency range is that two-conductor transmission lines operate down to DC while waveguide operates only over a narrow range of frequencies. These limitations will be discussed later in this application note.

## Transmission Line Phase and Velocity Factor

In the context of transmission line propagation, what is the signal frequency, and associated wavelength, that makes a cable or other two-wire system a transmission line? There are numerous textbook definitions for a transmission line which include relationships between electromagnetic fields and voltage/current changes along the line. But considering the nature of a signal propagating along a transmission line, the answer is relatively simple: when its physical length is comparable to the wavelength for the signal of interest. At one extreme, when a cable or two-wire line is operating with a signal at DC or very low frequency, the voltage and current values are the same at all points along the line and therefore would not be considered a transmission line. As the operating frequency is increased, the voltage and current values become functions of position as the cable length becomes proportional to the wavelength.

As an example, for a coaxial cable with a length of 1.2 meters and VF=66%, Table 3 compares the number of wavelengths and associated phase length when the input signal is 1 kHz and again when the input is 1 GHz. The wavelength within a transmission line is calculated using the following equation.

$$\lambda \frac{v}{f} = \frac{c}{f} (VF)$$

Where  $v$  is the velocity of the signal in the cable,  $c$  is the speed of light,  $VF$  is the cable velocity factor (=0.66), and  $f$  is the frequency. When the input signal is 1 kHz, the associated wavelength is 198 km which makes the relative length of this coaxial cable only 0.000006 wavelengths long. Using the relationship that one wavelength is equal to 360 degrees, this cable's electrical length is only 0.002 degrees at 1kHz. In this case, the voltage and current are essentially independent of location along the cable. When the frequency is increased to 1 GHz, the signal wavelength is 0.198 meters and the cable is now approximately 6 wavelengths long. Six wavelengths at 1 GHz results in an electrical length of 2,200 degrees. In this case, the cable is considered a transmission line as the cable length is larger than the wavelength of the signal. For the 1 GHz case, the voltage and current, represented as complex signals having magnitude and phase, are now functions of their location along the coaxial cable.

Table 3. Calculated electrical length for 1.2 meter coaxial cable

Frequency	# of wavelengths	Phase Length (degrees)
1 kHz	0.000006	0.002
1 GHz	6.06	2,200

It is important to note that the calculation of wavelength requires the value for the velocity factor of the cable. In free space, a signal will travel at the speed of light, and when a two-wire transmission line is filled with air, it has the same velocity factor of 1 or 100% of the speed of light. When a transmission line is filled with a dielectric, the velocity is reduced by the specified velocity factor. For the cable shown in Table 3, the velocity factory is 0.66 or 66% of the speed of light and the associated wavelength is also reduced by this value.

Knowing that the cable phase is approximately 0 degrees at very low frequency and 2,200 degrees at 1 GHz, the behavior of the transmission phase as a function of frequency can be measured using the vector network analyzer (VNA) mode on a Keysight FieldFox handheld analyzer. Figure 5a shows the measured S11 phase response for the 1.2 meter 50 ohm coaxial cable over a range of 30 kHz to 1 GHz. Using the marker function on FieldFox, the measured phase at 1 GHz is 2,201 degrees which matches the calculated value except for the minus sign. As the VNA phase is always a number relative to the input, the negative sign corresponds to the fact that the cable represents a delay, or lag, in the signal when measured at the cable output.

As shown in Figure 5a, the transmission phase for this cable is a linear function over frequency. A linear phase slope represents a fixed time delay across the measured frequency range. A VNA, such as FieldFox, calculates delay from the slope of the measured phase using the following equation.

$$\text{Delay (secs.)} = -\frac{1}{360^\circ} \left( \frac{\Delta\theta}{\Delta f} \right)$$

Where  $\Delta f$  represents a user-selected "smoothing aperture". Apertures can be adjusted from 0.05% of the measured frequency range to 25% of the range. Increasing the aperture will reduce noise effects from the slope calculation.

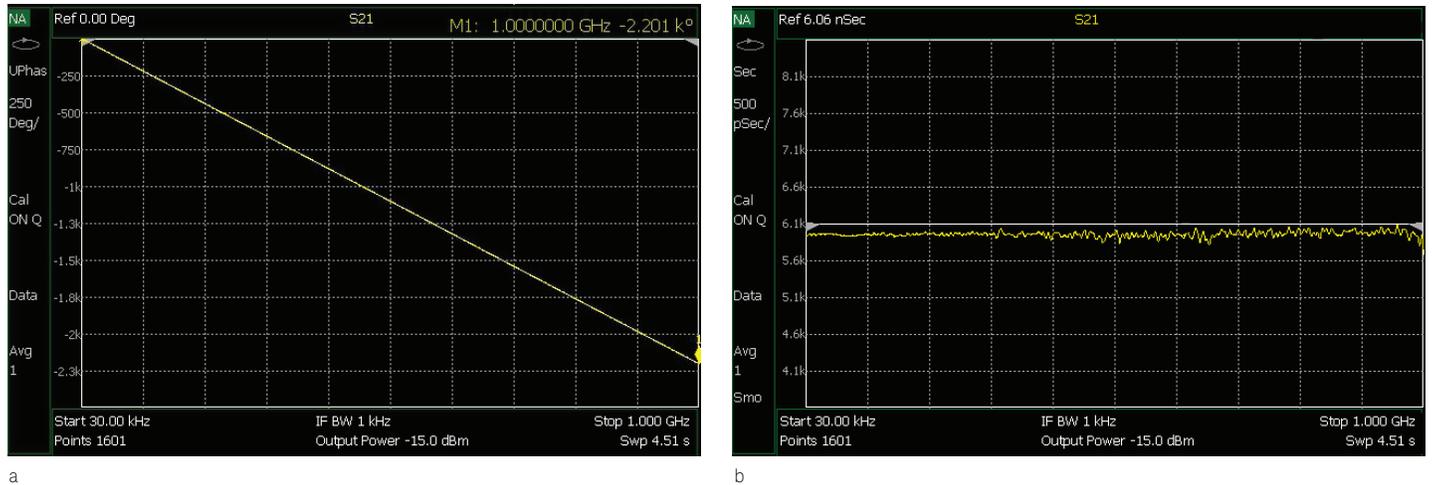


Figure 5. Measured (a) transmission phase and (b) delay as a function of frequency for a 1.2 meter coaxial cable

Figure 5b shows the transmission delay for this cable as a function of frequency. As shown in Figure 5b, the delay is a flat response with a value of 6.1 nanoseconds. If the phase slope is not linear, the calculated transmission delay will be a non-linear function of frequency resulting in distortion of the applied signal as it passes through the transmission line. When the transmission delay is not flat, the line is considered “dispersive”. Examples of dispersive medium are waveguide and microstrip transmission lines. Microstrip lines are dispersive as a portion of the electric field propagates in air and a portion propagates in the supporting dielectric. At higher frequencies, more of the field propagates in the dielectric resulting in a change in the transmission delay as the frequency is increased [7]. Waveguides will be discussed later in this application note.

It should be noted that the cable length can be calculated from the delay measurement if the cable’s velocity factor is known. In Figure 5b, the delay was measured at 6.1 nanoseconds. Knowing that the VF is 0.66 (66%), the cable length is calculated as 1.2 meters using the following equation.

$$Length = (delay) (v) = (delay) (VF) (c)$$

So why is it necessary to discuss the phase and/or delay through a transmission line when most of the test specifications are related to the performance of magnitude quantities such as return loss and insertion loss? While direct phase measurements are typically not important for most applications, unless working with phase matched components such as a feed network in a phased array antenna, the phase of a transmission line directly affects all magnitude quantities as reflections in the system add and subtract at different frequencies.

