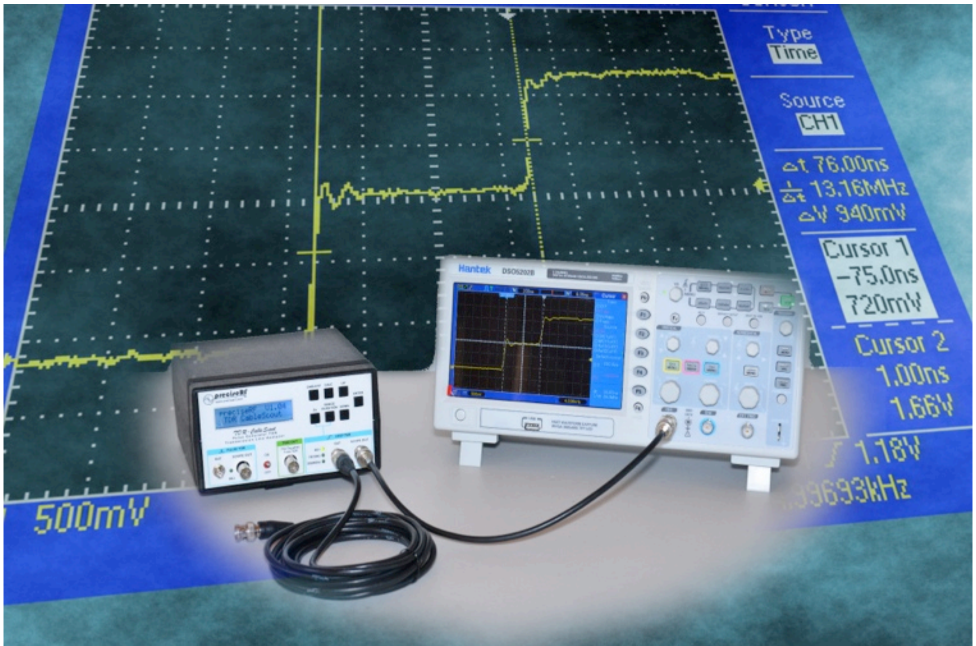

Compilation of Application Notes for Ham Radio Measurements - 2013

1. Transmitter Trapezoid Test
2. Transmitter Two Tone Test
3. Amplifier Linearity test with Spectrum Analyzer
4. Power Measurements with an Oscilloscope
5. Return Loss Bridge Measurements
6. Station Monitor Selection
7. TDR Transmission line Testing

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Precision Ham Radio Measurements

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Application Note #1

Transmitter Trapezoid Test

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How to measure amplifier non-linearity

How can the ham make practical amplifier measurements to check for splatter, IMD products and unwanted harmonic content? While it is helpful to understand the theory and math behind amplifier non-linearity's, all that is required is to make accurate measurements minimizing splatter and distortion is an understanding of the basic concepts.

Basically there are three methods to measure non-linearity and IMD products; Spectrum analyzer, Trapezoid test and the Two Tone test. Each has its advantages and disadvantages.

Trapezoids station monitor testing

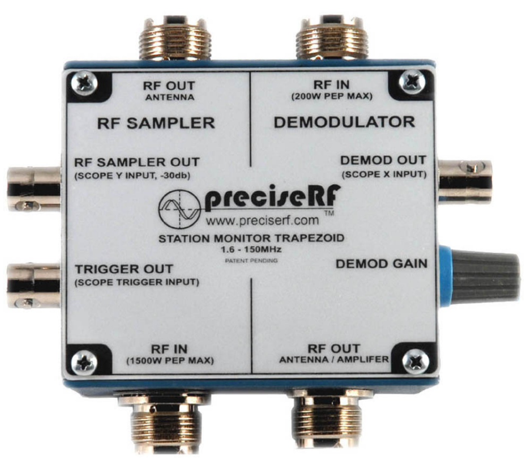

Figure 1 Station monitor

Figure 1. The station monitor provides all the components to make accurate transmitter tests. It includes a demodulator with separate level control for the scope's X input, an RF sampler for the scope's Y input, and a scope trigger output.

This station monitor allows for precise tuning of the entire transmitter chain with transceiver output of up to 100 Watts driving linear RF amplifiers. It features a wide band sampler, a high performance demodulator, a variable base band output, and an oscilloscope trigger output. A Linear RF amplifier generally amplifies an RF signal from 20-100 Watts by 20dB or more to about 500-1,500 Watts. Its performance and modulation can be characterized using a low-cost oscilloscope with a trapezoid display.

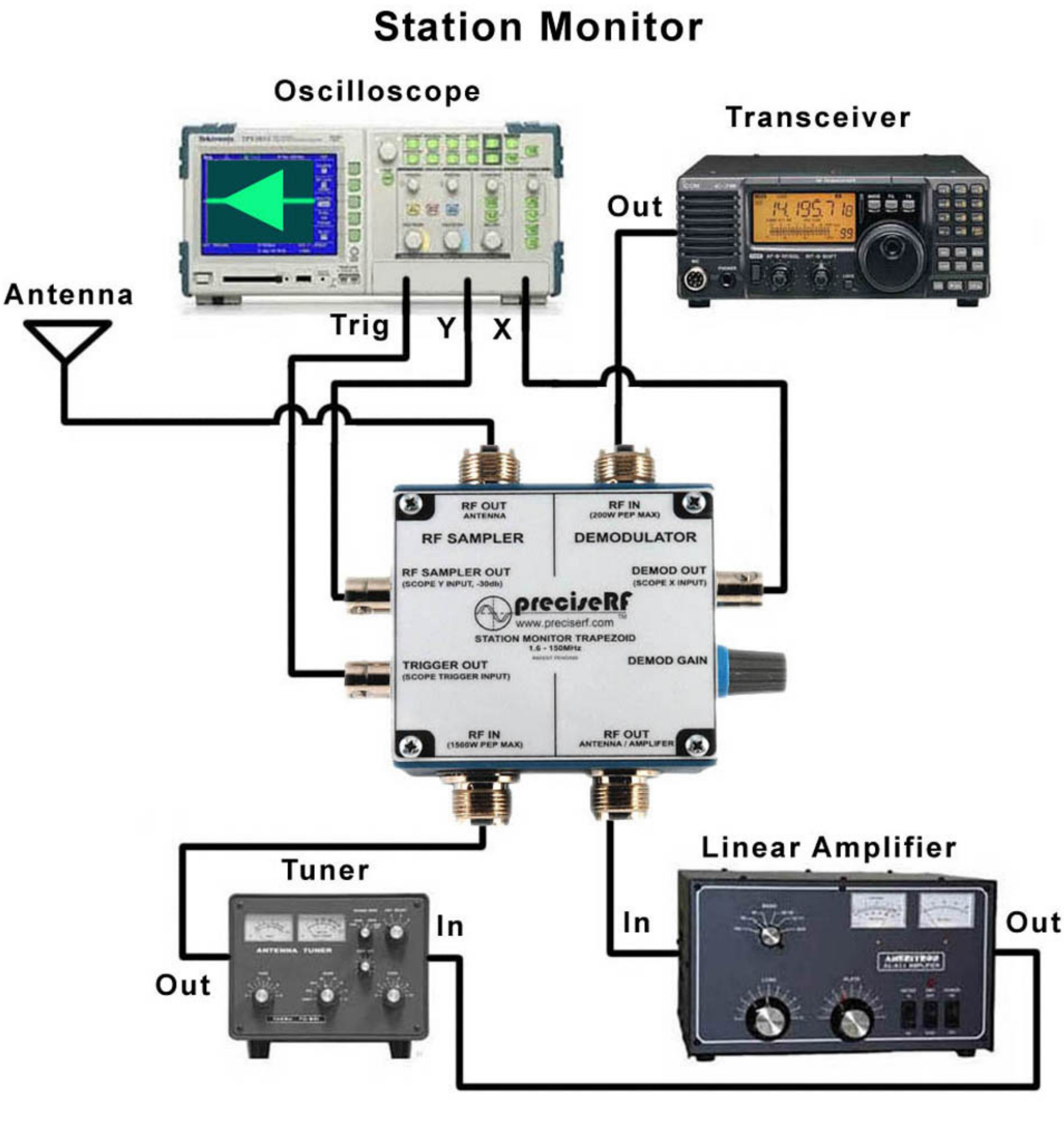

Figure 2 Block diagram station monitor

Figure 3. The RF signal is sampled and demodulated by the station monitor. The outputs are connected to the scope's X and Y inputs and external trigger input for monitoring the signal quality..

Figure 3 Station monitor connection

This is done by sampling the amplifier's output by using an RF sensor. This sensor is connected to the oscilloscope's vertical (Y) input. See the Figure 2. block diagram..

The input of the amplifier is driven by a transceiver which usually outputs less than 100W. Its output drives the input to the amplifier and also a wide band demodulator which extracts the baseband from the modulated carrier. It is this baseband that is connected to the oscilloscope's horizontal (X) input. This display yields a trapezoid pattern. This pattern compares the transceiver's output to the amplifiers output.

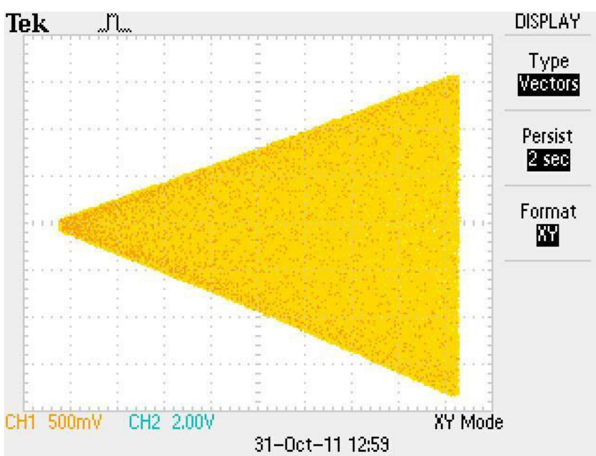

Figure 4. Shown here is the amplitude of the modulated RF increasing in a linear fashion, from minimum at left to maximum at right. When the demodulated horizontal (X) is reversed the trapezoid will be reversed seen here.

Figure 4 Station monitor display

If the amplifier is linear without any distortion and not overdriven, the trapezoid pattern will be a linear undistorted triangular waveform. To rely on such a measurement, the demodulator and signal samples must be linear and free of distortion.

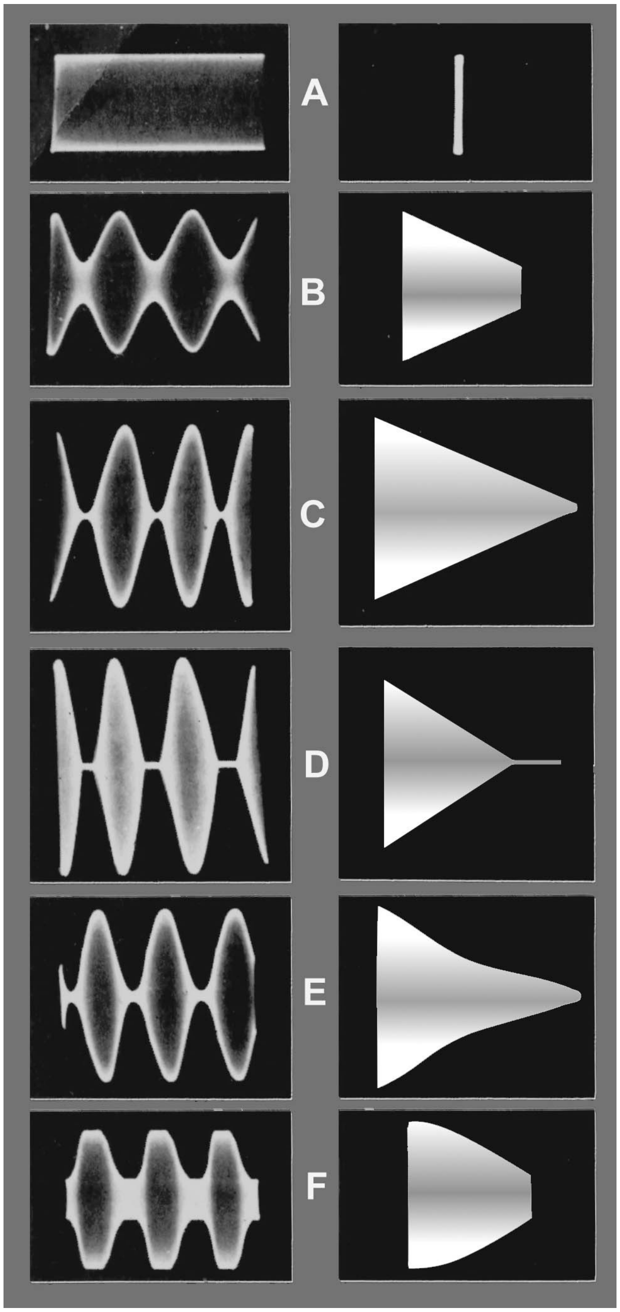

Figure 5 Oscilloscope patterns showing various forms of modulation of an RF amplifier

Figure 5. Wave-envelope patterns; with their corresponding trapezoidal patterns. The wave envelope patterns were obtained with a linear oscilloscope sweep having a frequency one-third of sine-wave audio modulating frequency, so that three cycles of the modulation envelope may be seen.

A. Unmodulated carrier

B. Approximately 50% modulation

C. 100% modulation

D. Shows modulation in excess of 100%

E. Modulation improper transmitter adjustment

F. Indicates improper modulation, incorrect bias, and clipping

HF amplifier distortion measurements

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However, if overdriven or not properly tuned, the potential distortion products can cause severe problems such as unintelligible modulation. RF power being transmitted out of band, thus causing interference with other radio communications. The interfering signals are the result of harmonic and intermediation products – sometimes referred to as "splatter".

Efficiency

Another byproduct of improper linear amplifier operation is inefficiency. Power that is not converted to a useful signal is dissipated as heat. Power Amplifiers that have low efficiency have high levels of heat dissipation, which could be a limiting factor in a particular design. This can have an adverse effect on the components, particularly the final output tubes or transistors.

Instability

Another undesirable amplifier phenomenon is instability. Instability in RF amplifiers may manifest itself as oscillation at almost any frequency, and may damage or destroy the amplifying device. This unwanted RF energy is called spurious oscillation.

These spurious oscillations can arise at specific or very wide ranging frequencies and over a particular bias, drive level, temperature or output load impedance.

Responsibility

In the amateur radio service, the control operator (i.e. ham) is responsible for ensuring that all emitted signals including RF linear power amplifiers are operated in accordance with those prescribed by their license privileges and do not exceed the maximum allowed distortion by the FCC.

Some practical theory

In practice and to ensure efficiency, many linear amplifiers operate as Class B. In Class B the conduction angle for the amplifying device (tube or transistor) is approximately

180°. Thus, the amplifying device conducts only half of the time, either on positive or negative half cycle of the input signal.

The same as in Class A, the DC bias applied to the amplifying device determines the Class B operation. Class B amplifiers are more efficient than Class-A amplifiers. The instantaneous efficiency of a Class-B PA varies with the output voltage and for an ideal PA reaches π/4 (78.5 %) at PEP. However they are much less linear. Therefore a typical Class-B amplifier will produce quite a bit of harmonic distortion that must be filtered from the amplified signal.

PDC = (2*VCC*V) / (Π*R);

PLOAD = V2 / (2*R);

η (Efficiency Class-B) = (Π*V) / (4*VCC)

A common configuration of Class B amplifiers is push-pull. In this configuration, one amplifying device conducts during positive half cycles of the input signal and the second transistor conducts during the negative half cycle. In this way, the entire input signal is reproduced at the output. In the push-pull arrangement, the DC components and even harmonics cancel, (but odd harmonics add), thus the output contains the fundamental signal only. Note that the cancellation of odd harmonics is only valid if the amplifier is not driven hard.

Power amplifier linearity

When two or more signals are input to an amplifier simultaneously, the second, third, and higher-order intermodulation components (IM) are caused by the sum and difference products of each of the fundamental input signals and their associated harmonics. The rated PEP of a Power Amplifier is the maximum envelope power of a two-tone signal for which the amplifier intermodulation level is -30dBc. When two signals at frequencies f1 and f2 are input to any nonlinear amplifier, the following output components will result:

Fundamental: f1, f2

Second order: 2f1, 2f2, f1 + f2, f1 - f2

Third order: 3f1, 3f2, 2f1 ± f2, 2f2 ± f1,

Fourth order: 4f1, 4f2, 2f2 ± 2f1,

Fifth order: 5f1, 5f2, 3f1 ± 2f2, 3f2 ± 2f1, + Higher order terms

The odd order intermodulation products (2f1-f2, 2f2-f1, 3f1-2f2, 3f2-2f1, etc) are close to the two fundamental tone frequencies f1 and f2.

The nonlinearity of a Power Amplifier can be measured on the basis of the generated spectra (i.e with a spectrum analyzer) than on variations of the fundamental waveform (i.e. oscilloscope). The estimation of the amplitude change (in dB) of the intermodulation components (IM) versus fundamental level change, is equal to the order of nonlinearity.

For a one dB increase of fundamental level (f1 and f2), the level of IM2 will go up by 2dB, the level of IM3 will go up by 3dB, and so on. As a relation between the degree of nonlinearity (third, fifth, etc) and the frequency of the side tone (such as IM3, IM5, etc). It can be mentioned with the IM5 tones are not affected by third-degree nonlinearities, but IM3 tones are functions of both third- and fifth-degree (and higher) nonlinearities. That means at low signal amplitudes, where the fifth-order distortion products can be neglected, the amplitudes of the IM3 tones are proportional to the third power of the input amplitude (see below).

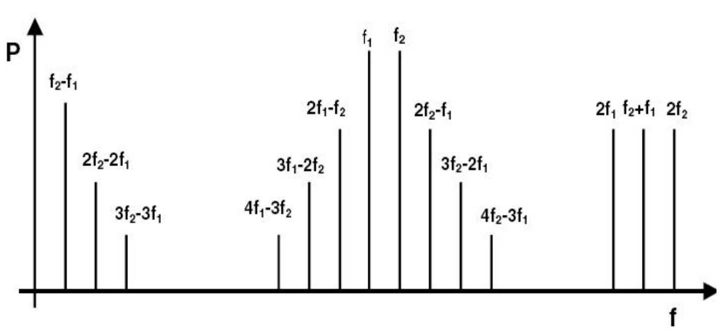

Figure 6 Spectrum of IMD products

- end -

Application Note #2

Transmitter Two Tone Test

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Basically there are three methods to measure non-linearity and IMD products; Spectrum analyzer, Trapezoid test and the Two Tone test. Each has its advantages and disadvantages.

Two tone signal test using an oscilloscope

The Two Tone, third-order intermodulation distortion (IM) test measures the degree of nonlinearity of an electronic device with a definable dynamic range, such as an amplifier. Nonlinear RF Amplifiers may spread signals into adjacent channels, and or frequencies, which can cause Cross Modulation. This is based on the same phenomena as third order intermodulation for nonlinear amplifiers with two-tone inputs. Fortunately, when making a Two Tone test, the signal distortion is relatively easy to spot. Testing ham radio HF linear amplifiers can be done by injecting a two tone test signal (usually 700 Hz and 1900Hz) into the amplifier input and observing the modulation envelope with an oscilloscope.

This test can be done by simply inputting the signal into the transceiver's mike jack from the line out of the two tone test generator, or in the alternative, using the speaker output to drive an external speaker placed close to the microphone. The drawback to using this technique is that the microphone and speaker performance may impact the test result. For this reason separate level and balance control are provided.

