Agilent
Radar Measurements

Application Note

Radar systems today are common and have a variety of uses including scientific, avionic, automotive and military. Even the police officer has a radar system to catch us if we violate the speed limit.

Given the broad range of radar applications, a variety of radar technologies have emerged to meet the unique needs in performance, cost, size, and capability. For example, many police radars use continuous wave (CW) radar to assess Doppler shift from moving cars, but range information is unnecessary. Hence, more advanced capabilities and features take a back seat to low cost and small size. At the other extreme, sophisticated phased-array radars may have thousands of transmit/receive (T/R) modules operating in tandem and may use a variety of sophisticated techniques to improve performance such as: side lobe nulling, staggered pulse repetition interval (PRI), frequency agility, real-time waveform optimization, wideband chirps, and target recognition capability.

Following a brief review of radar basics this application note will focus on the fundamentals of measuring basic pulsed radars as this is the basis for most radar systems. Where appropriate the application note will discuss adaptations of certain measurements for more complex or modulated pulsed radar systems.

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The Fundamentals of Radar Operation

The essence of radar is the ability to gather information about a target — location, speed, direction, shape, identity or simply presence — by processing reflected radio frequency (RF) or microwave signals in the case of primary radars or from a transmitted response in the case of secondary radars.

In most implementations, a pulsed-RF or -microwave signal is generated by the radar system, beamed toward the target in question and collected by the same antenna that transmitted the signal. This basic process is described by the radar range equation found on page 6. The signal power at the radar receiver is directly proportional to the transmitted power, the antenna gain (or aperture size), and the radar cross section (RCS) (i.e., the degree to which a target reflects the radar signal). Perhaps more significantly, it is indirectly proportional to the fourth power of the distance to the target. Given the large attenuation that occurs while the signal is traveling to and from the target, having high power is very desirable, but difficult because of practical problems, such as heat, voltage breakdown, system size and, of course, cost.

Pulse characteristics

The characteristics of a pulsed radar signal largely determine the performance and capability of the radar. Pulse power, pulse repetition rate, pulse width and pulse modulation are traded off to obtain the optimum combination for a given application. Pulse power directly affects the maximum distance, or range, of a target that can be detected by the radar.

Pulse repetition frequency (PRF) determines the maximum unambiguous range to the target. The next (non-coded) pulse cannot be sent until the previous pulse has traveled to the target and back. (Coded pulses can be sent more frequently because coding can be used to associate responses with their corresponding transmitted pulse.)

Pulse width determines the spatial resolution of the radar: pulses must be shorter than the time it takes for the signal to travel between the target details; otherwise, the pulses overlap in the receiver.

The pulse width and the shape of the pulse also determine the spectrum of the radar signal. Decreasing the pulse width increases signal bandwidth. A wider system bandwidth results in higher receiver noise for a given amount of power, which reduces sensitivity. Also, the pulse spectrum may exceed regulated frequency allotments if the pulse is too short.

The pulse shape can be the familiar trapezoidal pulse with rapid but controlled rise and fall times, or any of a number of alternative shapes such as Gaussian and raised-cosine. The pulse shape can determine the signal bandwidth and also affects the detection and identification of targets and therefore varies based on application.