## Frequency Domain Measurement of Return Loss

Figure 6 shows the measured return loss as a function of frequency for a 50 ohm coaxial cable terminated in a 50 ohm load. The return loss is measured as the “S11” S-parameter using the VNA mode on FieldFox. The frequency range for this measurement covers 30 kHz to 1 GHz. From the figure, there is an observable set of maximums and minimums that are relatively equally spaced in frequency. This response is typical for the interaction of two reflections positioned at each end of a transmission line.

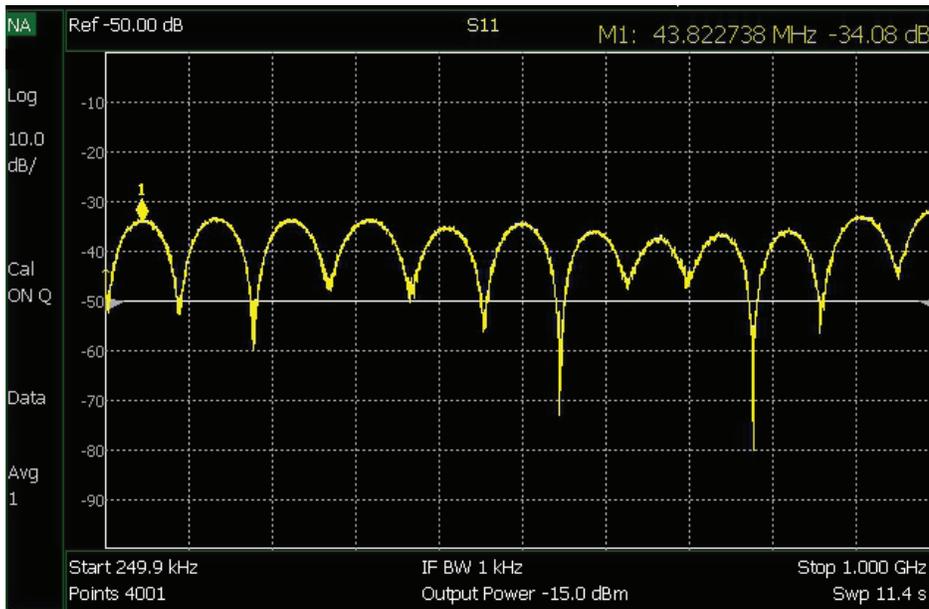


Figure 6. Measured S11 for a cable terminated in a load

Assuming that there are no discontinuities along the length of a cable and the termination is close to an ideal 50 ohm load, the measured reflection at the input is a combination of reflections from the connector and connector-to-cable interface from both ends of the cable. As the input connector is close to the VNA test port, it introduces little phase shift to this reflected signal as shown in Figure 7 as vector “1” with an angle equal to zero. The portion of the signal that passes through the cable will include a phase shift of “phi” degrees due to the cable length. Using the 1.2 meter cable from the previous discussion, this phase shift would be 2,200 degrees at 1 GHz. When the transmitted signal hits the output connector, a portion is reflected back to the VNA source. If the two connectors are of the same type, it can be assumed that the magnitude of the reflection from the output connector will be the same as the reflection from the input connector. After the reflected signal from the output passes back through the cable, the total phase shift will be two “phi” or 4,400 degrees at 1 GHz. These two reflected signals, each having an equal amplitude but a relative phase difference of two “phi”, will add into a single vector shown in Figure 7 as S11. As the frequency is swept over a wide range, the second reflection will sweep around a circle relative to the input reflection. It should be noted, as before, that the phase “phi” is actually a negative value as transmission through the cable represents a lag or corresponding negative phase shift.

If the relative phase difference between the two reflections is a multiple of 180 degrees, the reflections will subtract and the measured S11 will be a null in the frequency response as shown in Figure 7. If the relative phase difference is a multiple of 360 degrees, the reflections will add to a peak at that particular frequency. The measured S11, shown in Figure 6, displays this effect of vector addition and subtraction from these two reflections. If the reflection magnitude of the input and output connectors are identical, the peaks in the measured S11 response will be 6 dB higher than the reflection from a single connector.

It should be noted that the additional simplified vector example shown in Figure 7 does not include re-reflections between the two connectors. As the individual connectors have a relatively low return loss, the amplitude of the re-reflections is too small to be observed on the measured response. If the connectors had a poor return loss, then the re-reflections would introduce additional ripple in the S11 response as a function of frequency. Identifying the location of discontinuities and faults is generally not possible using only the frequency response data. By measuring the same system in the time domain, it is possible to troubleshoot the location of faults. The next section will discuss an example of time domain measurements using FieldFox on the 1.2 meter cable.

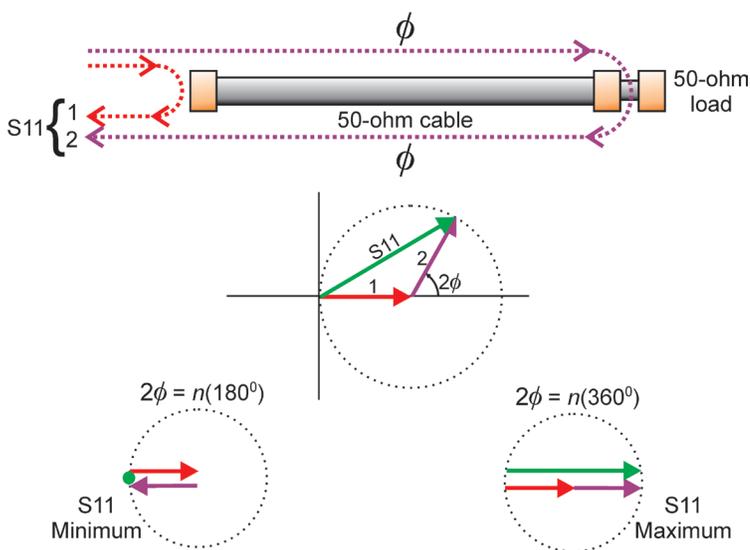


Figure 7. Interaction of reflected signals along a transmission line

## Time Domain Measurement of Return Loss

Continuing with the previous example, if one of the connectors was faulty, or there was damage to a portion of the cable, the measured frequency response of S11 may not provide any guidance to where the fault occurred. In this case, converting the frequency response into a time domain response can provide additional information regarding the physical location of the fault. Figure 8 shows the time domain response of the 1.2 meter cable terminated in a short. This measurement was recorded using the VNA mode in FieldFox and configured to display the Time Domain (TDR) step response. For this response, the analyzer was formatted to display the impedance ( $Z$ ) as a function of time. A basic TDR Step response is the result of exciting a transmission line with a stepped waveform and measuring the input to the DUT as the step travels down the

line and reflects back from any discontinuities. In this example, the step is referenced to time equal to 0 ( $t=0$ ) and the analyzer calculates the impedance of the transmission line as a function of time. Initially the impedance is approximately 50 ohm which is the impedance of the coaxial cable. At a later time, the step arrives at the short. As a short circuit implies a zero voltage across its terminals, the reflected step must be the negative of the incoming step. When the reflected step arrives back at the input it completely cancels the forward signal. The measurement in the time domain is an impedance reported as 0 ohm occurring at the time it takes for a signal to propagate twice through the cable. In the measurement shown in Figure 8, the two-way transit time is reported as 12 nanoseconds measured at the point of the transition in the step. Knowing the velocity factor of the cable and dividing the total transit time in half, the measured time to the short can be converted to the physical length of the cable which is 1.2 meters as expected.