Fig 1. This Two Tone test generator provides a standard 2-tone (700 and 1900 Hz) audio source. This type of testing is most commonly used as a measure of transmitter linearity for amateur radio equipment.

Figure 1 The TTG1 Two Tone test generator

Results of 2-tone IMD tests can be found in every ARRL review of new transceivers and power amplifiers.

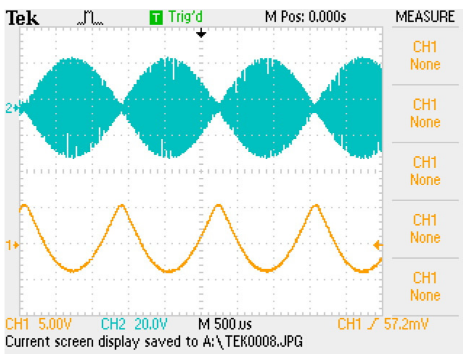


Figure 2 Scope Two Tone test display

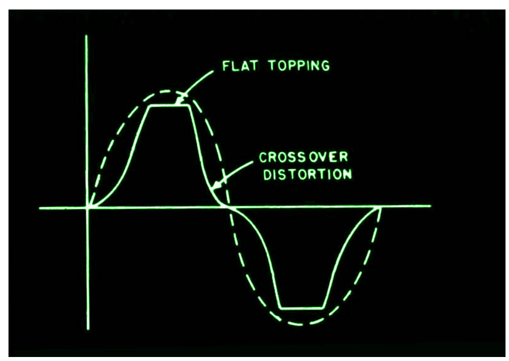
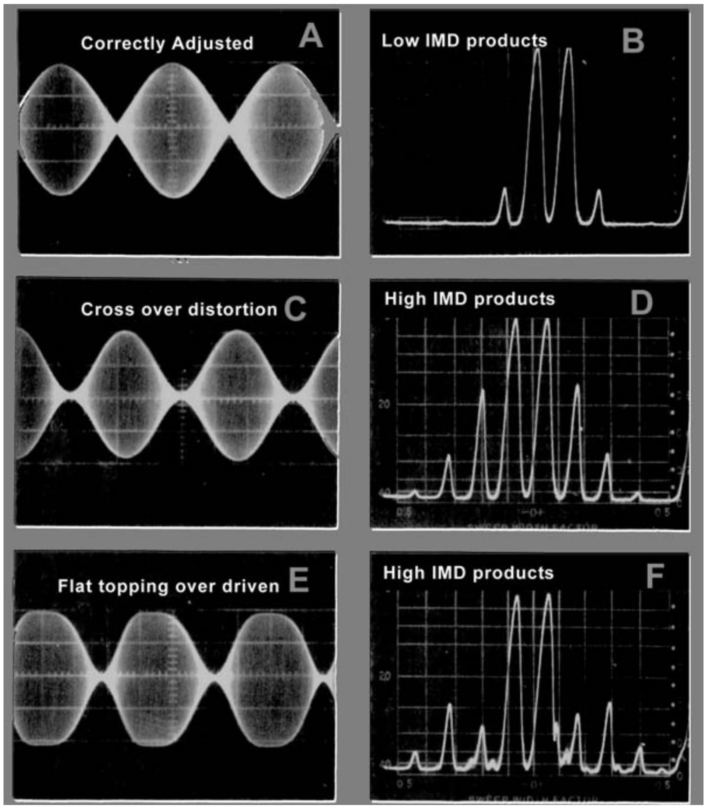


Figure 3 Two Tone test characteristics

Fig. 2 Top trace, note the lack of flat topping and or cross over distortion as when compared to the example shown in Fig 3.



A & B Properly adjusted transmitter.

C & D Crossover distortion resulting in high IMD products.

E & F Flat topping over driven modulation causing splatter and high IMD products.

Figure 4 Tow Tone test, left column scope display, right column spectrum analyzer display

HF amplifier distortion measurements

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However, if overdriven or not properly tuned, the potential distortion products can cause severe problems such as unintelligible modulation. RF power being transmitted out of band, thus causing interference with other radio communications. The interfering signals are the result of harmonic and intermediation products – sometimes referred to as “splatter”.

Efficiency

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Instability

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These spurious oscillations can arise at specific or very wide ranging frequencies and over a particular bias, drive level, temperature or output load impedance.

Responsibility

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Some practical theory

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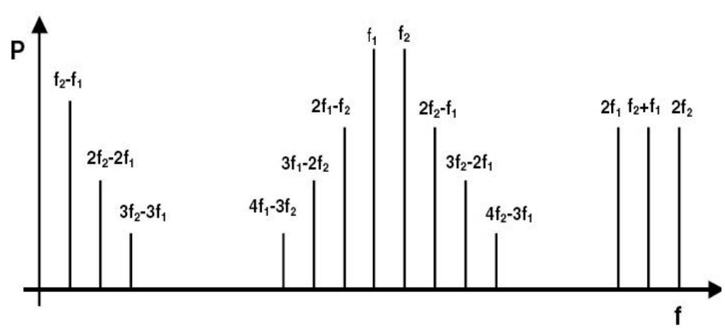

Figure 5 Spectrum of IMD products

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Application Note #3

Transmitter Spectrum Analyzer Test

How to measure amplifier non-linearity

How can the ham make practical amplifier measurements to check for splatter, IMD products and unwanted harmonic content? While it is helpful to understand the theory and math behind amplifier non-linearity's, all that is required is to make accurate measurements minimizing splatter and distortion is an understanding of the basic concepts.

Basically there are three methods to measure non-linearity and IMD products; Spectrum analyzer, Trapezoid test and the Two Tone test. Each has its advantages and disadvantages.

RF spectrum analyzer

The spectrum analyzer method is used to make quantifiable laboratory grade measurements. The spectrum analyzer test measures the amplifier's RF spectra distribution as a function of frequency, generally in MHz (X axis) and magnitude (Y axis). By comparing the level of the fundamental frequency to the level of any unwanted harmonic products (spurious) a precise quantifiable results can be obtained. The results are generally shown in the amount that the spurious products are down compared to the fundamental frequency and are express in dB. For example if the ratio is -20 dB than the spurious power is 1/100 that of the fundamental, and if the level was down -30 dB than the spurious power would be 1/1000 that of the fundamental. It should be noted that the spurious signals could be multiples of odd and even harmonics.

For convenience, some spectrum analyzer are equipped with a tracking generator. The tracking generator, by sweeping the amplifier's frequency, allows for measuring the performance at more than one frequency.

The frequency range of the spectrum analyzer should be at least five times the highest fundamental frequency to be measured. In the case of a 50MHz signal (6 meter) test, that would require a spectrum analyzer with a 250MHz frequency range or higher.

An alternative to the spectrum analyzer is using an oscilloscope that provides a Fast Fourier Transform (FFT) display. The FFT display is derived from the mathematical transformation of the signal's magnitude versus time components and then displayed as magnitude versus frequency. However, the oscilloscope's bandwidth needs to also be at least five times the fundamental frequency.

Either of these methods yields predictable and accurate measurement results. When comparing spectrum analyzer results to oscilloscope FFT results, and so long that the signal being measured is with the dynamic and frequency range of the measurement instrument input, the results are generally within .5dB.

Measurement considerations

While it is relatively straight forward to make these measurements, you need to consider the transceiver and or amplifier signal levels and the modulation being applied to their input. One can't simply just connect the high power RF output from the amplifier or transceiver directly to the input of the spectrum analyzer. This may result in damage to the instrument and may even cause injury to people. The power levels of even QRP transceivers are about 5 Watts and for transceivers can be as high as 200 Watts. Linear amplifiers can have levels of greater than 1,500 Watts. The levels are way out of the input range of a typical spectrum analyzer or oscilloscope. One could use attenuators. These would have to be wideband and high power. That means lots of money. The most economical way to condition the desired signal level is to use an RF sampler/coupler (see Figure 2).

While it is possible to just modulate the transmitter by speaking or whistling in to the microphone and observe the results, it is not recommended. Nonetheless, if such technique is used, it has been found that repeatable speaking the letter "X" will give the best results. The "ae" sound produces medium frequency levels for a long enough duration, and the "ks" sound produces higher frequency components to make some useful measurements. However, to make repeatable quantifiable measurements, a precision signal source, such as in Fig 3. below, a Two Tone test generator, is required.

Figure 1 3GHz spectrum analyzer

Fig. 1 The DSA1030A series spectrum analyzer with advanced measurement capabilities. It uses digital IF technology which guarantees the reliability and performance required meeting the most demanding RF applications for measuring amplifier no-linearity. This spectrum analyzer will make virtually all ham radio measurements and test. Price is about $5,000

Figure 2 HFS-1.5 HF sampler/coupler -30dB

Fig. 2 This wideband RF sample/coupler inductively couples a sample of high power RF (up to 1.5 KW PEP) passing from the RF IN to the RF OUT connector. This signal is coupled at -30 dB, a power reduction of 1000:1. The sampled signal is very useful for analyzing HF signals on an oscilloscope and spectrum analyzer.

Figure 3 TTG1 Two tone test generator

Fig . 3 The TTG1 Two Tone test generator is an excellent oscillator to test SSB transmitter performance such as Inter-modulation Distortion (IMD). The TTG1 was created to deliver a standard 2-tone (700 and 1900 Hz) audio source for testing of SSB transceivers and linear amplifiers. This type of testing is used as a measure of transmitter linearity for amateur radio transmitters.

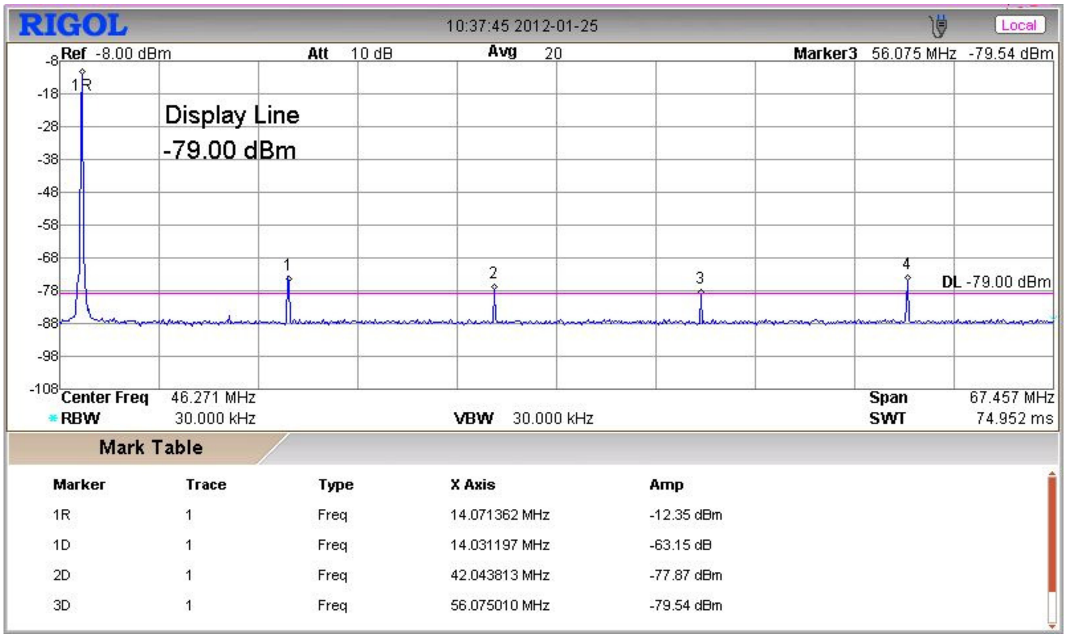

Figure 4 Spectrum analyzer display

Fig. 4 This example is a typical spectrum analyzer display. The fundamental frequency is the large spike at the start of the sweep. The harmonic contents are the smaller spikes. They are down from the fundamental frequency about 70 dB. The FCC minimum levels for the amateur radio service for unwanted IMD products is around -20dB. So in the example in Fig.4, this level would be considered excellent performance.

Summery

Using a spectrum analyzer or an oscilloscope with an FFT for amplifier distortion measurements provides the most accurate and repeatable results. However, there are always tradeoffs, see Fig. 5 below.

Test Equipment	Spectrum analyzer, RF Sampler/coupler, dummy load, signal source such as a two tone test generator.
Advantage	Results are displayed in quantifiable units, easy to duplicate, accurate to a dB or less.
Disadvantage	New laboratory grade spectrum analyzers are expensive.

Figure 5 advantages and disadvantages

Careful shopping on eBay will reveal that affordable used spectrum analyzers are available. It is recommended if such purchase is contemplated that it is calibrated with NIST certified equipment.

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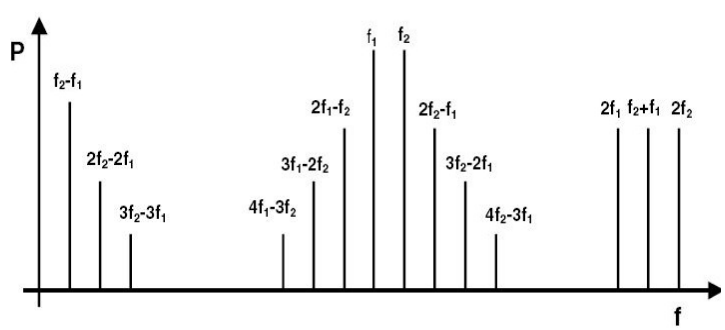

Figure 6 Spectrum of IMD products

Application Note #4
Measuring Transmitter Power with the Oscilloscope

HF Amplifier Power Measurements:

Power is often defined as peak power, carrier power, average power, Peak Envelope Power (PEP) and sometimes incorrectly as RMS power. In the United States the Federal Communications Commission uses PEP to set maximum power limits for amateur radio transmitters. The maximum power allowed on certain frequencies using SSB modulation is 1,500 Watts PEP. PEP is the average power supplied by the transmitter/linear RF amplifier to the transmission line and eventually the antenna, during one radio frequency cycle at the crest of the modulation envelope, under normal operating conditions.

What is Electric Power

Electric power is the rate at which electric energy is transferred by an electric circuit. The unit of power is the Watt. Joule heating, is ohmic heating and resistive heating, it is the process by which the passage of an electric current through a conductor releases heat. It was initially studied by James Prescott Joule in 1841.There is potential power (no heating), instantaneous power and average power. When one volt is applied across a one ohm resistor, one ampere of current flows though the resistor. Since P=IE then the resistor is dissipating one Watt.

When power is defined over time it is expressed in Joules. One Joule equals one Watt per second; that is synonymous to one Watt second. When power is referred to as “instantaneous power”, it is expressed as a fraction of a Joule; for example if the instant of that power lasts for one Millisecond, that equals one Millijoule or one Milliwatt second.

It should be noted that the power applied to an ideal antenna does not heat the antenna. The antenna radiates the power (less any losses, which indeed heat the antenna). This radiated power is eventually absorbed by the atmosphere, natural and manmade objects, and also at the radio receiver's antenna and receiving circuits; eventually the energy is transformed either to useful work, or heat.

PEP may be more difficult to measure than CW power. Nonetheless, PEP is the average power during one radio frequency cycle at the crest of the modulation envelope and continuous wave (CW) power is also an average power, thus they are equal.

All power measurements rely on these formulas:

For DC power measurements, use:

$$P = E^2/R$$

For AC, PEP measurements, use:

$$P = (E_{avg})^2/R$$

Peak Power

Measuring peak voltage with an oscilloscope is not difficult, and generally, the load impedance (R) in amateur radio transmitters and transmission lines equals 50 ohms (sometimes 300 ohms). Making a peak power measurement is straightforward by measuring the peak voltage. Some users find it simpler to measure the peak-peak voltage. In that case, divide the results by 2 to get peak voltage.

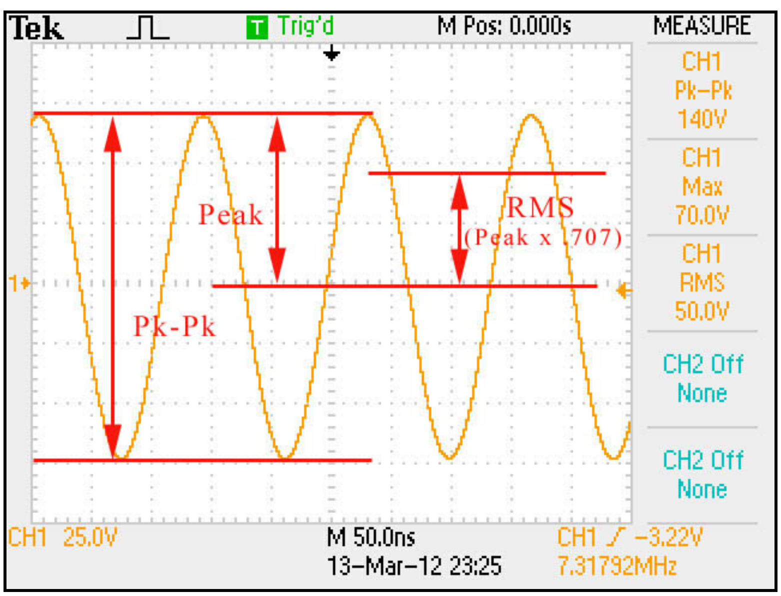

Figure 1, Oscilloscope display

See figure 1. Note, the Max (peak) voltage is 70 volts, which is half of the Pk-Pk voltage. To solve for peak power, square the peak voltage (*E*) and divide by 50 (the load impedance) So:

$$(70)^2 = 4900/50 = 98 \text{ watts Peak Power}$$

Average Power

Let's dispel the myth of RMS power. There is no such thing as RMS power. RMS is an abbreviation of Root Mean Squared. The term "Mean" is just another word for average. With respect to power calculations, the AC RMS voltage is the equivalent to the DC voltage. For example, 25V RMS or 25V DC across a non-inductive 50ohm load results in identical power dissipation of 12.5 watts in either case.

The RMS value by itself is not the comparable heating power and it doesn't correspond to any useful physical quantity; no heat, no power. Recall P=IE, and I=E/R. Voltage (E), nor current (I) by themselves generate power. The power is only produced when a current is induced by a voltage across a load R. Finally, RMS and average values of nearly all waveforms are different. One exception is a steady DC waveform, for which the average, RMS, and peak values are identical.