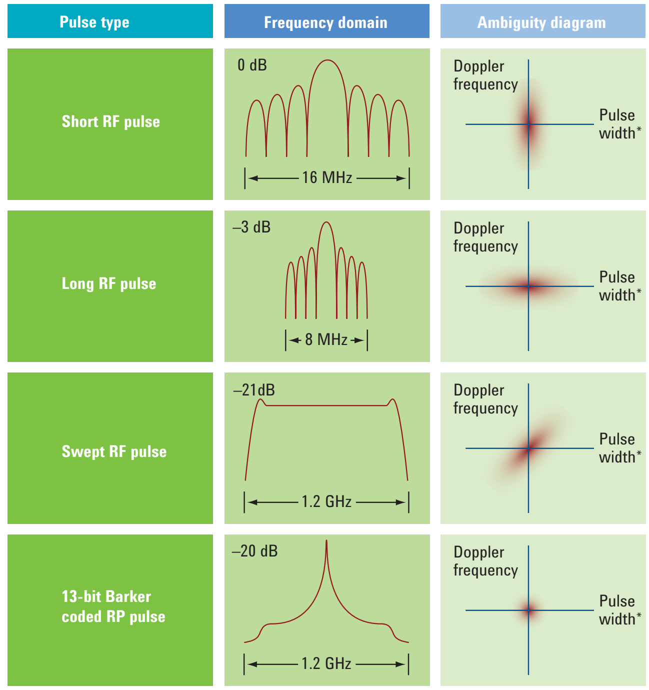
Short pulses with a low repetition rate maximize resolution and unambiguous range and high pulse power maximizes the radar's range in distance. However, there are practical limitations in generating short high-power pulses. For example, higher peak power will shorten the life of tubes used in an amplifier design. This conundrum would be the barrier to increasing radar performance if radar technology stopped here. However, using more complex waveforms and pulse compression techniques the power limitation on pulse width can be greatly mitigated.

Pulse compression

Pulse compression techniques allow relatively long RF pulses to be used without sacrificing range resolution. The key to pulse compression is energy. Using a longer pulse, one can reduce the peak power of the transmitted pulse while maintaining the same pulse energy. Upon reception, the pulse is compressed with a match correlation filter into a shorter pulse, which increases the peak power of the pulse and reduces the pulse width. Pulse compressed radar thereby realizes many of the benefits of a short pulse including: improved resolution and accuracy, reduced clutter, better target classification, and greater tolerance to some electronic warfare (EW) and jamming techniques. One area that does not realize an improvement is minimum range performance. Here the long transmitter pulse may obscure targets that are close to the radar.

The ability to compress the pulse with a match filter is achieved by modulating the RF pulse in a manner that facilitates the compression process. The matching filter function can be achieved digitally using the cross-correlation function to compare the received pulse with the transmitted pulse. The sampled receive signal is repeatedly time shifted, Fourier transformed and multiplied by the conjugate of the Fourier transform of the sampled transmit signal (or a replica). The output of the cross correlation function is proportional to the time shifted match of the two signals. A spike in the cross correlation function or matching filter output occurs when the two signals are aligned. This spike is the radar return signal, and it typically may be 1000 times shorter in time duration than the transmitted pulse. Even if two or more of the long transmitted pulses overlap in the receiver, the sharp rise in output only occurs when each of the pulses are aligned with the transmit pulse. This restores the separation between the received pulses, and with it, the range resolution. Note that the receive waveform is windowed using a Hamming or similar window to reduce the time-domain sidelobes created during the cross-correlation process.

Ideally, the correlation between the received and transmitted signals would be high only when the transmit and receive signals are exactly aligned. Many modulation techniques are used to achieve this goal including: linear FM sweep, binary phase coding (e.g., Barker codes), or polyphase codes (e.g., Costas codes). Graphs called ambiguity diagrams illustrate how different pulse compression schemes perform as a function of pulse width and Doppler frequency shift, as shown in Figure 1. Doppler shift can reduce detector sensitivity and cause errors in the time alignment.



* The term pulse width on the ambiguity diagram refers to the pulse width at the radar detector output.

Figure 1. An ambiguity diagram illustrates location accuracy versus Doppler accuracy. Shown in this figure are relative ambiguity diagrams for different types of radar pulses.

Although Doppler frequency shift can cause errors, it also gives radar operators important information about the target.

Doppler frequency

Most targets of interest are moving. Moving targets cause the frequency of the returned signal to be shifted higher if the target is moving toward the radar and lower if the target is moving away. This is the familiar Doppler frequency shift often associated with passing ambulances and trains. As many people who have received speeding tickets can attest, police radars using Doppler frequency shift can determine their car's (target's) radial velocity. In many radar systems, both the location and radial velocity are useful information.