The TDR Step response not only provides the type impedance at a discontinuity but also reports the location of the discontinuity relative to the input of the transmission line. This technique is very useful when troubleshooting transmission lines and will be discussed in more detail later in this application note.

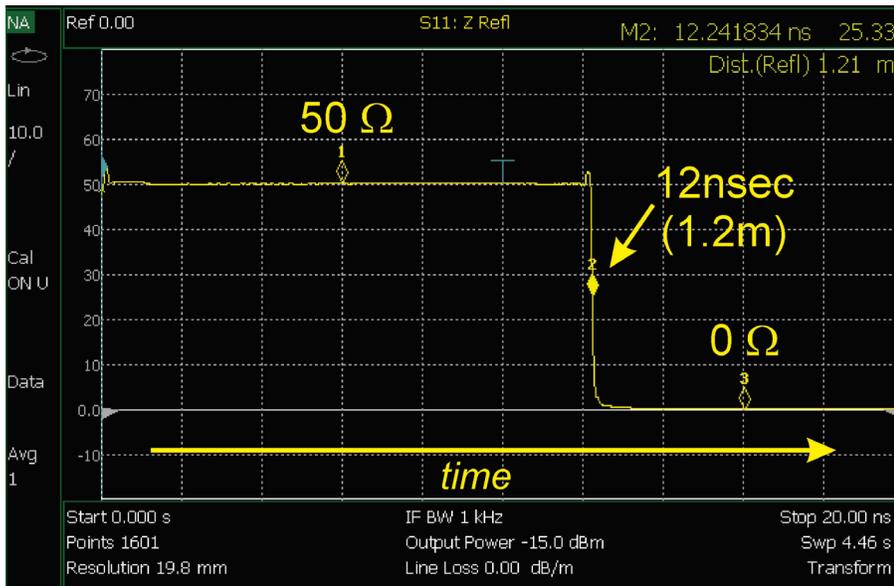


Figure 8. TDR response of a 50 ohm coaxial cable terminated in a short

## Frequency Range Limitations for Transmission Lines

While any transmission line can be measured over the frequency range of the available equipment, the specified frequency range is typically limited by the geometry, materials and construction of the line. Table 4 lists the operating frequency ranges for some of the most popular types of transmission lines and coaxial connectors.

The first category listed is twisted-pair datacomm cables which include CAT-3, CAT-5e and CAT-6a. These cables are specified to operate up to 16 MHz, 100 MHz and 250 MHz respectively. CAT-5e cables, with the “e” representing enhanced performance, have improved crosstalk between the four pairs of wires. CAT-6a, with the “a” representing augmented, have further improved crosstalk and noise performance achieved through

the addition of an outer shield. Next in this chart is twinaxial which is a two-wire cable with shielding. For example, twinaxial DXN-2600 has high frequency performance to 1 GHz. When discussing coaxial cable assemblies, it is important to distinguish the individual performance of the connector and the cable. For example, the larger 7/16 connector has an upper range of 7.5 GHz while the smaller diameter SMA will operate up to 18 GHz. When examining the performance of raw coaxial cable, the diameter of the cable places a limit on the upper frequency range. For example, the large diameter LMR-1700, is specified to 2.5 GHz, while the smaller diameter Accuphase TTL26 has an upper range at 26.5 GHz. When manufacturing a cable assembly, the frequency limitation for the complete assembly is that of the component with the lowest individual performance. For example, placing a 7/16 connector on a LMR-1700 cable restricts the upper frequency range to that of the cable.

It should be noted that all of the two conductor configurations will operate down to DC. The one type of transmission line which includes a low frequency limitation is waveguide. This single conductor “hollow tube” only operates over a narrow range of frequencies strictly determined by the physical dimensions of the guide. Rectangular waveguides are specified with a WR designation followed by numeric code. The WR stands for Waveguide Rectangular and the number corresponds to the larger dimension in mils divided by 10. For example, WR-90 waveguide has a cross sectional dimension of 400 mils (0.4 inches) by 900 mils (0.9 inches), therefore this waveguide is numbered using 90 (900mils/10). WR-90 has a specified operating range from 8.2 GHz to 12.4 GHz.

Table 4. Frequency Ranges for Cables, Connectors and Waveguide

Category	Type	Lower Frequency Limit (MHz)	Upper Frequency Limit (MHz)
Twisted Pair	CAT-3	DC	16
Twisted Pair	CAT-5e	DC	100
Twisted Pair	CAT-6a	DC	250
Twinaxial	DXN-2600	DC	1000
Connector-Coax	7/16	DC	7500
Connector-Coax	Type-N	DC	12000
Connector-Coax	SMA	DC	18000
Coaxial Cable	LMR-1700	DC	2500
Coaxial Cable	RG-214	DC	1100
Coaxial Cable	TTL26	DC	26500
Waveguide	WR-187	3950	5850
Waveguide	WR-137	5850	8200
Waveguide	WR-90	8200	12400
Waveguide	WR-62	12400	18000
Waveguide	WR-42	18000	26500

So what causes a transmission line to have a frequency limitation? Limitations are often associated with the physical geometry of the line that may result in propagation of undesired modes. Earlier in this note, it was briefly mentioned that the cross section of the outer conductor places a limit on the operating frequency for a transmission line. Table 2 included an equation for the frequency when the undesired  $TE_{11}$  waveguide mode would begin to propagate within a coaxial cable. The next two sections of this application note will discuss the conditions when higher order modes begin to propagate on a transmission line and their affect on the measured insertion loss.

## High Frequency Modes in Coaxial Cable

To understand the effects of high frequency modes propagating in a coaxial cable, a 1 meter length of RG-214 was measured beyond the specified frequency range for this cable. The cable ends were terminated in SMA connectors. The specifications for RG-214 cable include an upper frequency limit at 11 GHz. As the SMA connectors are specified to operate up to 18 GHz, this cable assembly is limited to the frequency range of the cable, namely 11 GHz. It should be noted that high frequency modes can be created in both the cables and the connectors.

The effects of higher order modes are shown in Figure 9. In this example, FieldFox was configured to display the RG-214 cable loss ( $S_{21}$ ) over the frequency range of 30 kHz to 18 GHz. As shown in the figure, the cable loss is well behaved to 13 GHz. At 13.5 GHz, there is a sudden drop in the measured  $S_{21}$  due to the first higher order mode propagating on the line. Table 2 above includes an equation to estimate the frequency at which a  $TE_{11}$  waveguide mode can propagate inside a coaxial conductor. For the geometry of RG-214, the calculated  $TE_{11}$  frequency is estimated at 13.1 GHz. Cable manufacturers and industry committees often include a safety margin as a percentage of this calculated  $TE_{11}$  cutoff frequency, and in the case of RG-214, the upper limit is specified at 11 GHz.

Without going into too much detail about electromagnetic theory, a coaxial line can be examined as a two-conductor system or as a single-conductor waveguide. As a two-conductor system, the signal propagates along the transmission line with the electric field starting at the inner conductor and extending radially outward to the outer conductor, as shown in Figure 10a. This is the desired mode of propagation and called the TEM mode for “transverse electromagnetic”. The TEM mode is the dominant mode in all two-conductor transmission lines. As the operating frequency increases, a single-conductor waveguide mode can also be launched on the line as the signal wavelength approaches the dimension of the outer conductor. As shown in Figure 10b, the  $TE_{11}$  circular waveguide mode propagates as the electric field lines start and end on the outer conductor. The  $TE_{11}$  frequency, listed in Table 2, is also called the cutoff frequency for this circular waveguide mode. Below the cutoff frequency, a waveguide mode will not propagate in the line.