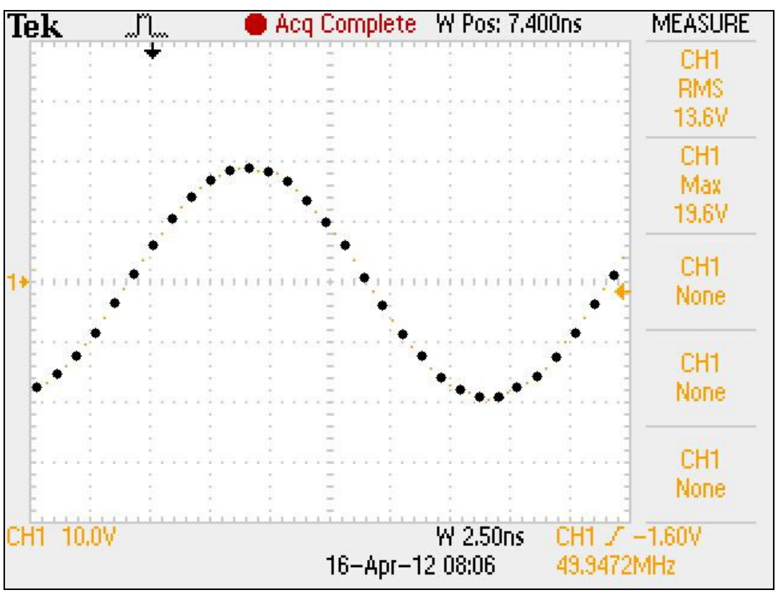

Figure 2 sampled sine wave

See figure 2. If one were to sample a waveform at regularly spaced times, and then add up their values and divide that total by the number of samples taken, one would have approximately the average value of whatever the waveform represents; this could be voltage, current or power. The less the time intervals between samples, the more accurate the average will be. The mathematical integration is a method to find what the value would be if we could minimize the time interval really close to zero. That's important if we want to calculate the exact average value of some waveforms. The corresponding formula for a continuous function (or waveform) $f(t)$ defined over the interval

$$T_1 \leq t \leq T_2$$

$$f_{\mathrm{rms}} = \sqrt{\frac{1}{T_2 - T_1} \int_{T_1}^{T_2} [f(t)]^2 \, dt},$$

Thus substitution volts for f we arrive at the familiar equation:

$$V\mathrm{rms} = \sqrt{\frac{1}{T_2 - T_1} \int_{T_1}^{T_2} [f(t)]^2 \, dt}$$

Fortunately this mathematical integration can be reduced to the RMS values. And for an ideal sine wave that happens to be the peak waveform value multiplied by .707. So power is still power, whether PEP or average. The correct way to express average AC power is P_{avg}. as a result:

$$P_{avg} = (E_{RMS})^2 / R$$

We know that oscilloscopes are great tools to measure voltages of AC waveforms. For a perfect sine wave, multiplying peak voltage times .707 will give us RMS voltage. Some of the newer scopes make this step easy and calculate the RMS voltage directly. These calculations are quite accurate even for non-sinusoidal waveforms. See Figure 1, what is the average power if the peak voltage is 70 volts?

Step one: Calculate the RMS voltage (70 x .707 = 49.5 volts). You'll note that in Figure 1, the calculated RMS voltage is 50 volts not 49.5 volts. That is because the oscilloscope measurement was done on a somewhat imperfect sine wave, thus giving a slightly higher reading.

Step two: Square the voltage, and divide by 50 (the load impedance):

$$(49.5)^2 = 2450/50 = 49 \text{ watts Avg.}$$

A common error some make calculating average power is they multiply PEP by .707, i.e. 98 watts times .707. This results in an <u>incorrect</u> answer of 69.3 watts. Power is always calculated by squaring the voltage and dividing the result by the load impedance.

PEP Power

The International Telecommunication Union (ITU) Radio Regulations define the terms Peak Envelope Power (PEP) as:

"PEP means the average power supplied to the antenna transmission line by a transmitter during one radio frequency cycle at the crest of the modulation envelope taken under normal operating conditions."

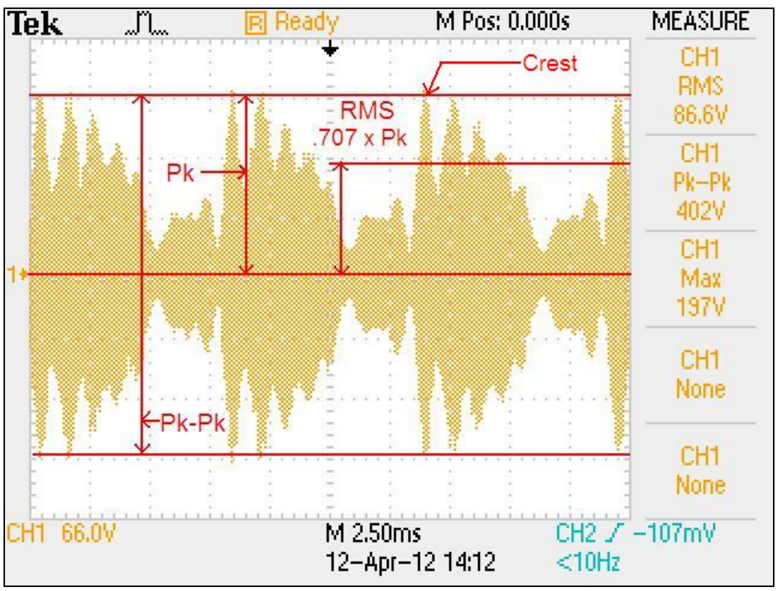

Figure 3 SSB Modulation

Understanding the definition of PEP, the question then arises what is meant by "*radio frequency cycle at the crest of the modulation envelope*"?

See figure 3. The crest of the modulation envelope is the peak value; this oscilloscope labels it as "Max". It is 197 volts. Note, that the scope calculated the RMS voltage as 86.6 volts which would be correct for a steady RF carrier. However, here the modulation is not a steady carrier, but instead represents the minimum and maximum modulation levels as an envelope.

The modulation envelope duration is 25mS over the entire display duration. At 7.3 MHz, this duration contains 182,500 individual radio frequency cycles. Since the scope calculates RMS voltage over all these cycles, we cannot rely on that calculation. So how do we obtain the RMS voltage for one radio frequency cycle? We know by examining the scope display that there must be at least one radio frequency cycle at the crest of the modulation envelope. The peak value of that cycle 197 volts. We also know that RMS voltage equals .707 times the peak voltage; so 197 x .707=139.3 volts. To calculate PEP power we again use this formula:

$$P_{avg}=(E_{RMS})^2/R$$

PEP means the average power. So we can substitute PEP for Pavg. Thus:

$$PEP=(E_{RMS})^2/R$$

Accordingly, applying this formula yields:

$$PEP = (139.3)^2=19,404/50=388\ watts.$$

Amplitude Modulation (AM) Power

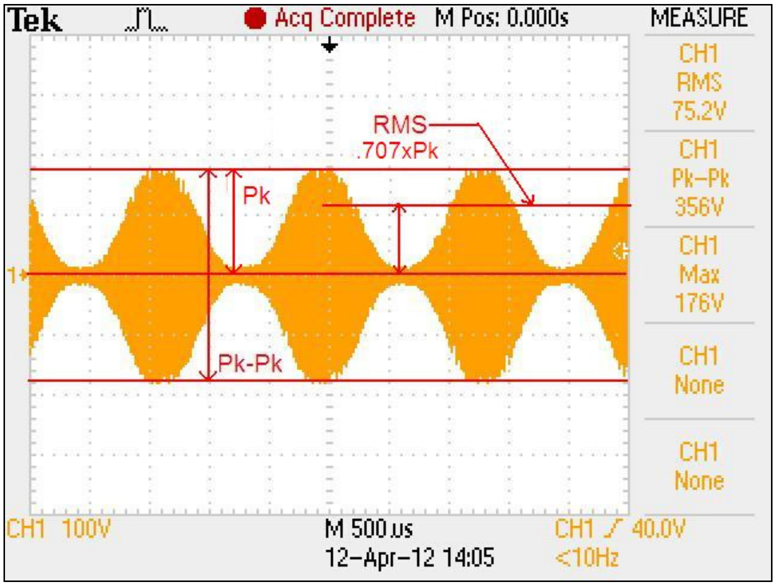

Figure 4 Approximately 90% AM modulation

See Figure 4, the PEP output of an AM transmitter at full modulation is four times its carrier PEP; in other words, a 100-watt amateur transceiver is usually rated for no more than 25 watts carrier output when operating in AM mode.

RF Power Meters

Absent a PEP function, virtually all analog power meters measure average power. Many low cost power meters are notoriously inaccurate as they are generally not calibrated to a known power standard. The Bird model 43®, and Heathkit® HM-102 are an exception. Their accuracy is guaranteed to better than 5% of full scale. The HM-102 employs an internal accurate calibration standard to which the meter is calibrated. The Bird® 43 slugs are individually calibrated at the factory against an accurate RF power source.

As a result, these two analog power meters are generally much more accurate than 5%. The accuracy of any meter can be verified to better 2% (the scope vertical amplifier specification) when an oscilloscope is used to measure the power and comparing that result to an unknown power meter.

PEP RF Power Meters

You may notice that your oscilloscope PEP measurements are consistently higher than that obtained by most RF watt meters. There is nothing wrong. That is because the oscilloscope, with its very fast rise time, can measure PEP based on peak voltages.

Most commercially available watt meters display average power only. Some RF meters employ a "PEP" function. They do this with a sample and hold circuit. This circuit needs to have a fast rise time, i.e. considerably faster than the PEP envelope components. Even then, some of these meters may not accurately measure the true PEP power. As a result, their PEP reading can be significantly lower.

One of the most reliable ways to confirm the accuracy of any analog or digital power meter is by using an oscilloscope with a calibrated vertical amplifier and sufficient bandwidth (normally twice the measured frequency).

Direct sampling power measurements with the Oscilloscope

Most modern oscilloscopes have a maximum calibrated vertical amplifier deflection factor of about 5 volts/division. With an accurate X10 voltage probe, the vertical amplifier can display up to 50 volts/division. Given the 8 vertical divisions normally found on an oscilloscope display, the maximum voltage measurement with a 10x probe is 400 volts peak to peak. That is the equivalent of 141.4 volts RMS, which in turn calculates to 400 watts into 50 ohms. Any measurement of greater power requires an RF sampler (discussed below).

The method used to make the power measurements is called direct sampling. Set the scope to the appropriate volts/division setting, sample the voltage applied to the center conductor of transmission line with an accurate 10x probe which in turn is connected to the scope's vertical input. This can be done with a "T" connection. For HF frequencies, the measured voltage will be quite accurate. The closer the sampling point is to the transmitter the better.

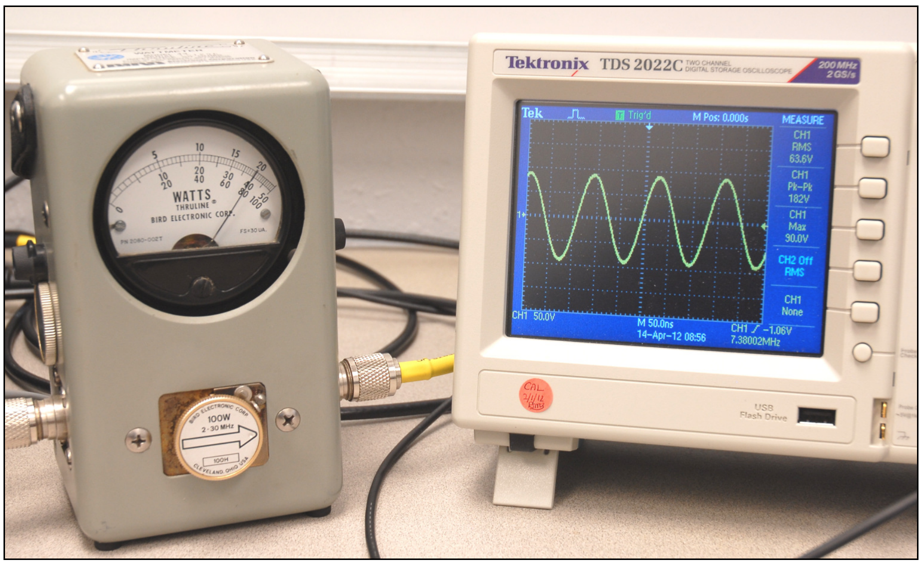

Figure 5 direct power measurement

The primary errors will be the result of the combined uncertainty of the specifications of the vertical amplifier and the 10x probe. However, these can be checked with an accurate signal generator such as a Tektronix® SG 503 constant amplitude calibration generator. My experience has been that the accuracy achieved is better than 2% and correlates with a Bird 43 ® power meter. In figure 5 the scope measured 63.6 volts RMS which equals 80.89 watts, and the Bird 43® indicates 80 watts. That's within +/- 1% well within the specifications of either instrument.

Problems with Direct Voltage Measurements

While it is possible to make direct high power measurements with an oscilloscope, such as 1500 watts, it is rarely done. For example, at 1500 watts, the RMS voltage is 274 volts, 388 volts peak, and 775.1 volts peak to peak. These levels can result in damaged equipment and potential injury or death to the operator. It is possible to use high power attenuators to reduce the voltage levels to a practical level. However, these attenuators must be able to take the full brunt of the maximum power to be measured.

The RF Sampler or Coupler (sampler)

Fortunately using an oscilloscope with an appropriate sampler is uncomplicated and provides accurate results of better than 1 db. One way is to measure the power is to load the amplifier directly into an antenna capable of radiating the maximum applied power. The preferred and most practical way, is to load the amplifier in a "dummy load", which is usually 50ohms. These dummy loads are inexpensive and simple to connect. But how do you measure the high voltage input to the dummy load? The best way to do this is with an RF sampler. The sampler reduces the power to a manageable level. The most common power reduction is a 1000 to 1. This equals -30 dB of the RF being sampled.

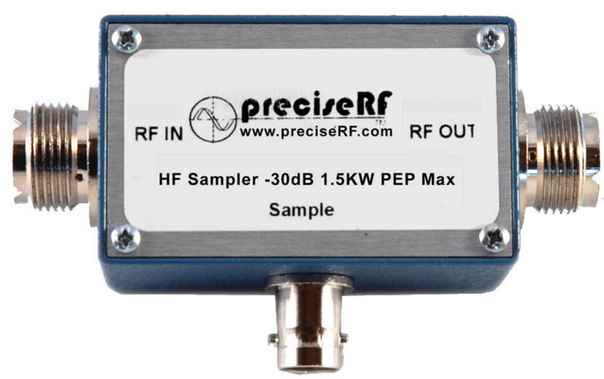

Figure 6 The HFS-1.5 HF sampler/coupler covers 2MHz to 50MHz. Employing a wideband transformer; it samples the high power RF (up to 1.5 KW PEP) from 2MHz to 50MHz.

Figure 6, The HFS-1.5 HF sampler allows for a maximum input of 1.5 KW, at HF frequencies from 2MHz to 50MHz with a sample transfer coefficient of - 30dB or a 1000:1 reduction in power. Most samplers, including this example, provide a "Sample" port which requires a 50ohm termination. Nearly all scopes have 1 Meg ohm vertical input impedance and do not have the necessary 50ohm input impedance. As a result, they require a 50ohm pass-through terminator at the scope's vertical input. However, when connecting the sampler to a spectrum analyzer, a pass-through terminator is not required since most spectrum analyzers have a 50ohm input impedance.

Using the Sampler to Measure Power

An ideal sampler with a -30 dB power reduction (1000:1) has a voltage gain/loss of 31.62. Therefore, to find the equivalent voltage (peak, pk-pk, or RMS) for -30dB, multiply the sampled voltage times 31.62. Using peak voltage allows for calculating power at any instant in time. If the peak voltage at the sampler port equals 2.83 volts, than the actual RMS voltage is 2.00 volts (2.83 x .707). Multiplying the RMS voltage by the gain/loss ratio 2.00×31.62 which we get 63.3 volts RMS. Calculate power using:

$P = E^2/R$ equals 80.05 watts

When comparing that measurement using a calibrated -30 dB sampler to the direct measurement in figure 5, (80.05 watts) correlates well with the previous direct measurement of 80.89 watts, and the Bird 43® indication of 80 watts.

Here is a sampler with a -20 dB power reduction (100:1) it has a voltage gain/loss of 10. To find the equivalent voltage (peak, pk-pk, or RMS) for -20 dB, multiply the sampled voltage times 10. For instance, if the peak voltage at the sampler equals 5 volts, that equals 3.54 volts RMS. The voltage at the sampler input is 3.54×10 which equals 35.4 volts RMS. Calculate power using:

$P = E^2/R$ equals 25.06 watts

A much quicker way to obtain PEP and average power measurement is to refer to Table 2 below. Connect the output of sampler port to the oscilloscope input (use a 50 ohm terminator at the scope input) and note the peak voltage. Move to the right and read the peak and average power in watts, or power expressed in dBm.

Other handy formulas

To obtain the voltage ratio from a given dB, use this formula, where $A=dB$

$$V2/V1=10^{(A20)}$$

To obtain the RMS voltage for any power use:

$$V=\sqrt{(PxR)}$$

So for a 100 watts and 50 ohms than the RMS voltage is calculated thus:

$$V=\sqrt{(100x50)}=70.71 \text{ volts RMS}$$

What are dB

Decibels state a power ratio, not an amount. They tell how many times more (positive dB) or less (negative dB) but not how much more or less in absolute terms. Decibels are logarithmic, not linear. For example, 20 dB is not twice the power ratio of 10 dB. Use this equation to find decibels: A = 10*log10 (P2/P1) (dB) where P1 is the power being measured, and P1 is the reference to which P2 is being compared. To convert from decibel measure back to power ratio: P2/P1=10^(A/10). Voltage is more easily measured than power, making it generally more convenient to use: A = 20*log10(V2/V1), Where A=voltage ratio. The equation for obtaining voltage ratio from dB is V2/V1 = 10^(A/20).

What are dBm

dBm is an abbreviation for the power ratio in decibels (dB) of the measured power referenced to one milliwatt (mW). It is used in radio and microwave equipment as a convenient measure of absolute power because of its capability to express both very large and very small values in a short form.

Compare dBW, which is referenced to one watt (1000 mW). Since it is referenced to the watt, it is an absolute unit, used when measuring absolute power. By comparison, the decibel (dB) is a dimensionless unit, used for quantifying the ratio between two values, such as signal-to-noise ratio.

Zero dBm equals one milliwatt. A 3 dB increase represents roughly doubling the power, which means that 3 dBm equals roughly 2 mW. For a 3 dB decrease, the power is reduced by about one half, making –3 dBm equal to about 0.5 milliwatt. To express an arbitrary power *P* as *x* dBm, or vice versa, the following equations may be used:

$$X=10\log_{10}(100P) \text{ or } x=10\log_{10}P+30 \text{ and}$$

$$P=10^{(x/10)}/1000 \text{ or } P=10^{(x-30)}/10$$

Table 1. Relationship of dBm to power and application

dBm level	Power	Application
80 dBm	100 kW	Typical transmission power of FM and TV radio station with about a 31 mile range.
60 dBm	1 kW	Typical combined radiated RF power of microwave oven elements. Approximate maximum RF output power from a ham radio transceiver allowed.
50 dBm	100 W	Typical thermal radiation emitted by a human body. Typical maximum output RF power from a ham radio HF transceiver.
40 dBm	10 W	Typical PLC (Power Line Carrier) Transmit Power.
37 dBm	5 W	Typical maximum output RF power from a handheld ham radio VHF/UHF transceiver.
36 dBm	4 W	Typical maximum output power for a Citizens' band radio station (27 MHz) in many countries.
33 dBm	2 W	Maximum output from a UMTS/3G mobile phone (Power class 1 mobiles). Maximum output from a GSM850/900 mobile phone.
27 dBm	500 mW	Typical cellular phone transmission power. Maximum output from a UMTS/3G mobile phone (Power class 2 mobiles).
20 dBm	100 mW	Bluetooth Class 1 radio. Maximum output power from unlicensed AM transmitter per U.S. Federal Communications Commission (FCC) rules.
–73 dBm	50.12 pW	"S9" signal strength, a strong signal, on the S-meter of a typical ham or shortwave radio receiver.
–127.5 dBm	0.178 fW	Typical received signal power from a GPS satellite
–174 dBm	0.004 aW	Thermal noise floor for 1 Hz bandwidth at room temperature (20 °C)

See Table 1 for typical power levels. It provides an insight into the relationship of dBm to power and application. Note the very wide power range in dBm that can be easily expressed

Making Precision RF Power Measurements

Table 2. It shows calculations for power measurements using an ideal -30dB sampler. Affordable samplers usually have a specification of -30dB +/- 1 db. A 1dB error could have an effect on the measurement accuracy.