Doppler frequency shift can reduce the sensitivity of location detection. Recall that the output of the cross-correlation filter used for detection is proportional to the match between the received and transmitted signals. If the received signal is slightly lower or slightly higher in frequency, then the output of the cross-correlation filter is somewhat lower. For a simple pulse, the response of the cross correlation filter follows the familiar SIN X/X shape as a function of Doppler frequency In extreme cases, the receive signal may be so shifted in frequency that it correlates with one of the sidelobes of the transmit signal. Note that short pulses have a relatively wide initial lobe in the sin(x)/x response and so they tend to be Doppler tolerant compared to longer pulses. Figure 1 compares the ambiguity diagrams for the short and long pulses. In other pulse compression schemes, such as Barker coding, the matching filter output drops off much faster than the SIN X/X of the simple pulse, which makes them Doppler intolerant. Doppler shift in linear FM modulated pulses can create an error in the location information because the highest cross correlation occurs where the swept frequencies in the receive pulse are best aligned with the swept frequencies in the transmit pulse. This offset is directly proportional to the Doppler shift.

The Radar Range Equation

The radar range equation describes the important performance variables of the radar and provides a basis for understanding the measurements that are made to ensure that radar performs optimally. This section will describe the basic derivation of the range equation and examine the important performance variables identified by the equation. The remainder of this application note will then discuss the measurement methods and options that are available to evaluate these performance variables.

The derivation starts by analyzing a simple spherical scattering model of propagation for an isotropic radiator, or point source antenna. Assume, for simplicity, that the antenna equally illuminated the interior of an imaginary sphere with equal power density in each unit of surface area, where the surface area of the sphere is:

$$A_s = 4(\pi)R^2$$

Where:

A_s = area of a sphere
R = radius of the sphere

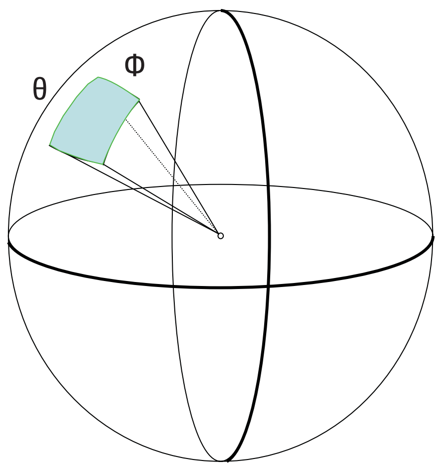


Figure 2. Ideal isotropic antenna radiation

The power density is found by dividing the total transmit power, in watts, by the surface area of the sphere in square meters.

$$\rho = \frac{P_t}{A_s} = \frac{P_t}{4\pi R^2}$$

Where:

ρ = power density in watts per square meters
P_t = total transmitted power in watts

Since radar systems use directive antennas to focus radiated energy onto a target, the equation can be modified to account for the directive gain G of the antenna. The gain of the antenna is defined as the ratio of power directed toward the target compared to the power from an ideal isotropic antenna.

$$\rho_T = \frac{P_t G_t}{4\pi R^2}$$

Where:

ρ_T = power density directed toward the target from directive antenna
G_t = gain of the directive antenna

This equation describes the transmitted power density that strikes the target. Some of that energy will be reflected off in various directions and some of the energy will be reradiated back to the radar system. The amount of this incident power density that is reradiated back to the radar is a function of the radar cross section (RCS or σ) of the target. RCS (σ) has units of area and is a measure of target size, as seen by the radar. With this information the equation can be expanded to solve for the power density returned to the radar antenna. This is done by multiplying the transmitted power density by the ratio of the RCS and area of the sphere.