Returning to Figure 9, the drop in S21 at 13.5 GHz is therefore the result of energy coupling into the TE<sub>11</sub> waveguide mode within the RG-214 cable. This energy will remain trapped inside the coaxial cable as only the dominant TEM can effectively leave through the SMA connector which is dimensioned not to allow propagation of the TE<sub>11</sub> mode at that frequency. Also, the TE<sub>11</sub> will propagate down the cable with different velocity relative to the TEM mode. Having signal energy shared between different propagating modes will create distortion in the signals should the multiple modes re-combine at the output. It may be worth noting that there are numerous other waveguide modes that can propagate on a coaxial line but the TE<sub>11</sub> mode has the lowest frequency for propagation and thus limits the frequency range of the coaxial line.

With the information presented above and in an earlier section of this application note, optimizing the diameter of the outer conductor of a coaxial line creates an engineering tradeoff as smaller diameters are required for high frequency operation while, using equations listed in Table 2, larger diameters are required for lower insertion loss, higher power capability and higher breakdown voltage.

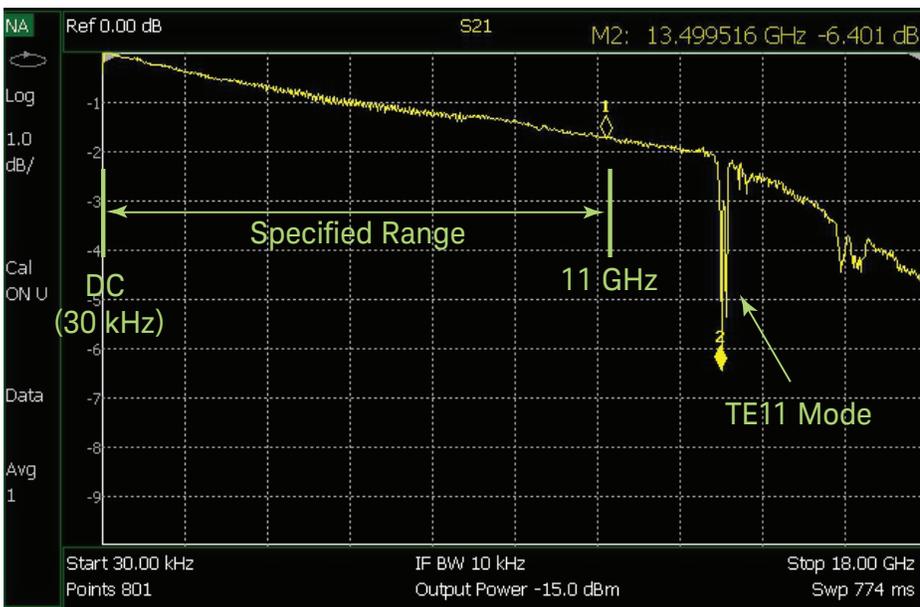


Figure 9. Measured insertion loss (S21) for RG-214 coaxial cable

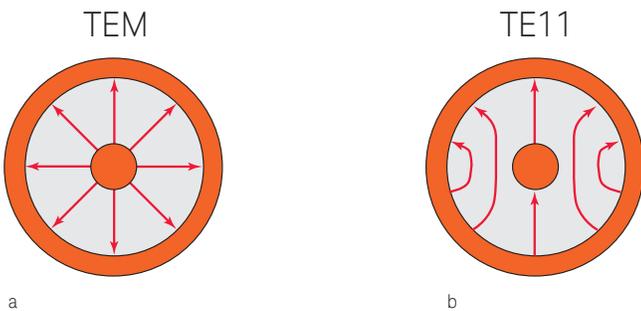


Figure 10. Propagating modes in coaxial transmission line

## High Frequency Modes in Waveguide

Waveguide is another type of transmission line with a frequency range specified by the propagation of higher order modes. Figure 11 shows an S21 measurement for a short section of X-Band WR-90 rectangular waveguide. In this example, FieldFox was configured to display the insertion loss above and below the specified frequency range for this waveguide type. The measurement was taken over the frequency range of 30 kHz to 18 GHz while the WR-90 waveguide is only specified from 8.2 GHz to 12.4 GHz. At the frequencies above the specified range, like coax, higher order modes create a condition where the transmission response is not well behaved. Unlike coax, there is a high level of attenuation at the low-end of the frequency range occurring below the cutoff frequency for the waveguide.

As waveguide is a single-conductor line, the dominant TEM mode found in coax will not propagate in waveguide. In this case, the TE<sub>10</sub> mode is the dominant mode in rectangular waveguide and the electric field distribution is shown in Figure 12. The TE<sub>10</sub> mode is “cutoff” from propagating until the signal wavelength is approximately one-half the waveguide’s largest dimension, labeled in the figure as “a”. Here is the equation for calculating the cutoff frequency for TE<sub>m0</sub> modes in rectangular waveguide.

$$f_{c,m0} = \frac{m(c)}{2a\sqrt{\epsilon_r}}$$

Where **c** is the speed of light and **m=1** for the dominant TE<sub>10</sub> mode. For X-band WR-90 with its largest dimension standardized to 0.9 inches, the TE<sub>10</sub> mode is considered to be propagating above 6.56 GHz.

The S21 measurement in Figure 11 shows a rapid improvement in insertion loss as the frequency approaches up to the specified cutoff frequency of 6.56 GHz. Above the cutoff frequency, the insertion loss further improves and becomes very low above 7 GHz.

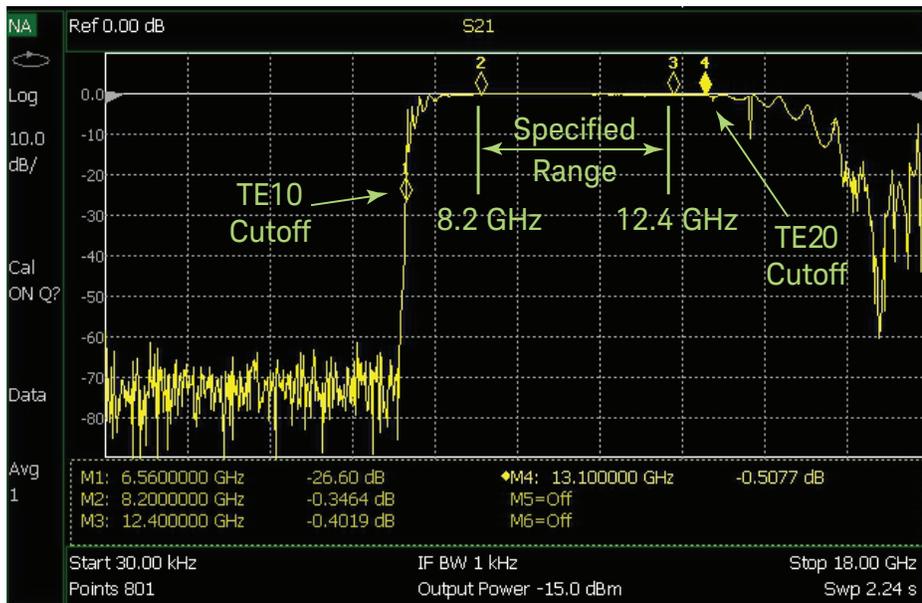


Figure 11. Measured insertion loss (S21) for WR-90 waveguide

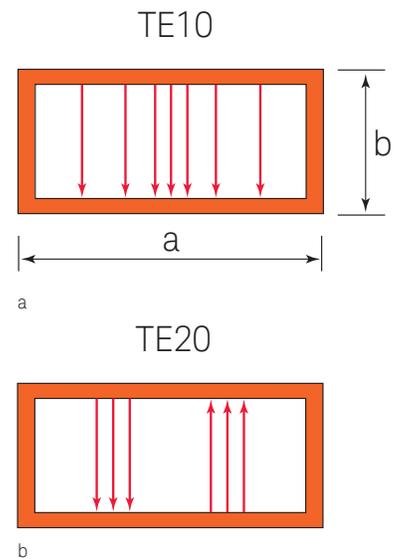


Figure 12. Propagating modes in waveguide transmission line

The specification for WR-90 starts at 8.2 GHz which places the operating range well above the cutoff frequency for this waveguide. At the high end of the range, the insertion loss is not well behaved above the calculated cutoff frequency of 13.1 GHz. Fortunately, the specified range is below 12.4 GHz and will not be effected by the propagation of higher order modes.