Table 2. Sine wave peak and average watts and dBm power, with a -30db sampler

Peak volts at sampler In/Out	Peak volts at sample port (-30db)	Peak Power Watts	Avg. Power Watts	Power in dBm
16	0.50	5	3	37.10
32	1.00	20	10	43.11
47	1.50	44	22	46.45
63	2.00	79	40	48.10
79	2.50	125	62	50.16
95	3.00	181	90	52.56
111	3.50	246	123	53.92
126	4.00	318	159	55.02
142	4.50	403	202	56.06
158	5.00	499	250	56.98
174	5.50	606	303	57.82
190	6.00	722	361	58.59
206	6.50	849	424	59.23
221	7.00	977	488	59.90
237	7.50	1123	562	50.51
253	8.00	1280	640	61.07
269	8.50	1447	723	61.61
285	9.00	1625	812	62.11
300	9.50	1800	900	62.55
316	10.00	1997	998	63.00
332	10.50	2204	1102	63.43
348	11.00	2422	1211	63.84
364	11.50	2650	1325	64.23
379	12.00	2873	1436	64.58
395	12.50	3121	1560	64.94

Consider an ideal sampler with a sample coefficient -30dB. That equates to a voltage gain/loss of 31.62. Then consider a sampler that has a sample coefficient of -31dB. That equates to voltage gain/loss of 35.48. Now consider a sampler with a sample coefficient of -29dB. That equates to a voltage gain/loss of 29.18.

Table 3. Calculates power for each of these examples based only on the voltage gain/loss, results in the following measurement uncertainty:

Table 3 Measurement uncertainty

	-29dB	Ideal -30dB	-31dB
Voltage gain/loss	29.18	31.62	35.48
Calculated power	17.03 watts	20.00 watts	25.18 watts
Approximate Error	17%	0%	21%

For all practical purposes, a 1dB gain or loss in transmitted and or received signals is insignificant and difficult to notice. However, when checking the accuracy of a component in the transmitter chain or using the device as a “ham shack” standard, measurement uncertainty is important. A 1dB error may be significant.

How to improve measurement accuracy

There are three practical ways to improve the accuracy. One expensive way is to compare your power measurement to a laboratory standard. Another, lower cost alternative is to purchase a sampler which was calibrated with NIST traceable equipment. These samplers have the measured transfer coefficient i.e. -30dB stamped on their enclosure. The third, and least expensive alternative, is to calibrate the sampler yourself. You can do this assuming that your oscilloscope has accurately calibrated vertical amplifiers.

See figure 7, a precise method to improve measurement accuracy is to measure the sampler’s RF input voltage (usually with a 10X probe) and compare that measurement to the sampler port output voltage (terminated into 50 ohms). The yellow (top) trace is the sampler port output, and the blue (bottom) trace is sampler RF input. To calculate the voltage gain/loss, using the Pk-Pk voltages will give greater resolution than RMS voltage. That is because Pk-Pk voltages are greater. In our example, we divide 152V/4.52V equals a measured voltage gain/loss of 33.62. From this we determine that the sampler coefficient is -30.53dB, well within the samplers published specification.

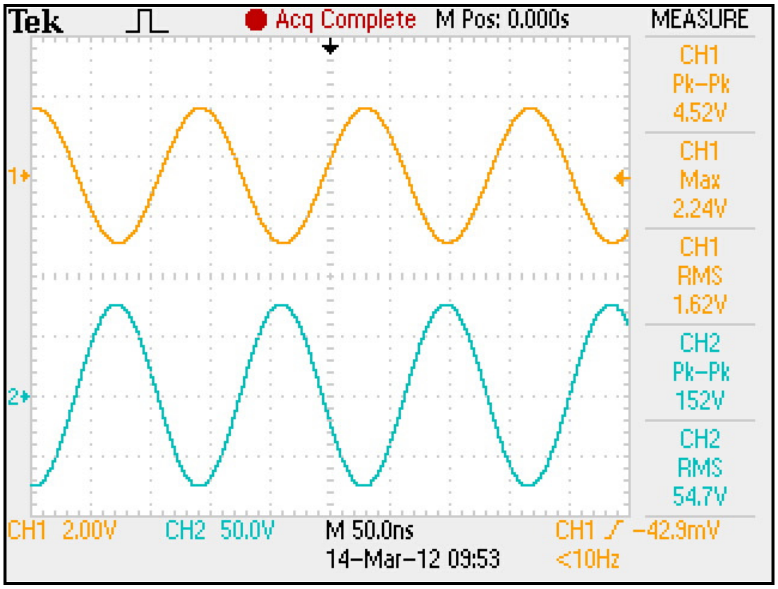

Figure 7 Sampler coefficient

Multiply the sampler port voltage by the sampling gain/loss value and plug it into V_{RMS}. In this case the sampler port voltage Pk-Pk is 4.52V. (The equivalent voltage on the sampler RF input equals 4.52x33.62=152volts). This equals Pk-Pk volts, so we need to calculate RMS:

$$(152 \times .707)/2 \; = 54.08 \text{ volts RMS}$$

That's pretty close to the measurement (54.7 volts) on our scope in figure 7. The minor difference is attributed to the scope's resolution and specification limits. Finally, to calculate the power we use:

$$P_{avg} = (E_{RMS})^2 / R \quad \text{resulting in}$$

$$(54.08)^2/50 = 58.5 \text{ watts}_{avg.}$$

These types of calculations/measurement will be well within +/- 1% of the scope's specifications. This accuracy is better than that obtained by the +/- 1dB uncertainty.

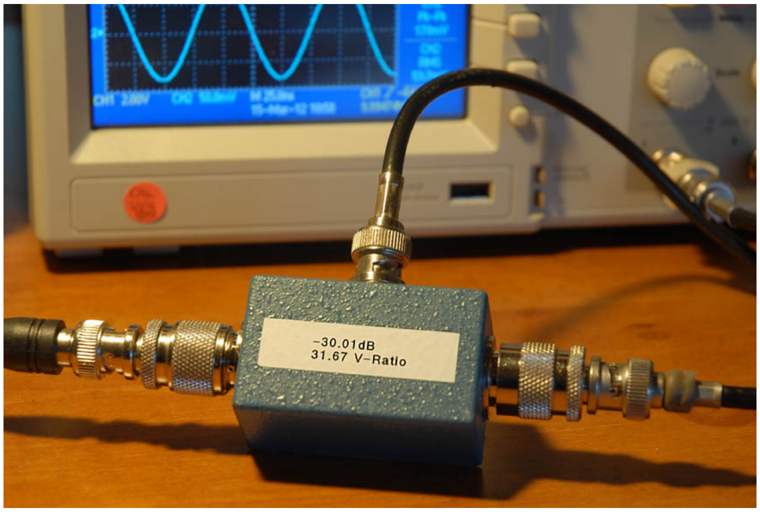

Figure 8, Calibrated Sampler

See figure 8, attach a calibration label on the sampler *"-30.01 dB, & 31.67 V-Ratio"*. Now when you use the "calibrated" sampler, you'll get accurate and consistent results. Table 4, provides a quick way to look-up voltage gain/loss ratios given a sample coefficient from 29dB to 31dB.

Table 4, dB versus voltage gain/loss

Coefficient (dB)	Voltage gain/loss
29.0	28.13
29.1	28.51
29.2	28.84
29.3	29.17
29.4	29.51
29.5	29.85
29.6	30.20
29.7	30.55
29.8	30.90
29.9	31.26
30.0	31.62
30.1	31.99
30.2	32.36
30.3	32.73
30.4	33.11
30.5	33.50
30.6	33.88
30.7	34.28
30.8	34.67
30.9	35.08
31.0	35.48

- end -

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Measuring Return Loss with an oscilloscope and a Return Loss Bridge (RLB)

An RF return loss bridge (RLB) is a wide band bridge which can be used to check the impedance of antennas, coaxial cables, and filters, etc. The ARRL Handbook defines return loss as:

> "*a measure of how closely one impedance matches a reference impedance in phase angle and magnitude. If the reference impedance equals the measured impedance level with a 0°phase difference, it has a return loss of infinity."*

For Ham radio applications This impedance is usually 50-ohms. The INPUT port is normally connected to a test frequency (an RF oscillator or tracking generator from a spectrum analyzer). The DET (detector) is usually connected to an oscilloscope or spectrum analyzer. A RLB is ideal for measuring filter response because return loss measurements are a more sensitive measure of pass band response than insertion-loss measurements.

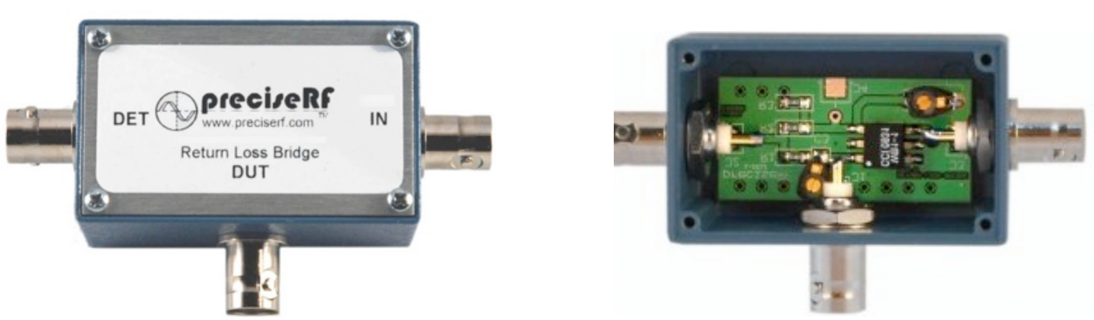

Figure 1. Return Loss Bridge

An RF Return-Loss Bridge

Figure 1. The RLB-I (internal reference – three ports) is a high performance RLB. It is carefully designed specifically for Ham radio applications. It uses a wide band 1:1 minia-

ture SMD 750 MHz transformer. The bridge reference resistors are precision 50 ohm SMD devices. The circuit board employs computer optimized 50-ohm strip line technology.

How to measure return loss with an oscilloscope

There are a number of ways to measure return loss. The method described below relies only on the accuracy of a low-cost step attenuator and removes the scope and signal generator accuracy uncertainty.

> Note: For maximum accuracy, the oscilloscope input impedance should be 50-ohms (this may require an external 50-ohm feedthrough terminator) the step attenuator should be 50-ohms and the signal generator should be 50-ohms as well.

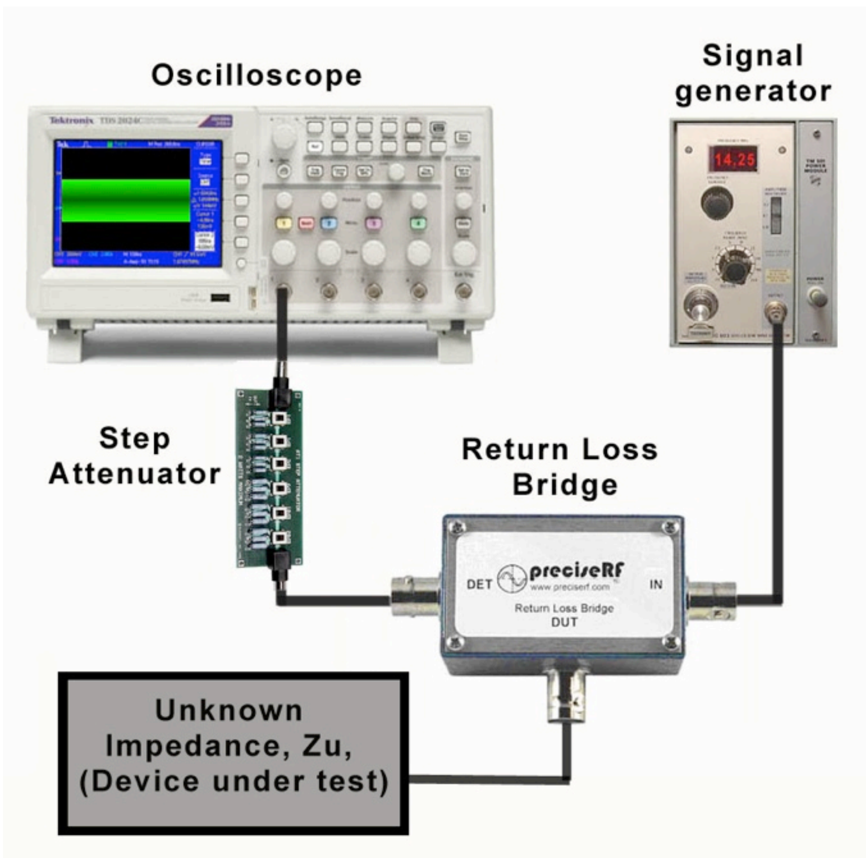

Figure 2. Return loss measurement setup

1. See Figure 2 above. Apply the output of the signal generator to the RF INPUT port of the RLB. It may be necessary to attenuate the generator output to avoid overloading the device under test.

2. Connect the bridge DETECTOR port to an oscilloscope through a step attenuator and leave the (DUT) port of the bridge open circuited.

3. Set the step attenuator for a relatively high level of attenuation of approximately 40 dB, and note the oscilloscope deflection.

4. Adjust the signal generator level and the oscilloscope vertical VOLTS/DIV setting for a convenient six divisions of amplitude as observed on the oscilloscope display.

5. Now connect the unknown impedance of the device under test to the bridge DUT port. The scope reading will decrease.

6. Adjust the step attenuator to produce the same reading obtained when the DUT port was open circuit. The difference between the two measurements is the return loss, measured in dB (as taken from the attenuator setting).

Example: Assume the step attenuator initial setting was 40 dB with six divisions of vertical signal displayed. After connecting the device under test to the DUT port, the step attenuator has to be adjusted to 10 dB of attenuation in order to get as close to six divisions of vertical deflection as possible. The difference between 40 dB and 10 dB is 30 dB. The return loss in this example is 30dB.

Accuracy Limitation

Since most low cost step attenuators have only 1dB of resolution, you may not be able top exactly match the initial divisions of the displayed signal. Thus, the accuracy is limited to about 1 dB. If greater accuracy is required, a spectrum analyzer and or precision RF power meter, or a very high accuracy oscilloscope may have to be employed.

Bridge Operation

By way of operation, the reference internal impedance (50-ohm) is compared to the DUT impedance. If the impedances are exactly equal, than the detector output will be essential zero (0). In practice this never happens. Most bridges have residual return loss from 30-40 dB (1000-1 to about 10,000-1).

The unknown impedance measured by this technique is not limited to amplifier inputs. Coax cables attached to a load, an antenna, a filter, or any other fixed impedance device can be characterized by return loss.

What is return loss

You may want to skip reading this section if all that interests you is the practical aspects of measuring return loss. Thanks to feedback from our customers, clarification as to the exact definition of return loss, particularly whether it should be expressed as a positive or negative quantity is useful.

I believe that return loss when expressed as a relative quantity, such as a dB, is a positive quantity, and when expressed as power as in dBm, it can be either positive or negative regardless whether or not it is measuring an active or passive device.

Notwithstanding the foregoing, there are a number of experts such as Dr. Trevor S. Bird, editor of the IEEE Antennas and Propagation Transactions, who have published articles addressing the science behind return loss measurements. Yet confusion still prevails. The internet provides a great deal of information, some of it questionable, so, for those interested, I suggest they research this issue on their own and draw their own conclusion.

Nonetheless, Wikipedia defines return loss as follows:

> *"..In telecommunications, return loss is the loss of signal power resulting from the reflection caused at a discontinuity in a transmission line or optical fiber. This discontinuity can be a mismatch with the terminating load or with a device inserted in the line. It is usually expressed as a ratio in decibels (dB);*

$$RL(\mathrm{dB}) = 10\log_{10}\frac{P_\mathrm{i}}{P_\mathrm{r}}$$

where RL(dB) is the return loss in dB, P_i is the incident power and P_r is the reflected power. Return loss is related to both standing wave ratio (SWR) and reflection coefficient (Γ). Increasing return loss corresponds to lower SWR. Return loss is a measure of how well devices or lines are matched. A match is good if the return loss is high. A high return loss is desirable and results in a lower insertion loss. Return loss is used in modern practice in preference to SWR because it has better resolution for small values of reflected wave.

Sign

Properly, loss quantities, when expressed in decibels, should be positive numbers. However, return loss has historically been expressed as a negative number, and this convention is still widely found in the literature. Taking the ratio of reflected to incident power results in a negative sign for return loss;

$$RL'(\mathrm{dB}) = 10\log_{10}\frac{P_\mathrm{r}}{P_\mathrm{i}}$$

where RL'(dB) is the negative of RL(dB). Return loss is identical to the magnitude of Γ when expressed in decibels but of opposite sign. That is, return loss with a negative sign is more properly called reflection coefficient. The S-parameter S_{11} from two-port network theory is frequently also called return loss, but is actually equal to Γ. Caution is required when discussing increasing or decreasing return loss since these terms strictly have the opposite meaning when return loss is defined as a negative quantity..."

See figure 3 below. Remember, a dB is a relative value and dBm are actual power levels. dBM can be either positive or negative. When using a return loss bridge, the detector port displays the reflected power from a device under test. When displaying this value on a spectrum analyzer, assuming the reference level is set to zero dBM, the reflected power from a device as measured from the detector port can very well be less than zero dBm. As a result, negative values may be shown if the output power is less than zero dBm. Remember, zero dBm is not zero power.