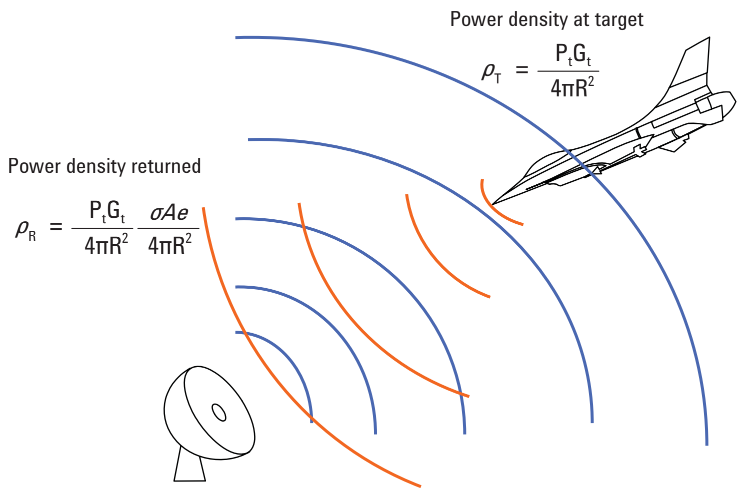


Figure 3. Transmitted and reflected power returned to the radar

$$\rho_R = \frac{P_t G_t}{4\pi R^2} \frac{\sigma}{4\pi R^2}$$

Where:

ρ_R = power density returned to the radar, in watts per square meter
σ = RCS in square meters

A portion of this signal reflected by the target will be intercepted by the radar antenna. This signal power will be equal to the return power density at the antenna multiplied by the effective area, Ae of the antenna.

$$S = \frac{P_T G_T \sigma A_e}{(4\pi)R^4}$$

Where:

S = signal power received at the receiver in watts
P_T = transmitted power in watts
G_T = gain of transmit antenna (ratio)
σ = RCS in square meters
R = radius or distance to the target in meters
A_e = effective area of the receive antenna square meters

Antenna theory allows us to relate the gain of an antenna to its effective area as:

$$A_e = \frac{G_R \lambda^2}{4\pi}$$

Where:

G_R = gain of the receive antenna
λ = wavelength of radar signal in meters

The equation for the received signal power can now be simplified. Note that for a monostatic radar the antenna gain G_T and G_R are equivalent. This is assumed to be the case for this derivation.

$$S = \frac{P_T G_T G_R \lambda^2 \sigma}{(4\pi)^2 R^4 4\pi} \longrightarrow S = \frac{P_T G^2 \lambda^2 \sigma}{(4\pi)^3 R^4}$$

Where:

S = signal power received at the receiver in watts
P_T = transmitted power in watts
G = gain antenna (assume same antenna for transmit and receive)
λ = wavelength of radar signal in meters
σ = RCS of target in square meters
R = radius or distance to the target in meters

Now that the signal power at the receiver is known, the next step is to analyze how the receiver will process the signal and extract information. The primary factor limiting the receiver is noise and the resulting signal-to-noise (S/N) ratio.

The noise power (theoretical limit) at the input of the receiver is described as Johnson noise or thermal noise, is a result of the random motion of electrons, and is proportional to temperature.

$N = kTB_n$

Where:

N = noise power in watts
k = Boltzmann's constant (1.38×10^{-23} Joules/degrees Kelvin)
T = temperature in Kelvin
B_n = system noise bandwidth

At a room temperature of 290 °K the available noise power at the input of the receiver would then be 4×10^{-21}W/Hz, –203.98 dBW/Hz, or–173.98 dBm/Hz.

The available noise power at the output of the receiver will always be higher than predicted by the above equation due to noise generated within the receiver. The output noise will be equal to the ideal noise power multiplied by the noise factor and gain of the receiver.

$N_o = GF_n kTB$

Where:

N_o = total receiver noise
G = gain of the receiver
F_n = noise factor

In addition to noise factor other limiting factors include oscillator noise (such as phase noise or AM noise), spurs, residuals, or images. These signals may or may not be noise-like but will impact the receiver's ability to process the received signals. For simplicity these factors will not be considered specifically in this derivation. However, phase noise and spurs are important radar measurements that can affect radar performance and are therefore included as part of the measurement discussion later in this application note.

As just noted, the available noise power at the output of the receiver will always be higher than the thermal Johnson noise. This is because extra noise is generated within the receiver. The total output noise will be equal to the Johnson noise power multiplied by the noise factor F_n and gain G of the receiver.

The gain of the receiver can be rewritten as the ratio of the signal output of the receiver to the signal input, ($G = S_o/S_i$). Solving for the noise factor F_n the equation then becomes:

$$F_n = \frac{S_i / N_i}{S_o / N_o} \quad \text{Where } N_i = kTB$$

By definition the noise factor is the ratio of the S/N in to the S/N out.