There are two other single-conductor waveguide configurations that are also found in industry, namely the circular waveguide and the elliptical waveguide. Circular waveguide is ideal for rotary joints and is often found in radar systems. Elliptical waveguide, such as Heliac, is often used in radar and wireless backhaul feeds requiring bends. Each configuration has numerous modes of propagation such as the dominant TE<sub>11</sub> mode for air-filled circular waveguide with a cutoff frequency calculated using the following equation [8].

$$f_{c,TE_{11}}(\text{GHz}) = \frac{3.46}{a\sqrt{\epsilon_r}}; a \text{ inches}$$

Where **a** is the radius for the circular waveguide. This is a similar equation to the equation previously shown in Table 2 for the TE<sub>11</sub> mode found in coaxial cable where the cutoff frequency for hollow circular waveguide is approximately half the frequency found for coaxial cable. The main difference is the TE<sub>11</sub> mode is the desired mode of propagation for circular waveguide but is an undesired mode for coaxial cable. Calculations of the cutoff frequencies for elliptical waveguide are more complicated as they are dependent on the major axis, minor axis and eccentricity. Formulas for the propagation characteristics for elliptical waveguide are provided in several references including [9].

Another key point regarding the frequency response of waveguide is related to the measured delay. Waveguide is considered a dispersive medium as its delay is a non-linear function of frequency. This dispersion will create distortion in wideband signals. Figure 13 shows the measured delay of a 2.5 foot section of WR-90 rectangular waveguide. The measurements show that lower frequencies have a longer delay than higher frequencies and the overall response is very non-linear. The reason for the different delay times is the result of how the travel path through the waveguide varies at different frequencies. For example, Figure 14 shows an illustration of a short length of waveguide where the signals travel in a crisscross pattern along the wider dimension of the waveguide. The angle of incidence, shown here as theta, is relatively small at the higher frequencies. At lower frequencies, the angle of incidence increases and eventually reaches 90 degrees when the frequency drops to the specified cutoff frequency. At cutoff, the signal is trapped as it bounces back and forth between the two side walls. The larger angle results in a longer time to travel through the waveguide as it must take a longer path. This effect is shown in Figure 13 as there is an increase in the measured delay at lower frequencies.

When using FieldFox to locate faults in waveguide systems, it is necessary to configure the analyzer to adjust the velocity factor (VF) to be a function of frequency. Here is the formula that properly corrects the waveguide VF as a function of frequency.

$$VF(f) = \sqrt{1 - \left(\frac{f_{cutoff}}{f}\right)^2}$$

For example, at 8.62 GHz the calculated VF is equal to 64.9% and at 11.98 GHz, the VF is equal to 83.6%. On FieldFox, the user is only required to select “waveguide” as the medium and the analyzer will make the appropriate corrections to the measured data. FieldFox also includes a table for selecting the type of waveguide used in the measurements, such as WR-90 in this case. Properly setting the waveguide VF is important when using the time domain response to troubleshoot the physical location of faults within the waveguide transmission system. With the correct VF setting, the measured distance to the discontinuity would be correctly displayed.

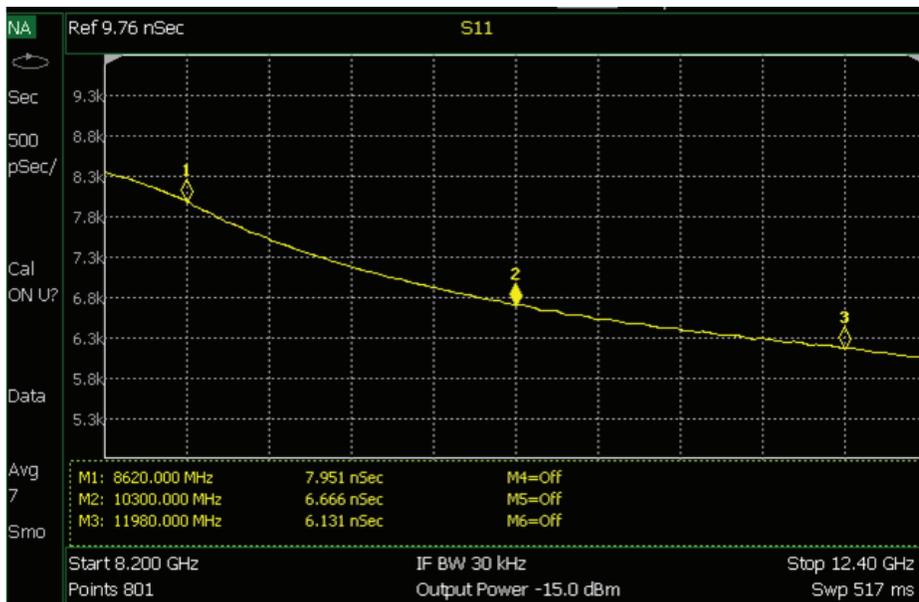


Figure 13. Measured delay of WR-90 waveguide as a function of frequency

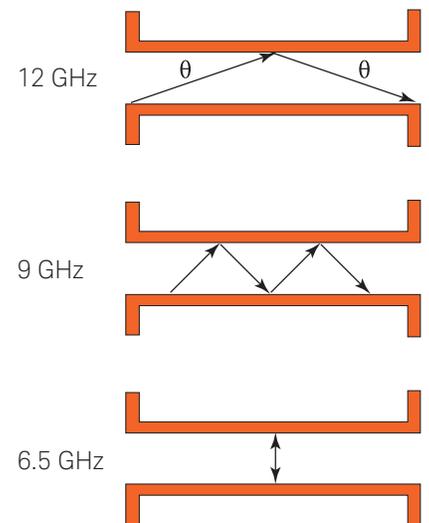


Figure 14. Diagram showing the propagation paths for WR-90 rectangular waveguide for operating frequencies above and below the cutoff frequency

## Troubleshooting Transmission Lines

Complex feed systems for radar, cellular, data communications and cable broadcast, to name a few, all require the verification of the RF performance for their transmission lines as a function of the frequency. If a transmission line is found to be faulty and performing outside the desired specification, then it becomes important to rapidly identify the location and type of failure for the repair. In some types of time domain measurements, including TDR and Low-Pass, it is also possible to identify the potential cause of the failure. Failures include loose or damaged connectors and waveguide flanges, water ingress into the system which is especially prevalent in outdoor installations, broken solder joints, cut cables, installation exceeding the specified minimum bend radius, and damage due to high power arcing.

Time domain measurement techniques include line sweeping, Distance to Fault (DTF), Time Domain Reflectometry (TDR) and Time Domain Transform (TDT). These techniques all result in identifying the physical location where a problem is found to exist along a transmission line. A traditional time domain measurement requires the introduction of an impulse or stepped waveform at the input to the transmission line. This impulse or step travels down the line until it encounters a discontinuity in the cable. Depending on the size of the mismatch created by the discontinuity, a portion of the incident waveform is reflected back to the source. In the ideal case, most of the incident signal passes by the discontinuity and continues to the intended load. Unfortunately, if the reflected signal is relatively large the cable may fail its insertion loss and/or return loss specification. Measuring the time delay for the return signal and dividing this time in half, it is possible to calculate the one-way travel time and associated distance to the damage. As discussed earlier in this application note, the distance calculation requires an accurate value for the velocity factor of the transmission line in order to properly convert the time measurement into a distance.