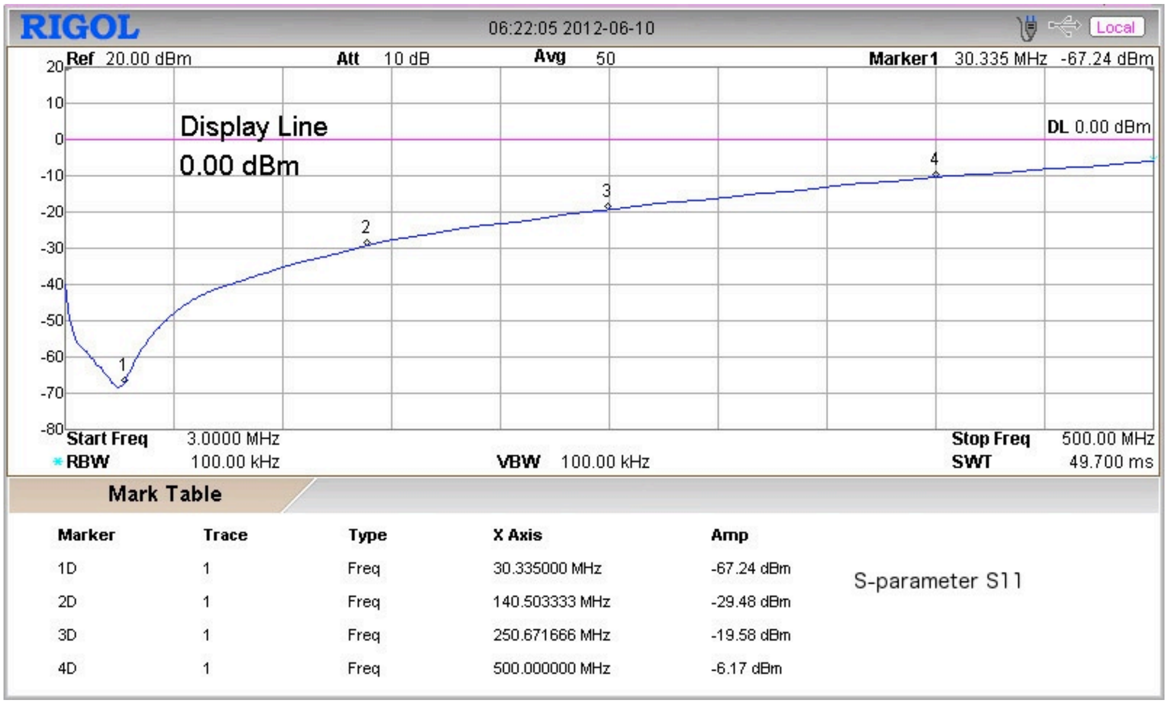

Figure 3. Reflected power expressed in dBm

You don’t have have to make these calculation manually. At the time of this writing there were a number of online SWR and Return loss calculators available free on the Internet. See below:

http://www.microwaves101.com/encyclopedia/calvswr.cfm

http://cgi.www.telestrian.co.uk/cgi-bin/www.telestrian.co.uk/vswr.pl

http://chemandy.com/calculators/return-loss-and-mismatch-calculator.htm

Return Loss versus VSWR

Return Loss(dB)	VSWR	Reflection Coefficient, Γ	Mismatch Loss (dB)	Reflected Power (%)	Forward Power(%)
1	17.39	0.891	6.868	79.43	20.57
2	8.72	0.794	4.329	63.10	36.90
3	5.85	0.708	3.021	50.12	49.88
4	4.42	0.631	2.205	39.81	60.19
5	3.57	0.562	1.651	31.62	68.38
6	3.01	0.501	1.256	25.12	74.88
7	2.61	0.447	0.967	19.95	80.05
8	2.32	0.398	0.749	15.85	84.15
9	2.10	0.355	0.584	12.59	87.41
10	1.92	0.316	0.458	10.00	90.00
11	1.78	0.282	0.359	7.94	92.06
12	1.67	0.251	0.283	6.31	93.69
13	1.58	0.224	0.223	5.01	94.99
14	1.50	0.200	0.176	3.98	96.02
15	1.43	0.178	0.140	3.16	96.84
16	1.38	0.158	0.110	2.51	97.49
17	1.33	0.141	0.088	2.00	98.00
18	1.29	0.126	0.069	1.58	98.42
19	1.25	0.112	0.055	1.26	98.74
20	1.22	0.100	0.044	1.00	99.00
21	1.20	0.089	0.035	0.79	99.21
22	1.17	0.079	0.027	0.63	99.37
23	1.15	0.071	0.022	0.50	99.50
24	1.13	0.063	0.017	0.40	99.60
25	1.12	0.056	0.014	0.32	99.68
26	1.11	0.050	0.011	0.25	99.75
27	1.09	0.045	0.009	0.20	99.80
28	1.08	0.040	0.007	0.16	99.84
29	1.07	0.035	0.005	0.13	99.87
30	1.07	0.032	0.004	0.10	99.90
31	1.06	0.028	0.003	0.08	99.92
32	1.05	0.025	0.003	0.06	99.94
33	1.05	0.022	0.002	0.05	99.95
34	1.04	0.020	0.002	0.04	99.96
35	1.04	0.018	0.001	0.03	99.97
36	1.03	0.016	0.001	0.03	99.97
37	1.03	0.014	0.001	0.02	99.98
38	1.03	0.013	0.001	0.02	99.98
39	1.02	0.011	0.001	0.01	99.99
40	1.02	0.010	0.000	0.01	99.99

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Application Note #6

Ham Radio Station Monitor

Performance and Selection

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The SMT Station Monitor and SMT-Pro Station Monitor are ideal for monitoring the performance of the entire transmitter chain for both AM and SSB operations. Selecting which SMT Station monitor depends on your measurements needs. For basic monitoring of high power RF (QRO) only, than the SMT Station Monitor is recommended. If you want to monitor both high power (CRO) and low power (QRP) with higher precision than the SMT-Pro is recommend. This discussion will give you some insight to help you make a more informed selection.

One of the best methods to monitor your station is by observing the demodulated RF being transmitted and or comparing the amplified RF to the un-amplified RF from the transceiver. This can be done with a trapezoid monitor and or two tone tests.

The station monitor consists of a wide band sampler, a high performance demodulator, a variable base band output and an oscilloscope trigger output, A linear RF amplifier generally amplifies an RF signal from .5 – 5 Watts by 30dB or more to about 500-1,500 Watts. Its performance and modulation can be characterized using a spectrum analyzer (expensive) or a low cost oscilloscope using a trapezoid display.

This is done by sampling the amplifier's output by using an RF sensor such as a sampler or coupler. This sensor is connected to the oscilloscope's vertical (Y) input.

The input of the amplifier is driven by a transceiver which usually outputs less than 100W. Its output drives the input to the amplifier and also a wide band demodulator which extracts the baseband from the modulated RF.

It is this baseband that is connected to the oscilloscope's horizontal (X) input. This display yields a trapezoid pattern. This pattern compares the transceiver's output to the amplifiers output. If the amplifier is linear without any distortion and not overdriven, the trapezoid pattern will be a linear undistorted triangular waveform see Fig1. To rely on such a measurement, the demodulator and signal samplers must be linear and free of distortion.

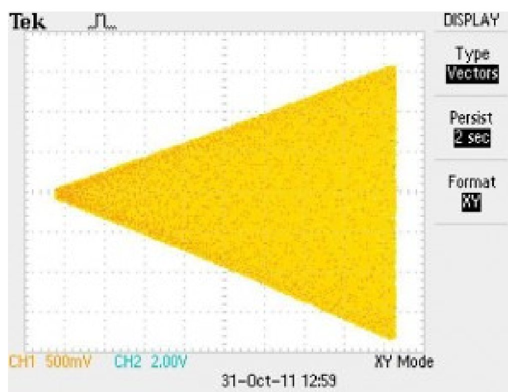


Fig. 1 Trapezoid test pattern

Most passive wide band demodulators employ an envelope detector. This detector is an electronic circuit that takes a high-frequency signal as input and provides an output which is the "envelope" of the original signal. Generally a diode rectifies the incoming signal, allowing current flow in only one direction.

Most practical envelope detectors use either half-wave or full-wave rectification of the signal to convert the AC audio input into a pulsed DC signals. Filtering is then used to smooth the final result. This filtering is rarely perfect and some "ripple" is likely to remain on the envelope follower output, particularly for low frequencies. More filtering gives a smoother result, but decreases the responsiveness; thus, real-world designs must be optimized for the application.

Low level detection

Another undesirable artifact is non-linearity of the detected baseband. This is caused by the diode's conduction voltage drop ranging from .2 volt to .8 volt depending on the diode type and current. The diode does not linearly detect in this diminished conduction region. For high level RF envelopes this small region usually represents only a small percentage of the envelope and can be ignored.

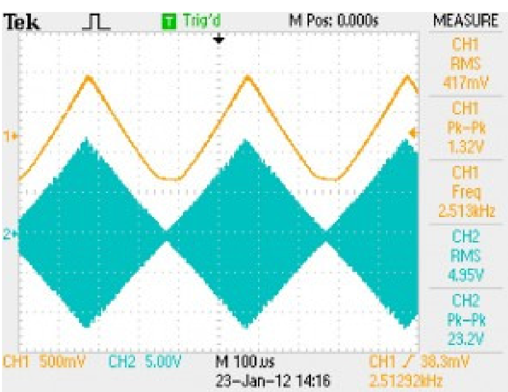


Fig. 2 Demodulator clipping no bias

See Fig 2. However, in low level RF envelopes encountered with QRP operation, this region represents a considerable portion of the modulation envelope. As a result, the baseband will exhibit significant non-linearity near the maximum modulation depth of the carrier

Unique diode bias circuit

One way to mitigate this undesirable phenomena is to bias the diode such that it is conducting throughout the 100% modulation envelope. The SMT-Pro Station Monitor incorporates such a bias currents source. See Fig 3. This provides exceptional baseband linearity over a wide input range for precise transmitter amplifier linearity measurements.

This discussion focuses on the trapezoid test technique. The RF sampler or RF coupler and demodulator performance can impact the quality of the test

Note the clipping at low modulation levels (upper trace) especially evident during the modulation trough. This phenomenon is caused by the inherent lack of the detector level bias current.

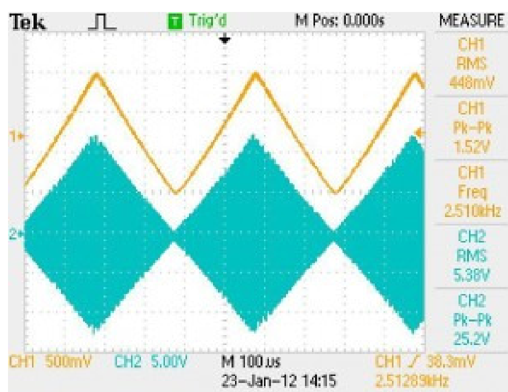


Fig. 3 Modulation envelope with bias

See Fig. 3 Note the demodulated product (upper waveform) clearly shows the improvement at the trough modulation level resulting from the addition of the detector level bias current option.

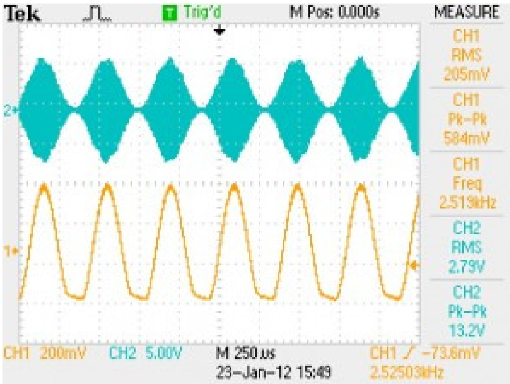


Fig. 4 Demodulated sine wave without bias

Fig. 4 Note the RF envelope (top waveform), detected product (bottom waveform) without bias current option. This is caused by the intrinsic lack of detector bias current at the modulation trough.

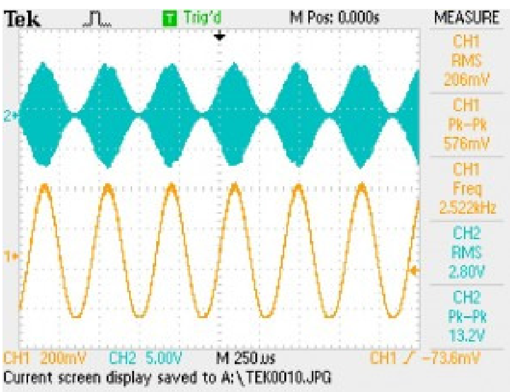


Fig. 5 Sine wave modulation with bias

Fig. 5 Note the lack of clipping at low modulation levels is especially evident during the modulation trough. This is the result of adding low level detector bias current available on the SMT-Pro option.

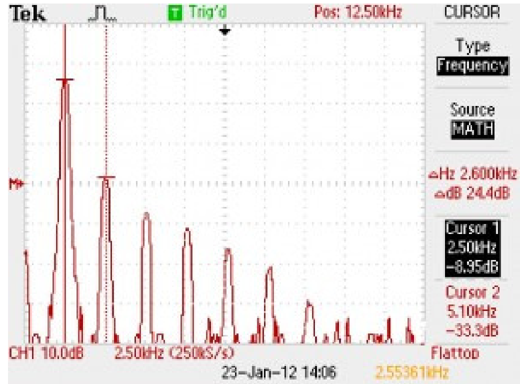


Fig. 6 THD without bias

Fig. 6 Note the significant harmonic distortion (-24.4 db) levels resulting from modulation through clipping without the detector bias option. This is especially noticeable at low power level such QRP operation. At higher levels this effect is significantly reduced.

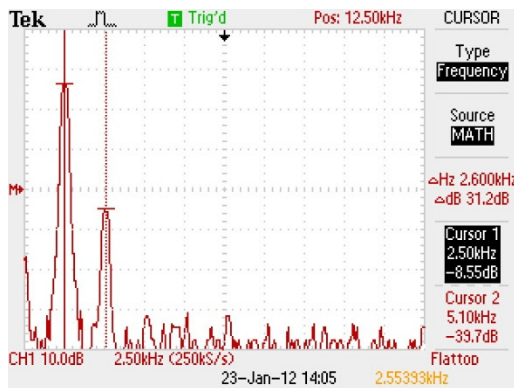


Fig. 7 THD with bias

Fig. 7 Note the significantly reduced harmonic distortion levels as a result of eliminating modulation through clipping with the detector bias option. This is especially noticeable at low power level such QRP operation. At higher levels this effect is significantly reduced. The second harmonic distortion is down -31.2 dB. At higher modulation levels of > 5watts, second order harmonic distortion is -45 dB. All other modulation products are virtually eliminated.

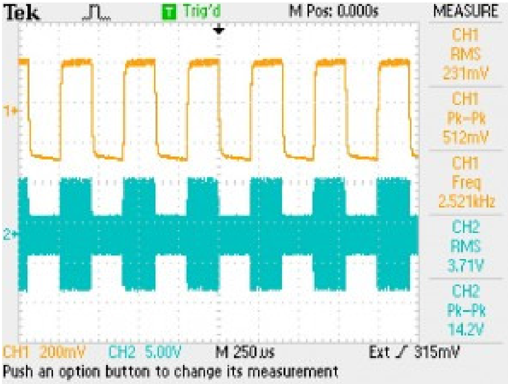


Fig. 8 Pulse response

Fig. 8 This graph shows the modulated RF envelope (lower trace) along with the demodulated signal (upper trace). Note that the transition times easily meet the bandwidth specifications (10-30,000 Hz) there is no spurious distortion ringing, overshoot present.

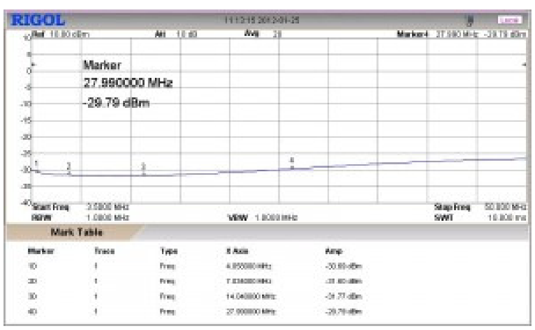


Fig. 9 Sampler bandwidth

Fig. 9 The nominal sampler output equals -30dB of the RF being sampled. This is a power ratio reduction of 1000:1. This also equals a voltage ratio reduction of 31.623. Decibels state a power ratio, not an amount. They tell how many times more (positive dB) or less (negative dB) a signal is but not how much more or less in absolute terms.

Decibels are logarithmic, not linear. For example, 20 dB is not twice the power ratio of 10 dB. Use this equation to find decibels: A = 10*log10 (P2/P1) (dB) where P1 is the power being measured, and P1 is the reference to which P2 is being compared. To convert from decibel measure back to power ratio: P2/P1=10^(A/10). Voltage is more easily measured than power, making it generally more convenient to use: A = 20*log10(V2/V1). Where A=voltage ratio. The equation for obtaining voltage ratio from dB is V2/V1 = 10^(A/20). Thus, to obtain the equivalent voltage (Peak, P-P, or RMS) for -30dB multiply the sampled voltage times 31.63. For example if the RMS voltage at the sampler = 5 volts, than the actual RF RMS voltage at the sampler input would be 5×31.63=158.15. The power (P=E^2/R) would be 500.2 Watts.

Fig. 10 The wideband graph of the sampler output is shown here. While the nominal sampler output equals is -30dB of the RF being sampled.

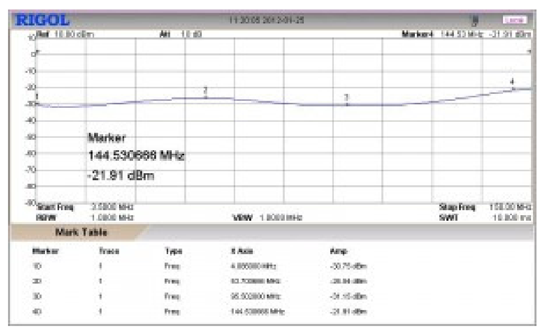


Fig. 10 2-150 MHZ bandwidth

Here at 4MHz it is -30.75dB, and 21.91 dB at 144.53 MHz Again, the equation for obtaining voltage ratio from dB is V2/V1 = 10^(A/20). Thus, -21 dB equals a voltage-ratio of 11.22 and the power ratio equals 125.89. To obtain the equivalent voltage (Peak, P-P, or RMS) for -21 dB, multiply the sampled voltage times 11.22. For example, if the RMS voltage at the sampler = 5 volts, than the actual RF RMS voltage at the sampler input would be 5×11.22=56.1. The power (P=E^2/R) would be 62.9 Watts.

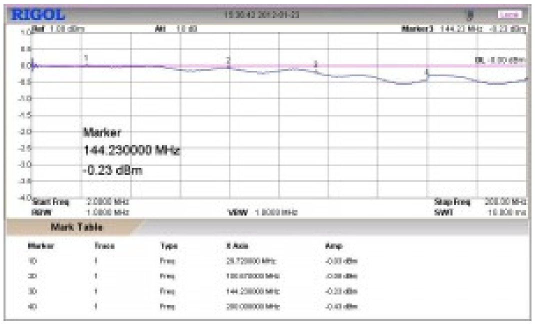


Fig 11 Insertion loss with bias

Fig. 11 Station monitor insertion loss without detector bias current option is show here. The insertion loss (-.03dB at HF frequencies) is barely measurable. This is the current consumed by the detector only (a few micro amps). For practical purposes this loss can be ignored since it equals a power radio of 1.00693:1, so again, virtually all the power passing through the sampler is delivered to the load.

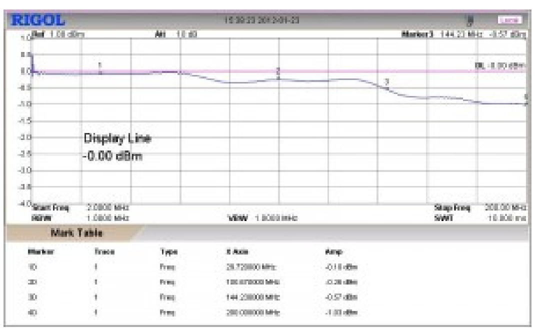


Fig. 12 Insertion loss with bias

Fig. 12 Station monitor insertion loss with detector bias current option is show here. The insertion loss (-.25dB at HF and 1dB at 200MHz) is barely measurable. This is the current consumed by the detector and its biasing current supply (a few micro amps). For practical purposes this loss can be ignored since it equals a power radio of 1.059:1. Here, virtually all the power passing through the sampler is delivered to the load.