FieldFox measures a DTF, TDR and other time domain responses by first measuring the frequency response of the transmission line and then transforming this measurement into the time domain. The time domain transform uses a mathematical conversion known as the Inverse Fast Fourier Transform (IFFT) [10]. As swept frequency data is transformed into time domain results, TDR on a network analyzer is sometimes referred to as Frequency Domain Reflectometry (FDR) to differentiate between instruments that measure TDR using only time measurements versus TDR based on frequency-to-time conversion.

As an example, Figure 15 shows a DTF measurement of two short 50 ohm coaxial cables connected together with a coaxial adapter. The shorter cable is connected to port 1 on FieldFox and the second cable is terminated in a 50 ohm load. Markers are placed at the three peaks in the measured DTF response. The first marker, shown on the far left, reports a distance of 0 meters. This marker represents the interface between the calibrated FieldFox and the first coaxial cable. The second marker reports a distance of 4 meters. This marker is located at the adapter between the two cables. It also indicates that the length of the first cable is 4 meters. The third marker is located at the 50 ohm load and is reported at 13.8 meters. This measurement can be used to calculate the length of the second cable which is 9.8 meters (13.8 meters – 4 meters). There is a noticeable drop in the measured amplitude to the right of the 50 ohm load signifying the location of the end of the cable. As this reflection measurement represents two-way signal paths, FieldFox properly adjusts the marker values and x-axis formatting to the appropriate one-way lengths. It should be noted again that the cable's velocity factor (VF) must be entered correctly into FieldFox otherwise the distance measurements will not be correct. It should also be noted when two cable types having different velocity factors are included in the measurement, such as the case when a short jumper cable is connected to a longer system cable, the velocity factor of the longer cable should be entered into FieldFox. Ideally, the short jumper cable should be included as part of the FieldFox user calibration, such as *QuickCal*, and therefore its effects would be removed from the DTF measurement. Additional information regarding FieldFox calibration can be found in the Keysight application note "Techniques for Precise Measurement Calibrations in the Field Using FieldFox Handheld Analyzers" [11].

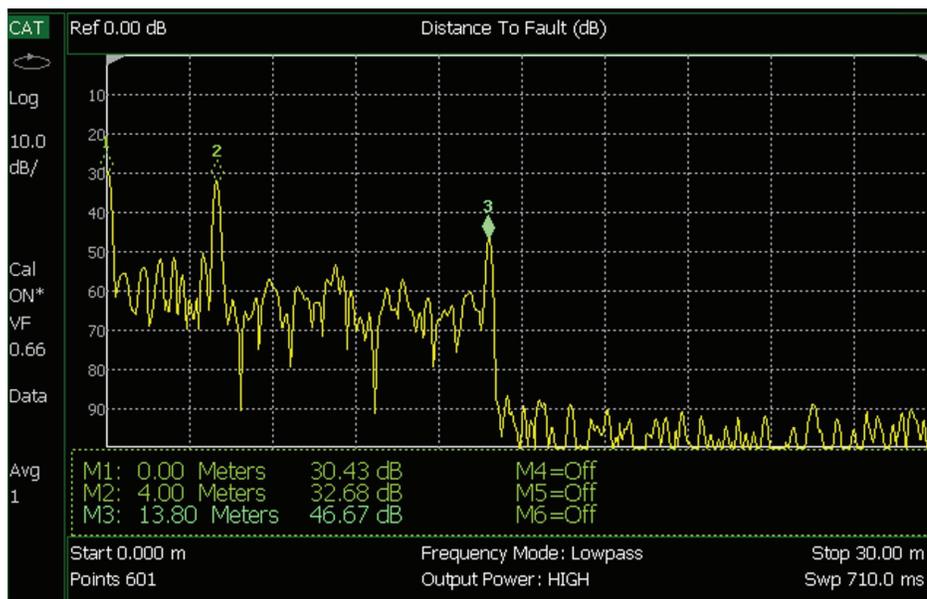


Figure 15. Distance to Fault (DTF) measurement for two connected coaxial cables terminated in 50 ohm load

This simple example shows the importance of DTF measurements when attempting to locate the physical location of a discontinuity along a transmission system. The peaks in this example represent the magnitude of single reflections from a discontinuity but it is also possible to determine the type of reflection using TDR and low-pass modes available on FieldFox handheld analyzers.

## TDR and Time Domain Measurements of Coaxial Cable

Figure 16 shows a photo of a coaxial cable with damage in two areas along the cable. The first problem, labeled “A”, is a bend in the cable that has exceeded the manufacturer’s specification for minimum bend radius. For this cable, the specified bend radius should be 1-inch or larger. For this damaged cable, the bend at location “A” is well below this radius creating an undesired reflection from this part of the cable. The next fault, located at “B”, is a cut through the outer conductor of the cable. At this location, the braided shield has been partially removed exposing the inner dielectric of the coax. Both of these faults can be examined using any one of several time domain techniques available on FieldFox. One technique previously discussed in Figure 15, is the DTF. DTF will only report the location of these faults. Two additional techniques, known as TDR and Impulse response, can be used to characterize the type of fault including discontinuities that are inductive or capacitive. Figures 17a and 17b show measurements for the damaged cable in TDR and Impulse modes respectively. The TDR response, also referred to as stepped response, shows that the cable impedance is generally 50 ohms across most of the time domain response until a discontinuity is encountered. The locations of the discontinuities occur at the input connector, the bend at A, the cut at B and the 50 ohm termination at the end. Of all the discontinuities on this cable, the cut “B” in the outer conductor has the largest mismatch as shown by the magnitude of the largest peak in this time domain response. The cut on the TDR response has a single peak in the positive direction. This indicates an inductive mismatch which is typical for cuts in the outer conductor of a coaxial cable.

Another very useful time domain mode is the Impulse response shown in Figure 17b. Impulse response also provides information about the type of discontinuity by examining the positive and negative going peaks such as those observed at location B. This set of peaks, positive then followed by negative, also reports that the fault is inductive. If the fault was capacitive, the peaks would start in the negative direction and then be followed with a positive peak.