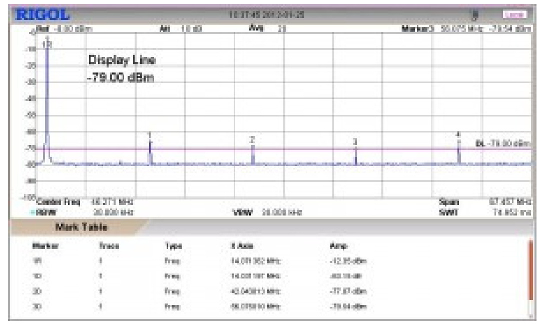
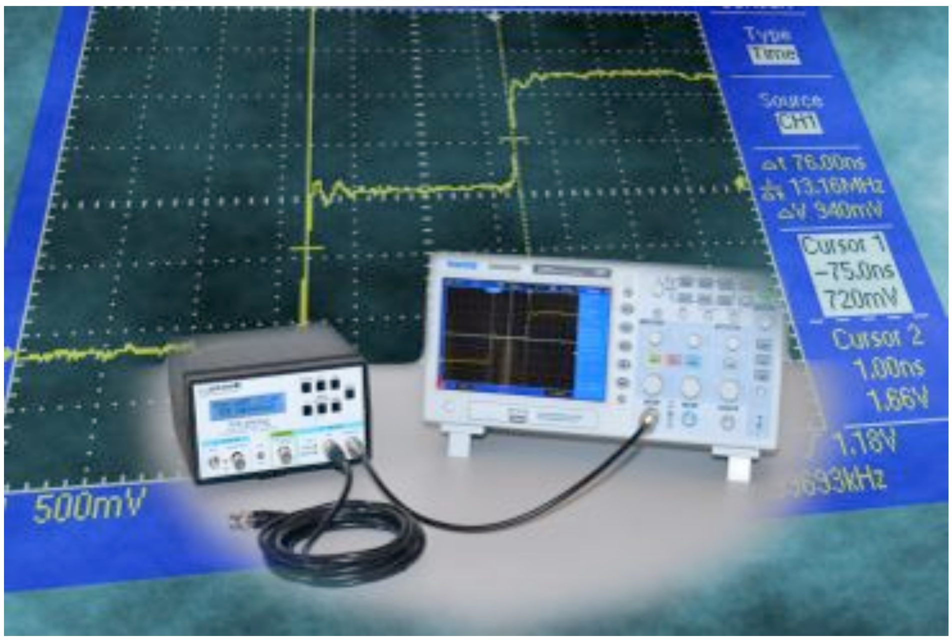


Fig. 13 Spurious emissions

Fig. 13 The graph of spurious emission of a typical linear amplifier (Ameritron AL811H) passing through the SMT & SMT-Pro station monitor. Note, the worst case harmonic spurious emission is -63.15 dB from the fundamental frequency of 14MHz. This is the residual spurious emission of the AL811H amplifier. Remarkably, when not overdriven, the AL811H despite its ancient design, exhibits excellent spurious emission characteristics. The additional spurious emissions contributed by the station monitor measurements were negligible. It should be noted that the power of these spurious emission are at least 100 times (20dB) better than the minimum allowed by the FCC*: §97.307 Emission standards (d) For transmitters installed after January 1, 2003, the mean power of any spurious emission from a station transmitter or external RF amplifier transmitting on a frequency below 30 MHz must be at least 43 dB below the mean power of the fundamental emission.*

Most, if not, all of the foregoing oscilloscope measurements can be made with an inexpensive 30MHz oscilloscope. By using appropriate accessories and techniques, the ham radio operator can maximize the RF transmitted signal performance.

Using TDR for Measuring Transmission Lines in Ham Radio Installations



This application note reviews the elements of transmission line measurement in the ham radio environment. It demonstrates how you can measure line impedance, return loss, SWR, velocity factor, distance to fault and line losses using pulse interrogation techniques. It focuses on the new preciseRF TDR-CableScout® pulse generator as a companion accessory to an oscilloscope.

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Precision Ham Radio Measurements

1. Transmission Lines

Virtually all ham radio installations require some type of transmission line. The purpose of a transmission line is to efficiently transfer RF energy from the transmitter to the transmitting antenna, or conversely, efficiently transfer RF energy from the receiving antenna to the receiver.

Ideal Transmission Line

The ideal transmission line matches the transmitter and antenna impedance precisely and delivers all input energy without losses. This usually occurs when all the power is transmitted without reflections and or resistive losses. Real-world transmission lines always have losses because either the source impedance (Zs) of the transmitter, the load impedance (Z_L) of the device receiving the energy (usually the antenna) or the line impedance (Zo) of the transmission lines are not matched.

Line Losses

There are other losses such dielectric, resistive and reactive losses which affect the performance of the transmission line. Many times these losses are due to manufacturing defects, poor quality, inferior connectors, environmental damage to the line such as UV radiation, moisture, physically kinked or broken cables, miss-matched cable types or excessive cable length.

The goal is to insure that the transmission line meets the expected performance requirements. In the ham radio applications, overall transmission lines performance is usually measured with SWR meters, return loss bridges or RF samplers with station monitors. These techniques work well and are low cost. They provide an overall check of the transmitter, transmission line and antenna (or resistive load) quality. Their main disadvantage is that they do not provide information as to where the fault is located, nor do they indicate the fault type, such as defective cables loose, connectors or other problems. Fortunately, time domain reflectometry (TDR) provides more information. Wikipedia defines a transmission line as:

> *"In communications and electronic engineering, a transmission line is a specialized cable designed to carry alternating current of radio frequency, that is, currents with a frequency high enough that their wave nature must be taken into account. Transmission lines are used for purposes such as connecting radio transmitters and receivers with their antennas, distributing cable signals, and computer network connections."*

2. TDR Basics

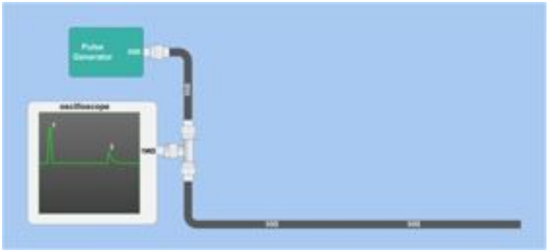


Figure 1. Simple TDR set-up

See figure 1. TDR principles are fairly easy to master. Think of it as cable radar. A pulse generator is connected via a “T” connection to an oscilloscope’s high impedance input and a pulse or step is injected (incident pulse) into the cable. The pulse and any reflections are then displayed on the oscilloscope for analysis.

Properly Terminated Line

See figure 2. If the conductor is of a uniform impedance and is properly terminated, the entire transmitted pulse will be absorbed in the far-end termination and no signal will be reflected toward the TDR. Any impedance discontinuities will cause some of the incident signal to be sent back towards the source. This is similar in principle to radar.

Figure 2. Terminated line

Since the pulse in transmission lines travels at a certain speed (.66 to .90 times the speed of light) depending on the cable type, it is possible to locate the reflection (fault) by measuring the round trip time and thus, locate the distance to the fault.

Open Line

See figure 3. The scope displays the incident pulse and any reflections. With an open

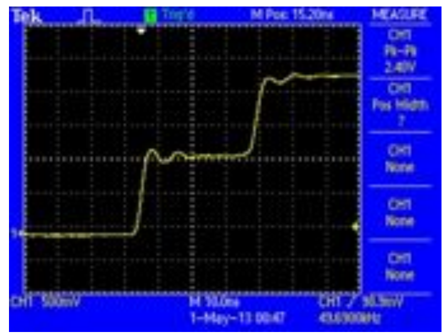


Figure 3. Open cable

cable for example, which is a very high impedance, increases in the impedance create a reflection that reinforces the original pulse.

Shorted Line

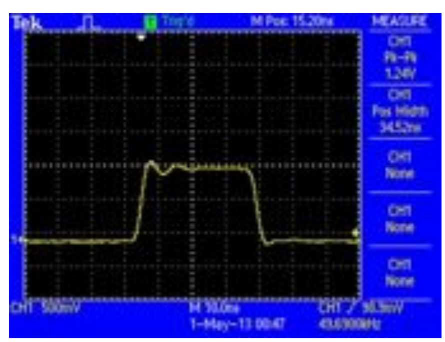


Figure 4. Shorted cable

See figure 4. A shorted cable, for example, has very low impedance, it creates a reflection that opposes the original pulse.

TDR Sensitivity

Because of TDR's sensitivity to impedance variations, it may be used to verify cable impedance characteristics, splice and connector locations, associated losses, and measure cable lengths.

Figure 5. Reflections

See figure 5. By analyzing the pulse amplitude, shape, and time, one can analyze the likely cause of the fault. In this example, the first reflection is negative going, indicating a decrease in impedance (most likely caused by a kinked cable). The second reflection is positive going, indicating an increase in impedance (most likely caused by a defective connector or braided shield failure).

Incident Pulse

See figure 6. The incident pulse is the pulse applied to the device under test (DUT). The amplitude is measured from the most negative level (generally ground) to the most positive level (excluding any aberrations not caused by reflections from the DUT).

Figure 6. The incident pulse

Reflected Pulse

The reflected signal, also called the reflected pulse, contains signals which are made of reflections caused by the line impedance (Zo) not matching the pulse generator source impedance (Zs). The amplitude is measured as a deviation from the most positive level of the incident pulse. This level can be either positive or negative.

See figure 7. This multiple exposure waveform shows the incident pulse (the first half of the screen), and the reflected signal of positive and negative values.

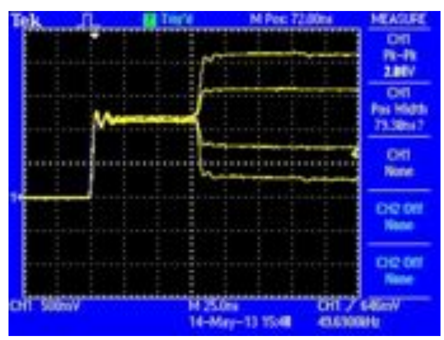


Figure 7. Multiple exposure of Zo variations

3. TDR Equipment

The commercial communications industry has long adopted TDR techniques to analyze transmission lines. A TDR measurement of a transmission line provides precise quantitative data of the line performance and identifies any faults.

Specifically, TDR measurements provide information such as distance to fault (DTF), reflection coefficient (p), transmission line impedance (Zo), return loss (RL), voltage standing wave ratio (VSWR), line length, line velocity factor (Vf), cable dielectric and resistive losses at specific frequency and cable length. While the measurement capability is impressive, the equipment costs are high and generally beyond the reach of ham radio operators.

Up until now, ham radio operators wanting to make TDR measurements either had to spend considerable money on a commercial TDR oscilloscope and pulse generator with integrated samplers and TDR computers, or compromise and use an ordinary oscilloscope and pulse generator and accept the limitations provided by this solution.

Equipment Choices

The electronic practitioner who wishes to make TDR transmission line measurements basically has these options:

1. If you have the money, buy a new TDR system such as the Mohr CT100, a Tektronix TDR scope or Angilent TDR scope. Starting at $18,000, these solutions are expensive, but they have the latest software and work well and have factory support.
2. Buy a used TDR system from eBay. Good price, but they may be difficult to get calibrated or serviced.
3. Compromise and use your scope, pulse generator and your trusty calculator. This works pretty well depending on your pulse generator performance. This choice is low cost, uses “T” connection, but has no low impedance scope input capability (needed for speed). All calculations must be done manually.

4. The TDR-CableScout ®

Recognizing the cost and performance limitations of the available choices, we created the *TDR-CableScout* ®. It was designed to be affordable, yet provide laboratory level accuracy and utility. *TDR-CableScout* ® takes advantage of the fact that low cost, high performance oscilloscopes are available from many sources. When used with an oscilloscope of sufficient bandwidth, measurements can be made rivaling those of commercial TDR systems at a fraction of the cost.

The *TDR-CableScout* ® includes a TDR computer because these scopes, while high enough in bandwidth (about 200MHz) do not have a TDR computer. This requires the user to make all the calculations manually. While not difficult to do, they are nevertheless tedious. Conventional pulse generators do not have the very fast rise time, selectable Zs and duration rates best suited for TDR work.

TDR for Ham Radio

The *TDR-CableScout* ® gives hams the means to analyze transmission lines and circuit board strip lines. Line impedance from 50 Ω to 600 Ω can be measured with 25 ps resolution. It

features isolated high speed samplers and separate device under test (DUT) outputs. This design allows a direct, fast Tr 50 Ω connection to the oscilloscope for accurate TDR measurements without the inconvenience and lower performance that the “T” connector solution offers.

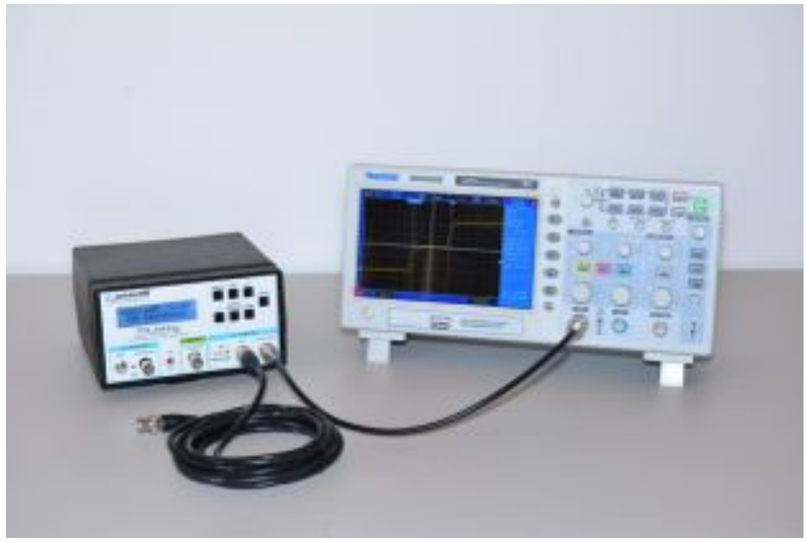
A transmission line library is included. It contains data for velocity factor (Vf), line impedance (Zo) and line loss data. The integrated TDR measurement computer takes the work out of TDR measurements, such as time to fault (TTF), reflection coefficient (p), cable length, velocity factor (V_F), line impedance (Zo), return loss (RL), SWR and cable loss.

The *TDR-CableScout* ® features both pulse and step TDR. The step TDR has a maximum range of 15 KM and time resolution of better than 1 ns. The pulse TDR features a ≤ 400 ps pulse width and ≤ 150 ps Tr.

The resolution is under 5mm, which is well suited for analyzing circuit board strip lines. A dedicated trigger output features a 100 ns pre-trigger to allow viewing of the TDR pulse leading edge when using sampling scopes without a delay line such as the 7S11 and 7T11 installed in legacy Tektronix 7000 scopes.

TDR-CableScout ® Method

The TDR measurement set-up consists of the *TDR-CableScout* ® pulse generator and an oscilloscope. The pulse generator provides all required TDR pulses with the proper amplitude and transition time (Tr) and source impedance (Zs). The scope displays the resultant TDR waveform and provides a means of measuring time and amplitude of these pulses.



The *TDR-CableScout* ® *with scope option*

The user inputs voltage and time values observed on the scope, and the pulse generator computer provides measurement results. The scope option shown above includes a 200 MHz Hantek DSO oscilloscope especially selected for TDR measurements.

Rise Time

The scope should have a calibrated vertical amplifier and calibrated time base. See figure 8. Three bandwidths displays created with a bandwidth limiter are shown. The scope bandwidth should be sufficiently high to identify reflections at the resolution needed for the application.

For short line distances and circuit board TDR, finer time resolution is required. Resolution is a function of the scope's rise time (Tr). Bandwidth is directly related to Tr and the commonly accepted mathematical relationship is BW=.35/Tr. While some users have used scopes with bandwidth as low 20 MHz, 100-200 MHz bandwidth scopes will work for most ham radio applications.

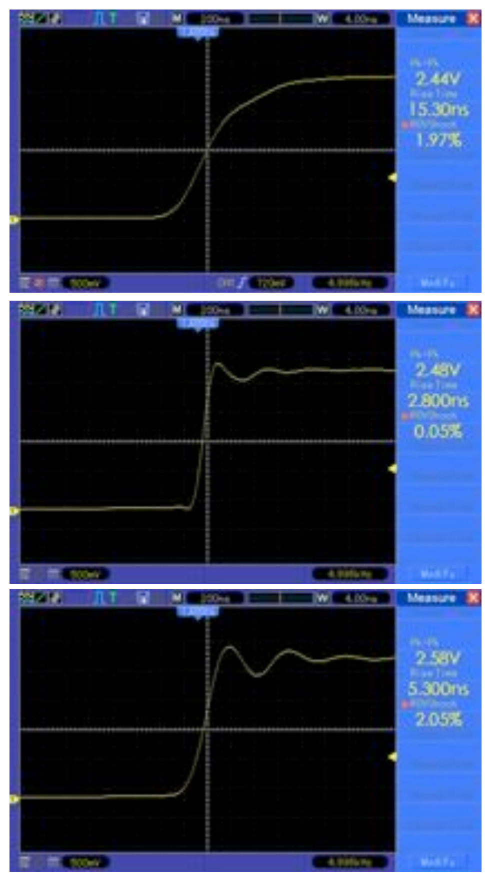


Figure 8. Tr displays of 20, 100 and 200 MHz BW scopes

Effects of Rise Time on TDR Resolution

The scope's rise time has a significant effect on TDR measurements are shown in figure 9. The example is a TDR measurements of a 12 foot piece of RG 58 coax with a published velocity factor (Vf) of .66 with the end open.

It demonstrates that this is about the shortest cable that can be tested with a 20 MHz scope. Note that impedance variations are clearly visible in the 100 MHz and 200 MHz scope.

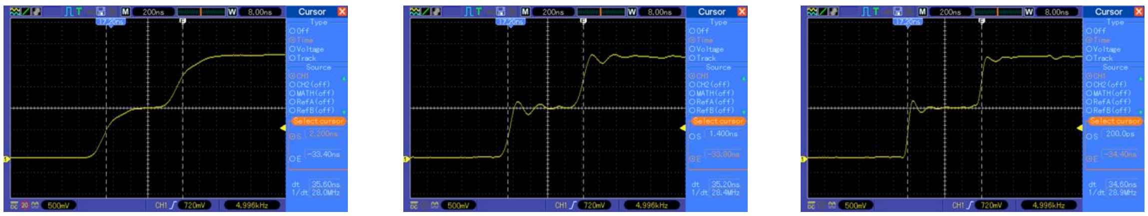


Figure 9. Scope rise time and bandwidth effects on TDR Measurement resolution

5. *TDR-CableScout* ® Controls

See figure 10. Here is a depiction of the *TDR-CableScout* ® pulse generator front panel and controls. Black is used for labels, blue is for outputs and green is for trigger functions. Green LEDs indicate the state of the current TDR mode and output impedance. The controls are grouped into five sections, see tables 1 and 2.

Figure 10. The *TDR-CableScout* ® *front panel*

Button	Table 1. Key Description
CABLE/Vf	Provides a cable selection from the library. Each cable includes the nomenclature, cable loss per 100 feet at 100 MHz, cable impedance (Zo) and velocity factor (Vf). A selection of a cable is used as a preset value for calculations performed by the TDR computer.
Zo	Sets the source impedance Zs and step or pulse TDR selection. Zs impedance is 50 Ω, 75 Ω and 300 Ω suitable for line impedance measurements ranging from 50 Ω to 600 Ω.
RANGE/ DURATIO N	Selects TDR range and duration ranging from 15 KM (100 us pulse width) to 75 Meter (500 ns pulse duration) in four ranges. Maximum resolution is 10 ps (using pulse TDR).
CALC	Selects calculation of distance to fault (DTF), cable length, velocity factor (Vf), reflection coefficient (p), return loss (RL), SWR, and line loss per 100 ft at 100 MHz.
UP	Use the UP and DOWN key to scroll increasing or decreasing numerical data or selectable items.
DOWN	Use the UP and DOWN key scroll increasing or decreasing numerical data or selectable items. Also used to display additional calculation results.
ENTER	Completes a data entry or sub calculation or function.

Table 2 Displayed Prompt Convention	
	Whenever prompted by the return symbol press the **ENTER** key to advance.
	Whenever prompted by the down arrow, press the **DOWN** key for additional measurement results.

Display

The display is a high contrast two line sixteen digit backlit LCD display. See figure 11. It provides the state of the current TDR function and shows various input and output conditions. It also indicates TDR measurement results.

Figure 11. LCD display

Step TDR Output

See figure 12. The Step TDR output consists of the DUT output and the scope output. The pulse waveform duration and frequency is settable from 5 KHz to 1 MHz. They are accurately set by a crystal oscillator timebase.

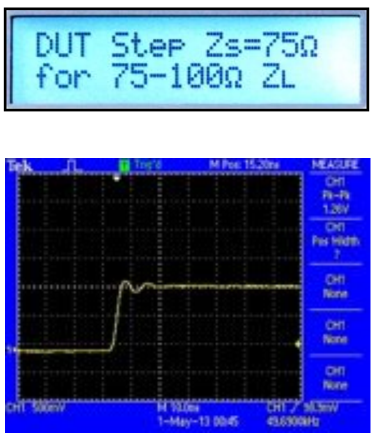


Figure 12. Step TDR

This allows for TDR range measurements from a few centimeters to 15 KM. Step TDR is most useful for measuring reflection coefficient, cable impedance (Zo), return loss (RL), voltage standing wave ratio (VSWR) and distance to fault. The source impedance (Zs) is calibrated and selectable from 50 Ω, 75 Ω and 300 Ω, using high speed SMA microwave relays set by the on-board microprocessor.