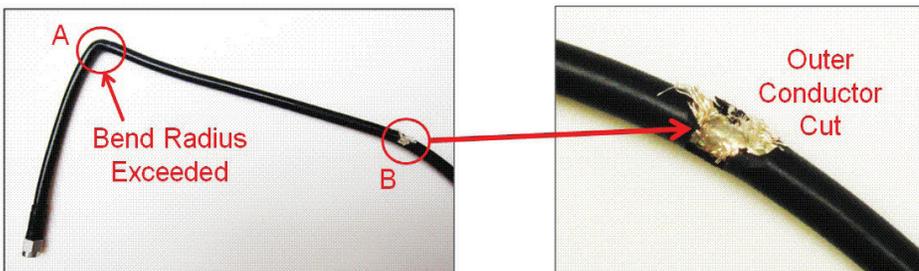


Figure 16. Photo of damaged coaxial cable

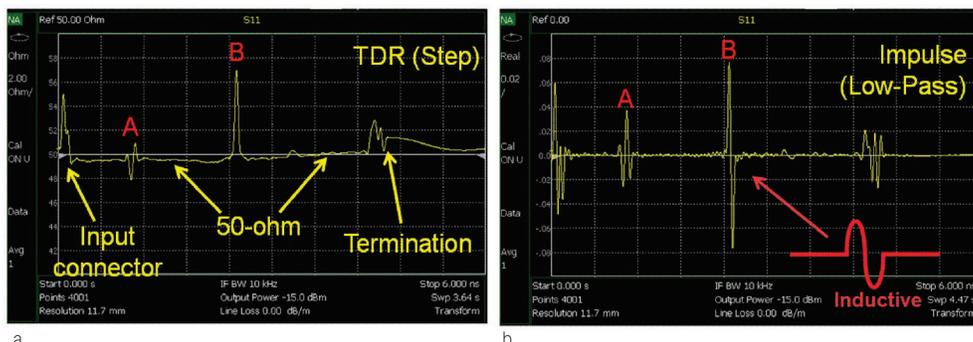


Figure 17. Time domain measurements of damaged coaxial cable

Table 5 summarizes the various types of discontinuities that can be identified with FieldFox using either TDR (Step) or Impulse modes. As discussed, an inductive or capacitive junction can be identified using either mode. As another example, should a transmission line be terminated in a load with a resistance that is larger than the characteristic impedance, the TDR response would show a step in the positive direction. If the load resistance is smaller, the step would move in the negative direction. We observed this effect in Figure 8 when a 50 ohm cable was terminated in a short circuit and the measured step started at the 50 ohm level and then dropped to 0 ohm at the short. The Impulse mode may also be used to identify changes in the impedance as a single positive peak when the load resistance is larger than  $Z_0$  or a single negative peak when the load is less. When measuring transmission lines with similar characteristic impedance, it is typically easier to interpret the display results when in Impulse mode. Many engineers familiar with traditional TDR instrumentation tend to use FieldFox in the TDR mode.

It is important to note that TDR (Step) and Impulse modes are only available for those transmission lines that can operate down to DC, namely two-conductor transmission lines. When using waveguide, the narrowband response restricts the time domain measurements to only “Bandpass” mode of operation. Bandpass mode is ideal for frequency limited DUTs but only provides the location of the fault. Determination of the type of discontinuity, such as inductive, capacitive or resistive is not available in the bandpass mode.

Table 5. Relationship between the type of discontinuity and the displayed waveform using TDR Step and Impulse modes

	TDR (Step)	Impulse (Low-Pass)
$R > Z_0$		
$R < Z_0$		
Inductor		
Capacitor		

## Time Domain Measurements of Waveguide

The application of time domain measurements to waveguide transmission lines is limited to the bandpass mode. As the frequency range of waveguide is limited to a narrow range of frequencies, FieldFox is configured to measure the frequency response over the specified bandwidth of 8.2 GHz to 12.4 GHz for (WR-90) and then the bandpass time domain transform is selected. For example, Figure 18 shows time domain measurements for a system of waveguide components. The transmission system under test starts with a coaxial-to-waveguide adapter connected to FieldFox. Next is connected a 6 inch length of straight rigid waveguide which is then followed by an 18 inch length of flexible waveguide. For the measurements shown in Figure 18a, the flexible waveguide, or “flexguide”, is either terminated in a matched waveguide load or the end flange is left open. Both traces in Figure 18a show a first peak at the coax-to-waveguide adapter

located at time equal to zero. For the measurement with the open-ended flexguide, there is a second large peak corresponding to the location of the open. When the flexguide is terminated using a matched load, the amplitude of the peak is very low in comparison. Markers placed at the peak of each reflection report the electrical distance and the associated physical location to the discontinuity. For example, the location of the open circuit at the end of the flexguide is measured at 675 mm which is the total length through the adapter, straight waveguide and flexguide. As mentioned earlier, FieldFox automatically corrects for the nonlinear velocity factor of the waveguide when the WR-90 type is selected from the “waveguide and cable electrical properties” table.

As a comparison, the transmission system was exposed to the environment and water leaked into the waveguide and flowed into the bottom of the flexguide. For the measurement shown in Figure 18b, the time domain response now displays a large peak that corresponds to the location of the water-filled waveguide. Once again, a marker is used to measure the physical distance to the water which is 384 mm from the input. It is interesting to note that the reflection from the open-ended waveguide is now “masked” by the large reflection from the water in the waveguide. Water is such a lossy medium for RF that any signal that propagates through the water in this system, reflects from the open and returns through the water a second time and is so highly attenuated that it is hardly observable at the input.

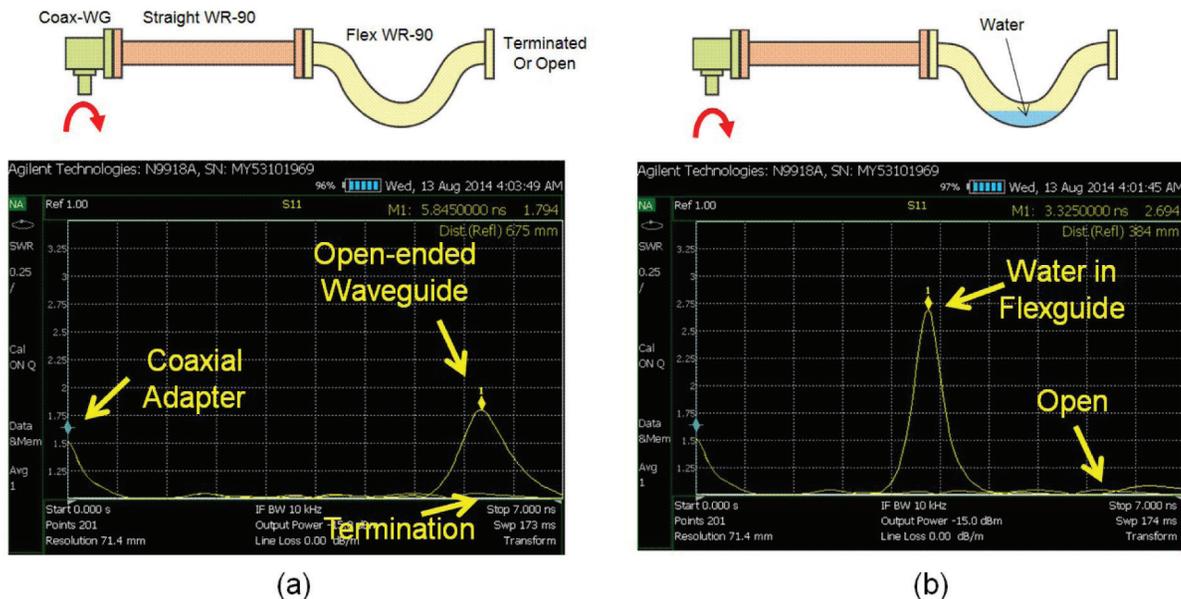


Figure 18. Time domain measurements of a system of waveguide components

## Measurement Techniques for Installed Transmission Lines

Once transmission lines, including coaxial cables and waveguides, are installed into a system, it is often difficult and costly to remove them in order to verify their operation and troubleshoot failures. Also with very long lengths of lines, access to both ends at the same time is typically impossible especially when attempting to connect the transmission line to the test instrumentation including FieldFox. Under these conditions, techniques that allow insertion loss measurements to be made from only one end of the transmission line are preferred.

Figure 19 shows three techniques available on FieldFox that can be used to measure the insertion loss of a cable or transmission line. Depending on the total cable loss and whether both ends of the cable will be accessible will determine which technique is best to use.

The traditional 2-Port method, shown on the left, is the desired configuration when both ends of the cable can be directly connected to FieldFox. This configuration will result in the highest measurement accuracy as FieldFox can apply a full 2-port calibration to remove all measurement errors associated with adapters and jumper cables required for the test. On FieldFox, a full two-port user calibration can be performed using a mechanical or electronic calibration kit. FieldFox also includes the unique “CalReady” and “QuickCal” calibrations which provide high accuracy measurements without the need for a calibration kit [11].