Pulse TDR Output

See figure 13. The DUT pulse output. The very fast pulse TDR is used to measure circuit board traces and line losses.

The SMA outputs a pulse to the device under test (DUT) with a 400 ps pulse width, a Tr of ≤ 150 ps and an amplitude of 2 V Pk-Pk. When PULSE TDR is enabled the green LED next to the DUT connector illuminates. The maximum range is 1.5 KM, pulse width is 10 us and the frequency is 50 KHz.

When using pulse TDR, the SMA connector is intended for the DUT and the BNC connector for the scope's vertical input. To preserve the high frequency detail, the scope vertical should be terminated into 50 Ω.

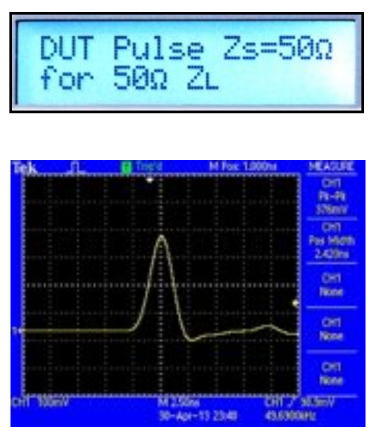


Figure 13. Pulse TDR

6. TDR Concepts and Terms

There are a number of primary TDR measurements concepts terms. These are the reflection coefficient (ρ), distance to fault (DTF), velocity factor (Vf) and line loss or cable loss. Some of these measurements also provide return loss (RL), line impedance (Z_0), VSWR and other parameters. What follows are the mathematical relationships governing these measurements:

The Reflection Coefficient

TDR measurements are based on a series of impedance ratios. TDR measurements are described in terms of a reflection coefficient, ρ (rho). The coefficient ρ is the ratio of the reflected pulse amplitude to the incident pulse amplitude:

$$\rho = V\ \text{reflected} \div V\ \text{incident}$$

For a fixed termination Z_L, ρ can also be expressed in terms of the transmission line characteristic impedance, Z_O and the load impedance Z_L.

$$\rho = V\ \text{reflected} \div V\ \text{incident}$$
$$= (Z_L - Z_0) \div (Z_L + Z_0)$$

Representing a matched load, a short circuit and an open load, ρ has a range of values from +1 to —1, with 0 representing a matched load. When ZL is equal to Zo, the load is matched. V reflected, the reflected wave, is equal to 0 and ρ is 0. There are no reflections:

$$\rho = V\ \text{reflected} \div V\ \text{incident}$$
$$= 0 \div V = 0$$

A Z_L reading of zero (0) implies a short circuit. The reflected wave is equal to the incident wave, but opposite in polarity. As seen below, the reflected wave negates part of the incident wave. The ρ value is -1.

$$\rho = V\ \text{reflected} \div V\ \text{incident}$$
$$= -V \div V = -1$$

When Z_L is infinite, an open circuit is implied. The reflected wave is equal to the incident wave and of the same polarity. The reflected wave reinforces part of the incident wave. The ρ value is +1.

Transmission Line and Load Impedance

The characteristic impedance Z_0, or the load impedance Z_0 can be calculated with the value of ρ:

$$Z_L = Z_0 * (1+\rho)\div(1-\rho)$$

Return Loss of the Transmission Line

The return loss (RL) of a transmission line is a conversion of the reflection coefficient (ρ) to dB. Return loss is expressed as a positive number and can be calculated by the equation as follows:

$$RL = -20\ \log 10\ (\rho)$$

VSWR of the Transmission Line

The voltage standing wave ratio (VSWR) represents the ratio between the maximum and minimum amplitude of the standing wave. VSWR can be calculated by the equation as follows:

$$VSWR = (V\ max \div V\ min)$$
$$= 1+\ \rho \div 1 -\ \rho$$

Cable Losses of the Transmission Line

Cable losses in the ham radio installation are caused by several factors. While both conductor loss and dielectric loss occur, conductor loss usually dominates. Conductor loss is caused by the finite resistance of the metal conductors in the cable which, due to the skin effect, increases with frequency. The result of this incremental series resistance is an apparent increase in impedance as you look further into the cable. So, with long test cables, the DUT impedance looks higher than it actually is.

The second problem is that the rise time and settling of the incident pulse is degraded by the time it reaches the end of the cable. This affects resolution and accuracy since the effective amplitude of the incident step is different than expected. This amplitude inaccuracy does not cause much error when the DUT impedance is close to 50 Ω, but for a larger or smaller impedance, the error can be significant.

Loss per unit of length is generally provided by the manufacturers. For example, RG 58 might be specified as 4.1 dB/100 feet. Given a constant amplitude sine wave generator and a known length of transmission line, one can measure the actual loss per unit of length and compare that to a specified cable length usually given as dB/100 feet.

Gaussian Pulse Loss Measurements

One can make loss measurements using a TDR pulse generator with a fast enough pulse Tr output by comparing the incident pulse amplitude to the reflected pulse amplitude.

Losses in a transmission line due to changes in frequency are proportional to the square root of the frequency. Thus, if the attenuation per unit length is known for a particular frequency f1, the loss of any other frequency f2 can be calculated from the following equation:

$$af2 = af1\left(\sqrt{f2 \div f1}\right)$$

where a is loss in dB

Since the *TDR-CableScout®* makes loss measurements using a very fast gaussian pulse, we can apply the above equation using a 200 MHz scope for our measurements.

See figure 14. The *TDR-CableScout®* generates a pulse with a Tr of ≤ 150 ps and a pulse width of ≤ 400 ps. When the pulse is displayed on a Tektronix 6 GHz DSO, the incident pulse Tr of ≤ 150 ps Tr is clearly evident.

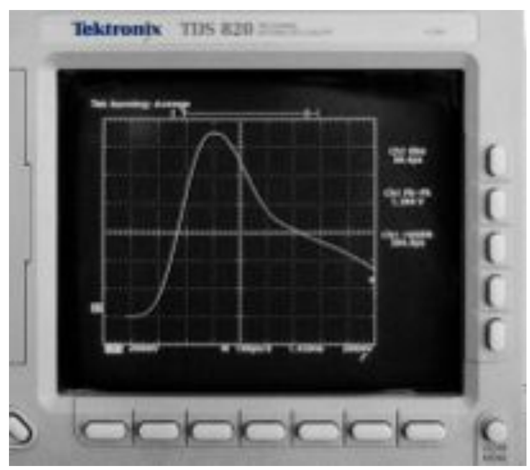


Figure 14. Gaussian pulse displayed on 6 GHz DSO

Given the relationship of BW=.35/Tr, we see that .35 ÷ 150ps = 2.3 GHz. The maximum FFT frequency is about 2.3 GHz.

See figure 15. When the same pulse is viewed on a 200 MHz scope, the displayed pulse Tr and pulse width will be stretched and the amplitude is decreased. However, the *generated* Tr, width and amplitude are unchanged, it's just that the scopes lower BW limit can't display them.

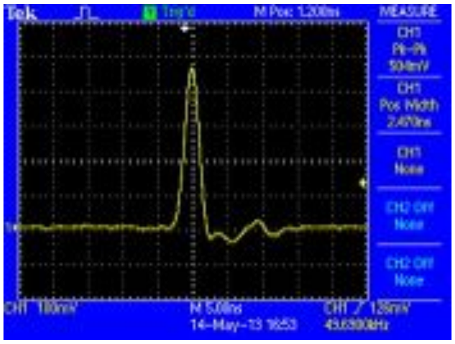


Figure 15. Gaussian pulse displayed on 200 MHz DSO

According to the ARRL on-line calculator, the Tandy RG 58 coax is specified to have a cable loss of 4.068 dB/100 feet, and at 2.2 GHz the calculated loss is 26.66 dB /100 feet.

Using a Rigol DSA1030 laboratory spectrum analyzer, we measured an actual cable loss of 5.2 dB/100 feet. At 2.2 GHz the measured loss was 21.3 dB /100 feet. The calculated results are pretty close when compared to real-world cables measurements.

Fundamentally, the *TDR-CableScout* ® makes the loss measurement at 2.2 GHz and then calculates the loss at 100 MHz, using the equation previously discussed:

$$af2 = af1 \left(\sqrt{f2 \div f1}\right)$$

where a is loss in dB

Summary

Understanding the frequency contents of gaussian pulses allows for measurements of cable losses using pulses with equivalent results as a constant amplitude sine wave generator measurement. This holds true as long as the measurement scope's bandwidth is ≥ 2 times the measurement parameter (loss @ 100 MHz). So, a 200 MHz scope will give reliable results.

Not all secondary factors affecting cable loss are taken into account using the gaussian pulse cable loss measurement method. For this reason, when making cable loss measurements a normalized cable length gives the most accurate result. Sample cable lengths of 25-50 feet give the best accuracy. However, for comparative cable loss testing, cables with identical length of just a few feet can be tested and high measurement certainty can be achieved.

7. Step-by-step TDR Measurements

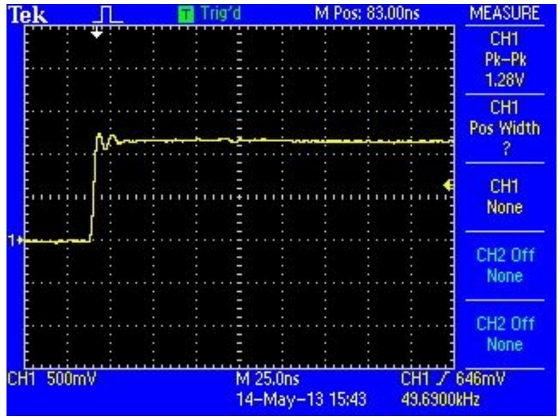
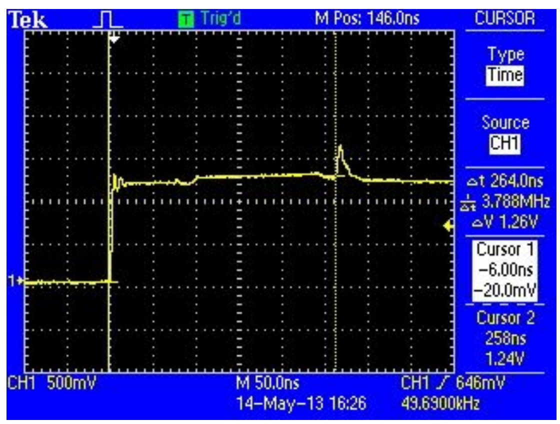
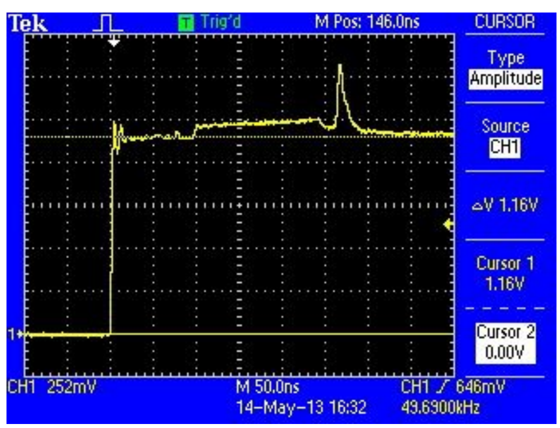
This section explains how to make TDR measurements using the TDR-CableScout ® measurement computer (accessible by using the **CALC** key). In each example you will be shown how to connect the scope and the DUT to the TDR-CableScout ®. Examples of the TDR-CableScout ® display and scope display are shown. Since the DUT transmission lines and cables you will be using are most likely different than the examples shown, your measurement results will reflect the actual cables you will be testing. The following examples are covered:

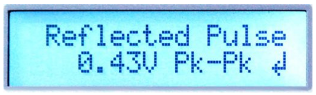
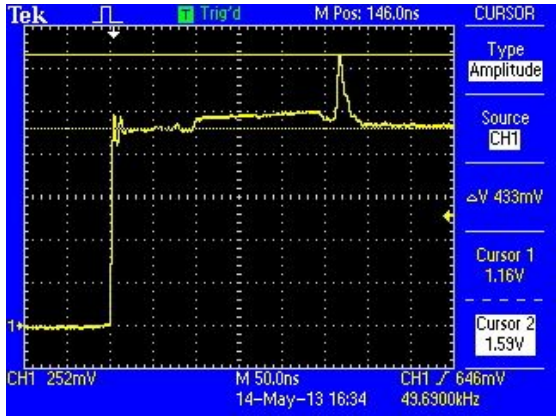
1. Reflection Coefficient, Return Loss, SWR and Zo Measurements.
2. Distanced to Fault (DTF) Measurements.
3. Velocity Factor Measurements.
4. Line Loss Measurements.

Reflection Coefficient, Return Loss, SWR and Zo Measurements

Hams worry about VSWR, and for good reason. Most transmitters do not tolerate a mismatched load with a VSWR > 2:1, and power that is supposed to go to the antenna is lost as heat. As previously discussed, VSWR is directly related to the reflection coefficient (ρ), return loss (RL), and line impedance (Zo). Using TDR to measure ρ, we get all four parameters in one measurement. See the step-by-step instructions below:

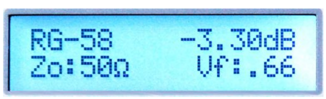
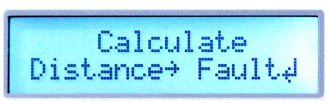
Reflection Coefficient Measurements Steps	Scope Display
 1. Using the **Zs** key, set the *TDR-CableScout* ® to **STEP TDR** and the appropriate Zs (Usually 50 Ω).	

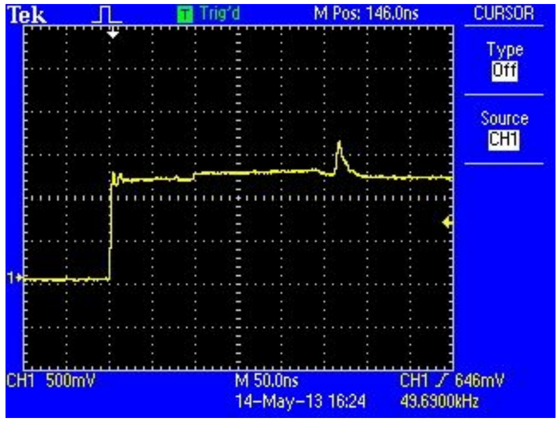
Reflection Coefficient Measurements Steps	Scope Display
2. Connect the Step TDR **SCOPE OUT** to the scope input, use a 50 Ω feedthrough terminator. Set the scope for the display shown.	
 3. Using the **CALC** key, select the Calc. Reflection Coefficient (p) function.	
4. Connect the cable to be tested to the **DUT** output. Adjust the scope to show both the incident pulse and the reflected pulse.	
 5. Adjust your scope for about 6 divisions amplitude. Measure the incident pulse Pk-Pk amplitude.	

Reflection Coefficient Measurements Steps	Scope Display
 6. Measure the reflected pulse Pk-Pk amplitude.	
 7. Press the **ENTER** key. The reflection coefficient will be displayed. Press the down key for additional results.	
 8. Results for Zo, return loss and SWR will be displayed.	

Distanced to Fault (DTF) Measurements

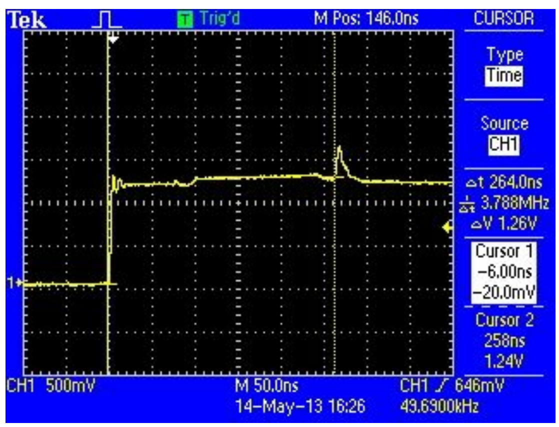
While it is good to know the reflection coefficient, return loss and VSWR, if there is a problem, it would be helpful to know where in the line the problem is located. Fortunately, if we know the cable type and its velocity factor (Vf), we can locate the fault quite easily. The Vf is a part of the cable specification. If you don't know the Vf, it can also be measured with the *TDR-CableScout*®. See the step-by-step instructions below:

Distance to Fault Measurements Steps	Scope Display
 1. Using the **Zs** key, set the output to **STEP TDR** and the appropriate Zs (usually 50 Ω).	
 2. Select the cable from the cable library using the **CABLE/Vf** key. If the cable is not found, set the Vf during the DTF measurement.	
 3. Using the **CALC** key, select the Calc. Reflection Coefficient (p) function.	
 4. Confirm the Vf of your cable or set the desired value.	

<table>
<tr><th>Distance to Fault
Measurements Steps</th><th>Scope Display</th></tr>
<tr><td>5. Connect the cable to be tested to the DUT output. Adjust the scope to show both the incident pulse and the reflected pulse.</td><td>


</td></tr>
<tr><td>


6. If the reflected pulse is not displayed, change the range with the RANGE/DURATION key.</td><td></td></tr>
<tr><td>


7. Press the ENTER Key. Measure the delta time from the incident to the reflected pulse. Enter this value into TDR-CableScout®.</td><td>


</td></tr>
<tr><td>


8. The distance to fault is displayed</td><td></td></tr>
</table>

Velocity Factor Measurements

The velocity factor (Vf), also called wave propagation speed or velocity of propagation (VoP) of a transmission medium is the speed at which a wavefront of an electromagnetic signal or a change of the voltage on a wire passes through the medium, relative to the speed of light.

See table 3. The speed of radio signals in a vacuum, for example, is the speed of light and so the velocity factor of a radio wave in a vacuum is unity (1) , or 100%. In electrical cables, the velocity factor mainly depends on the insulating material.