The center configuration in Figure 19 shows a technique for measuring the insertion loss from only one end of the cable. Often when a long cable is installed into a system, it is often difficult to physically connect FieldFox to both ends without introducing an equally long jumper cable into the test setup. Fortunately, the 1-port cable loss technique will eliminate the need to carry an extra-long, high-quality test cable as part of the equipment requirements for on-site testing. This simple 1-port configuration requires a single connection to one end of the cable and leaving the other end either open or terminated in a short. It is preferred at microwave frequencies to use a shorted termination to eliminate fringing fields found in an open-ended cable which could alter the measured results. In this configuration, FieldFox measures the S11 of the cable and calculates the one-way insertion loss from the two-way reflected measurement. This technique is ideal for cables whose insertion loss is less than 30 dB. When the insertion loss is larger than 30 dB, the dynamic range of the test system begins to reduce the accuracy of the measurement.

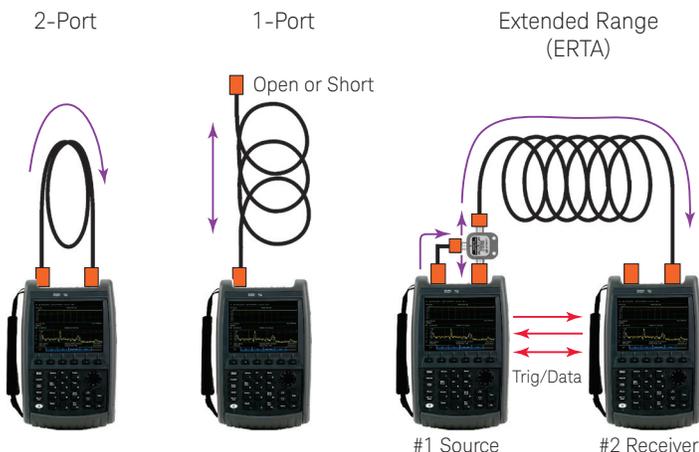


Figure 19. Test configurations for measuring insertion loss

The third configuration, shown on the right in Figure 19, requires the use of two FieldFox analyzers connected at opposite ends of the cable. This technique is referred to as Extended Range Transmission Analysis (ERTA) and is a test mode on FieldFox. With both analyzers configured in ERTA mode, one FieldFox acts as the stepping source and the other acts as a stepping receiver. Both the source and receiver are synchronized using a master/slave configuration and a shared trigger. The high dynamic range receiver found in FieldFox requires no calibration or warm-up time. At the source, a two-resistor power splitter, such as the Keysight 11667B, splits the signal between the local and remote analyzers. The measured insertion loss is displayed on both analyzers as a ratio of the two measurements. The ERTA configuration is ideal for transmission systems with high insertion loss (>30 dB). This configuration is also capable of measuring transmission paths which contain a frequency conversion element such as a frequency downconverter or upconverter.

## Measurement Configurations for Other Transmission Line Types

FieldFox, as most Keysight analyzers, are generally configured as 50 ohm instruments with coaxial test ports. When measuring transmission lines other than 50 ohm cables, it will be necessary to introduce an adapter between the analyzer and the DUT. Figure 20 shows several test configurations for measuring transmission lines other than 50 ohm coaxial cable. FieldFox is configured with either Type-N connectors, for models up to and including 18 GHz, or 3.5 mm connectors for the 26.5 GHz models. When measuring 50 ohm coaxial cable, the cable can often be directly connected to the analyzer's test ports. If the cable under test has different connector types, such as 7/16, TNC or 7 mm, adapters will be required to interface the instrument's test ports with the cable. Ideally, a calibration kit will be required to match the connector type of the cable under test. If the appropriate calibration kit is not available, FieldFox includes the *QuickCal* option to extend the calibration to the end of the test adapters.

When measuring a 75 ohm cable, adapters will be used to convert the 50 ohm test port to the 75 ohm characteristic impedance of the cable under test. In this case the adapters, such as the Keysight N9910X-846, will be connected to FieldFox. If a 75 ohm calibration kit is not available, *QuickCal* may also be used to improve the accuracy of the measurement by adjusting the FieldFox built-in calibration to include the adapters. If measurements using the Smith Chart format are required, then the FieldFox system impedance should be set to 75 ohm for proper readout of the measured impedance.

The measurement of waveguide also requires the use of adapters. In this case, coaxial-to-waveguide adapters are selected to interface the test ports of FieldFox to the flange type of the waveguide under test. Ideally, a waveguide calibration kit should be used for the highest measurement accuracy. Additionally, when measurements are made in the time domain, it is important to set the type of "medium" to "waveguide" so FieldFox can properly adjust the frequency dispersion characteristics of the analyzer.

Finally, the measurement of twisted pair cable, also known as a mixed mode measurement, requires the use of either a RJ-45 to SMA adapter or a balance to unbalanced (BALUN) adapter. For mixed mode measurements, FieldFox requires an option for these types of measurements and Keysight engineers can provide assistance when configuring all of the analyzer's options and adapters listed here.



Figure 20. Test configurations for measuring other types of transmission lines

## Conclusion

This application note has reviewed basic transmission line theory and limitations associated with coaxial and waveguide transmission lines. Troubleshooting techniques using time domain tools available on FieldFox were described. Frequency domain and time domain measurement examples were provided for coaxial cables and waveguide components. Test configurations for measuring different types of transmission lines using FieldFox handheld analyzers were also provided.

## References

- [1] Espenschied, Lloyd ; Strieby, M.E., "Wide Band Transmission Over Coaxial Lines," Transactions of the American Institute of Electrical Engineers, Volume: 53 , Issue: 10, Page(s): 1371 – 1380, 1934.
- [2] Clark, A.B., "Wide Band Transmission Over Balanced Circuits," Transactions of the American Institute of Electrical Engineers, Volume: 54 , Issue: 1, Page(s): 27 – 30, 1934.
- [3] Sams, H.W., *Reference Data for Radio Engineers*, Sixth Edition, International Telephone and Telegraph Corp., 1975.
- [4] Moreno, T., *Microwave Transmission Design Data*, Sperry Gyroscope Company, 1948.
- [5] Blackband, W.T., "The choice of impedance for coaxial radio-frequency cables," Proceedings of the IEE - Part B: Radio and Electronic Engineering, Volume: 102, Issue: 6, 1955 , Page(s): 804 – 814.
- [6] Fuks, R., "SMA Connectors with Extended Frequency Range," Microwave Journal, July 2007.
- [7] Gupta, K.C., et al., *Microstrip Lines and Slotlines*, Artech House, 2nd edition, pg 79-80.
- [8] Pozar, David M., *Microwave Engineering*, 3rd Edition, John Wiley and Sons, Inc., 2005.
- [9] Marcuvitz, N., *Waveguide Handbook*, MIT Radiation Laboratory Series, McGraw-Hill, 1951.
- [10] Brigham, E.O. *The Fast Fourier Transform*, Prentice-Hall, 1974.
- [11] *Techniques for Precise Measurement Calibrations in the Field Using FieldFox Handheld Analyzers*, Keysight Application Note, literature number 5991-0421EN.

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Related literature	Number
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<i>FieldFox Microwave Spectrum Analyzers, Technical Overview</i>	5990-9782EN
<i>FieldFox Microwave Vector Network Analyzers, Technical Overview</i>	5990-9781EN
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<i>FieldFox N9912A RF Analyzer, Technical Overview</i>	5989-8618EN
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