Vf	**Transmission line**
0.95 - 0.99	Open-wire "ladder" line
0.80	Belden 9085 twin lead
0.82	RG-8X Belden 9258 coaxial cable (foamed polyethylene dielectric)
0.66	RG-213 RG-58 coaxial cable (solid polyethylene dielectric)

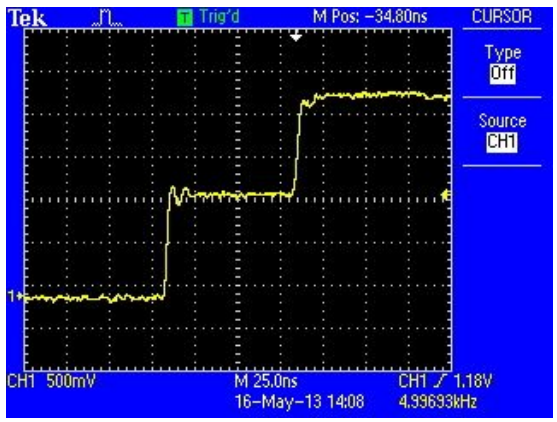
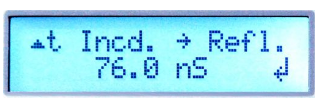
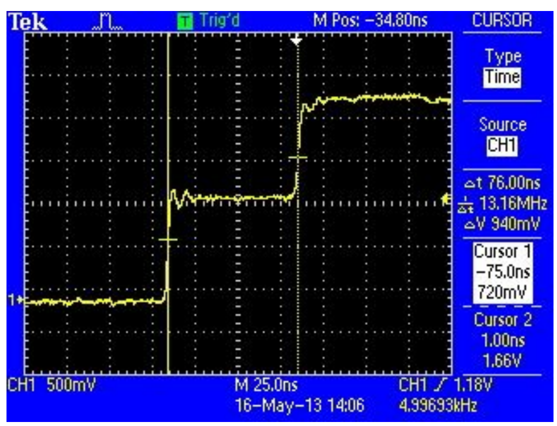
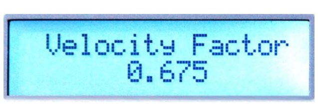
Table 3. Typical velocity factors (Vf) of transmission lines

The use of the terms velocity of propagation and wave propagation is confined to transmission lines and cables. In a ham radio and engineering context, these terms would be understood to mean a true speed or velocity in units of distance per time. Since Vf affects the accuracy of distance to fault measurements, a means to measure Vf is provided. See the step-by-step instructions below:

<table>
<tr><th>Velocity Factor Measurements Steps</th><th>Scope Display</th></tr>
<tr><td>


1. Using the **Zs** key, set the *TDR-CableScout* ® to **STEP TDR** and the appropriate Zs (usually 50 Ω).
</td><td></td></tr>
<tr><td>

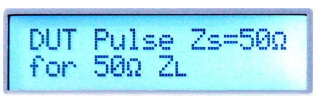
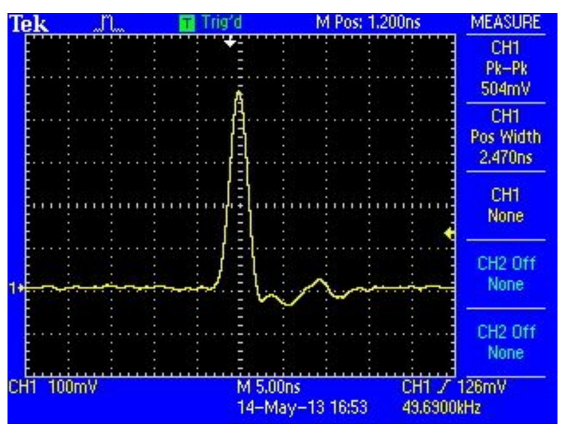

2. Select the measurement choice for Vf using the **CALC** key.
</td><td></td></tr>
</table>

Velocity Factor Measurements Steps	Scope Display
 3. Select Units, Meters or Feet.	
 4. Set the known cable length using the up and down keys.	
5. Connect the cable to be tested to the **DUT** output. Make sure the cable is open at the other end. Adjust the scope to display both the incident pulse and the reflected pulse.	
 6. Measure the delta time from the incident to the reflected pulse. Enter this value into the *TDR-CableScout* ®	
 7. Press **ENTER.** The velocity factor (Vf) is displayed - in this case it is .675. That's pretty close to the published value of .66.	

Line Loss Measurements

Line loss or cable loss is a function of frequency and the length of the cable. It is expressed in dB loss for a given length at a specific frequency. Doubling the length doubles the loss in dB. However, doubling the frequency does not double the loss as the losses in a transmission line due to changes in frequency are proportional to the square root of the frequency. The greater the frequency and length, the greater the loss.

Line loss is a published specification provided by the cable manufacturer. A fundamental contributor to line loss is the dielectric. Dielectric quality and condition can change over time due to environmental conditions such as moisture and mechanical stress. It is not uncommon for cables that have been in service a number of years to have increased line losses. If you want to transfer the maximum power from your transmitter to the antenna, a measurement of line loss is important. See the step-by-step instructions below:

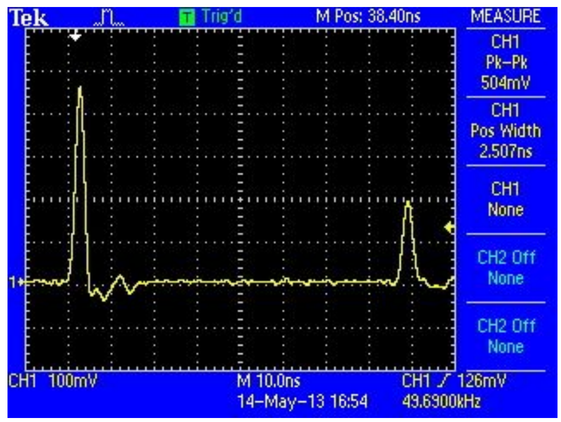
Line Loss Measurement Steps	Scope Display
 1. Using the **Zs** key, set the *TDR-CableScout* ® to pulse TDR.	
 2. Using the **CALC** key, select Cable Loss measurement.	
3. Connect the **PULSE TDR SCOPE OUT** to the scope input using a 50 Ω feedthrough terminator. Set the scope for the display shown.	

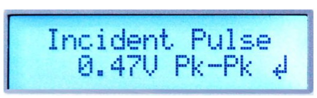
Line Loss Measurement Steps | **Scope Display**

4. Select Units, Meters or Feet.

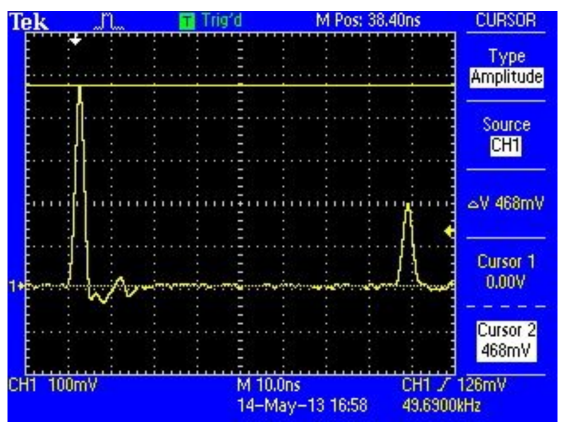
6. Set the Cable Length.

5. Connect the cable to be tested to the **PULSE TDR DUT OUT.** Adjust the scope to show both the incident pulse and the reflected pulse.

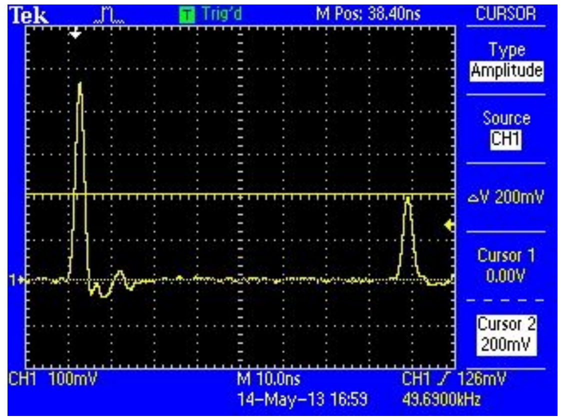




7. Using your scope cursors, measure the incident pulse Pk-Pk amplitude. Enter this value into the *TDR-CableScout*®.



<table>
<tr><th>Line Loss Measurement Steps</th><th>Scope Display</th></tr>
<tr><td>

8. Using your scope cursors, measure the reflected pulse Pk-Pk amplitude. Enter this value into the *TDR-CableScout* ®.</td><td>
</td></tr>
<tr><td>

9. The cable loss is shown on the display.</td><td></td></tr>
</table>

8. Common Transmission Line Faults

In this section we'll take a look at common line faults and what they look like on the oscilloscope. This will help you pin-point the source of the problem and how to fix it.

Common transmission line problems are:

1. Defective shield
2. Pinched cable
3. Line mismatches
4. Faulty connectors
5. Circuit board trace mismatches

TDR Sensitivities

We know that TDR basically measures only two parameters, impedance and time. They are changes in impedances at a given time in the line.

Ideal Line

See figure 16. Assume the ideal condition where Zs = Zo = ZL. Here, all the energy is absorbed by ZL and there are no reflections.

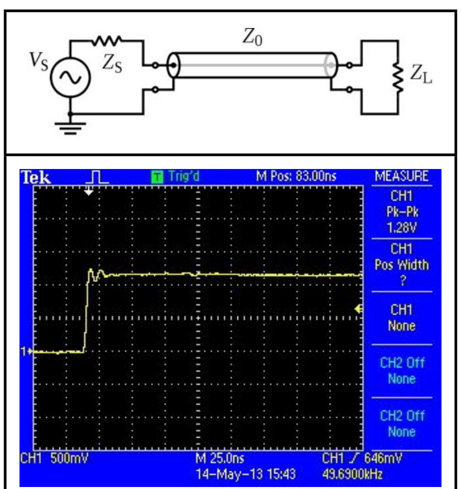


Figure 16. Ideal Transmission Line Zs=Zo =ZL

The ideal line does not exist in reality. There are always some faults; they may be small but they are there. In the following examples, I have purposely induced larger errors to better illustrate the concept.

Inductive Fault

See figure 17. In this example, there is an inductive component in part of the line. This may occur when the shield has been compromised or a connector is defective.

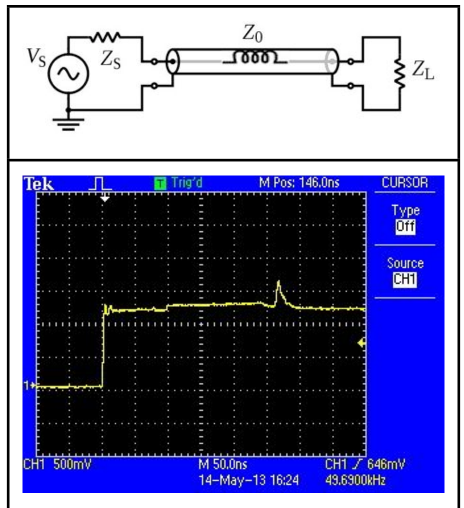


Figure 17. Faulty transmission line Zo contains an inductive component in series

Capacitive Fault

See figure 18. In this example, there is a capacitive component in part of the line. This may occur when the shield has been pinched close to the center conductor in a coaxial transmission line or the line is defective or the loss is due to the dielectric having a localized defect.

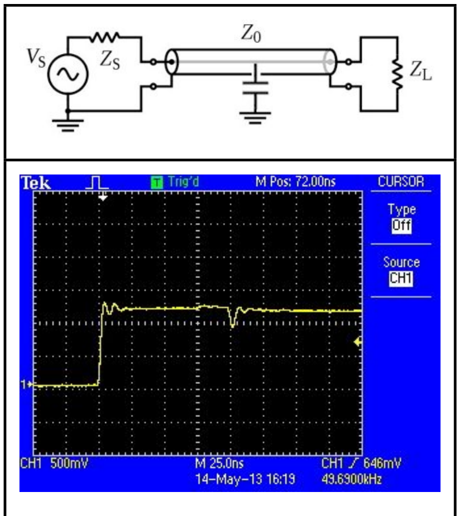


Figure 18. Faulty transmission line Zo contains a capacitive component to ground.

Impedance Mismatch ZL > Zo

See figure 19. In this example, there is an impedance mismatch in part of the line.

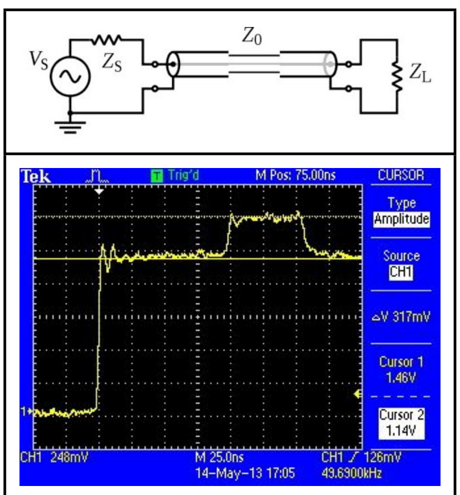


Figure 19. Mismatched transmission Line Zo>ZL=Zs

This may occur when line impedances with differing Zo are connected in series. In this case, a piece of RG 58 50 Ω coax was connected to a section of RG 59 75 Ω coax and then further connected to a section of RG 58. This problem is quite common and can result in unexpectedly high VSWR conditions.

Impedance Mismatch ZL < Zo

See figures 20 and 21. In this example, there is an impedance mismatch in part of the line.

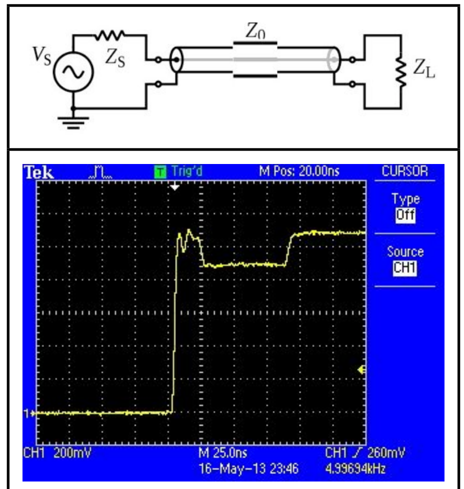


Figure 20. Mismatched transmission Line Zo<ZL=Zs

This may occur when line impedances with differing Zo are connected in series. I purposely set Zs to 75 Ω. In this case, a piece of RG 59 50 Ω coax was connected to a section of RG 58 75 Ω coax and then further connected to a section of RG 59. This problem is quite common and can result in unexpectedly high VSWR conditions.

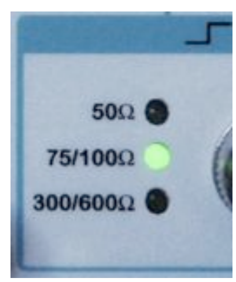


Figure 21. Zs set to 75 Ω

ECB Trace Impedance Variations

See figure 22. In the following example, a circuit board trace was examined for correct strip line design. The scope display clearly shows the impedance variations. The positive going Zo indicates trace width over a ground plane which is narrow (higher impedance) and the negative going variations indicate trace width which are wider (lower impedance).

As a starting point, some microwave engineers use .125” trace width over a ground plane to achieve a Zo of 50 Ω over Fiber Reinforced Plastic (FRP-4) PCB material.

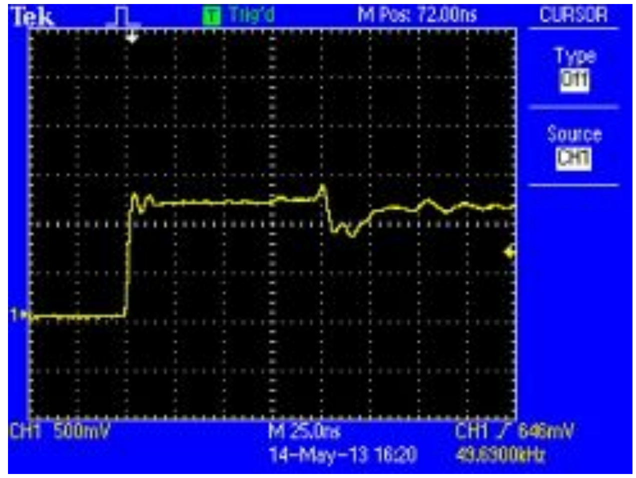


Figure 22. ECB strip line impedance variations

High Resolution Pulse TDR

See figure 23. The fast Tr ≤ 150 ps pulse allows for very high resolution measurements. The length of an SMA connector clearly reveals itself. A TDR measurement of an SMA cable with a Vf of .66 with one end open displays a major reflection at 9.88 ns.

This cable is 38.5 inches in length. The aberrations on the trailing edge are the reflections caused by the connector. Note a cable loss of 10 dB at the 2.2 GHz (the equivalent FFT frequency of the incident pulse). It has been my experience that when displayed on a high bandwidth scope, Zo changes in distance of just a few millimeters are clearly observable.

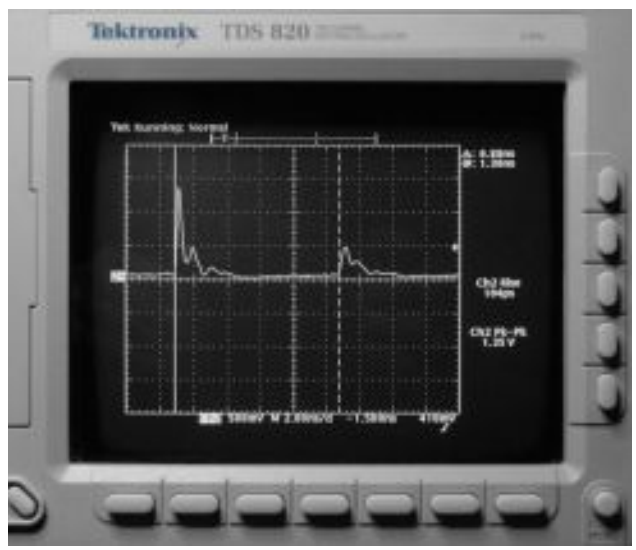


Figure 23. Detail of high resolution TDR measurement when viewed on a Tektronix TDS 820 6 GHz DSO

9. Zo Measurement by Z_L Substitution

See figure 24. We know that if Z_L = Zo there will be no reflections. So, if we could somehow measure Z_L, we would know the line impedance Zo.

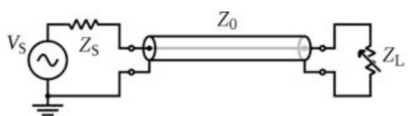


Figure 24. Zo terminated with variable ZL

The VZ500 Variable Terminator

See figure 25. An adjustable Z_L VZ500 is nothing more than a 500 Ω variable resistor which can be adjusted to match the line Zo. The only requirement is that it have low series inductance and low shunt capacitance.

Figure 25. VZ500 variable terminator

Adjust for Minimum Reflections

See figure 26. We have connected the VZ500 variable resistor to a length of RG 59 75 Ω cable and adjusted VZ500 for minimum reflections as observed on the oscilloscope display.

Figure 26. RG 59 cable connected to variable ZL

Reading the R_L Resistance

See figure 27. After confirming that R_L VZ500 resistor has been adjusted for minimum

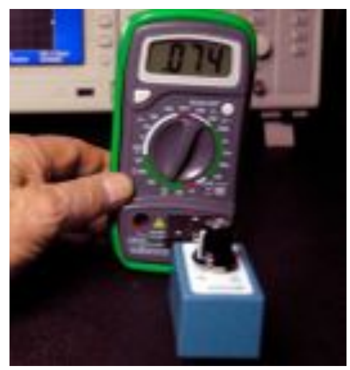


Figure 27. Measurement results of RG 59 cable

reflections, remove it from the cable end and measure the DC resistance with an ordinary ohm meter. In this example, we see that it reads 74 Ω. That is very close to the line's 75 Ω impedance.

10. Additional Information

James A. Strickland, Allen Zimmerman, Gordon Long and George Frye, all from Tektronix at the time, wrote a comprehensive Measurement Concepts paper, "TIME-DOMAIN REFLECTOMETER MEASUREMENTS" in the late 1960's. It is still considered the authoritative reference despite it being more than 40 years old. Of all the papers and information I researched in preparation of this application note, I found it to be most enlightening and fairly easy to read and understand.
http://www.davmar.org/TE/TekConcepts/TekTDRMeas.pdf

If you are into home-brewing circuit board level projects, this application note entitled, "Time Domain Methods for Measuring Crosstalk on PCB Quality Verification" will help you layout your circuit board traces for best-high frequency performance.

http://www.coe.montana.edu/ee/lameres/courses/eele461_spring12/information/TDR_AppNote_Tektronix_Xtalk_11499_EN.pdf

A detailed discussion is also available at the ARRL website:

http://www.arrl.org/files/file/Technology/tis/info/pdf/q1106037.pdf

http://www.arrl.org/files/file/Technology/tis/info/pdf/9706057.pdf

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Microwave & Analog design: Roger Stenbock W1RMS

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