The equation can be rewritten in a different form:

$$F_n = \frac{N_o}{kT_oB_nG} \quad \text{Where } G = \frac{S_o}{S_i}$$

Where:

F_n = noise factor
N_o = total receiver noise
G = gain of the receiver
S_o = receiver output signal
S_i = receiver input signal
T_o = room temperature
k = Boltzmann's constant
B_n = noise bandwidth of the receiver

Since noise factor describes the degradation of signal-to-noise as the signal passes through the system, the minimum detectable signal (MDS) at the input can be determined, which corresponds to a minimum output S/N ratio with an input noise power of kTB.

$Si \longrightarrow S_{min}$ when minimum S_o/N_o condition is met

$$S_{min} = kT_oB_nF_n\left(\frac{S_o}{N_o}\right)_{min}$$

Where:

S_{min} = minimum power required at input of the receiver
F_n = noise factor
$(S_o/N_o)_{min}$ = min ratio required by the receiver processor to detect the signal

Now that the minimum signal level required to overcome system noise is defined, the maximum range of the radar can be calculated by equating the MDS (S_{min}) to the signal level reflected from our target at maximum range (set S_{min} = to equation for S above).

$$S_{min} = kT_oB_nF_n\left(\frac{S_o}{N_o}\right)_{min} = \frac{P_T G^2 \lambda^2 \sigma}{(4\pi)^3 R^4_{max}}$$

From this equation we can solve for that maximum range of our radar.

$$R_{max} = \frac{P_T G^2 \lambda^2 \sigma}{kT_oB_nF_n(S/N)\,(4\pi)^3}$$

Where:

R_{max} = maximum distance to detectable target in meters
P_T = transmitted power in watts
G = antenna gain (assume same antenna for transmit and receive)
λ = wavelength of radar signal in meters
σ = RCS of target in square meters
k = Boltzmann's constant
T = room temperature in Kelvin degrees
B_n = receiver noise bandwidth in Hz
F_n = noise factor
S/N = min signal-to-noise ratio required by receiver processor to detect the signal

This equation describes the maximum target range of our radar based on transmitter power, antenna gain, RCS of the target, system noise figure, and minimum signal-to-noise ratio. In reality this is a simplistic model of system performance. There are many factors that will also affect system performance including modifications to the assumptions made to derive this equation. Two additional items that should be considered are system losses and pulse integration that may be applied during signal processing. Losses in the system will be found both in the transmit path (LT) and in the receive path (LR). In a classical pulsed radar application we could assume that multiple pulses would be received, from a given target, for each position of the radar antenna (since the radar's antenna beamwidth is greater than zero, we can assume that the radar will dwell on each target for some period of time) and therefore could be integrated together to improve the performance of our radar system. Since our integration may not be ideal we will use an integration efficiency term $E_i(n)$, based on the number of pulses integrated, to describe integration improvement.

Including these terms the radar equation yields:

$$R^4 = \frac{P_T G^2 \lambda^2 \sigma E_i(n)}{kTB_n F_n (S/N)(4\pi)^3 L_T L_R}$$

Where:

L_T = losses in the transmitter path
L_R = losses in the receive path
$E_i(n)$ = integration efficiency factor

The entire equation can be converted to log form (dB) to simplify the discussion:

$$40\text{Log}(R) = P_T + 2G_{dB} + 20\ \text{Log}_{10}\lambda + \sigma_{dBm} + E_{idB}(n) + 204\text{dBW/Hz} - 10\text{Log}(B_n) - F_n - (S/N) - L_{TdB} - L_{RdB} - 33\text{dB}$$

Where:

R = maximum distance in meters
P_T = transmit power in dBW
G = antenna gain in dB
λ = wavelength of radar signal in meters
σ = RCS of target measured in dBsm, or dB relative to a square meter
F_N = noise figure (Noise figure = noise factor converted to dB)
S/N = min signal –to-noise ratio required by receiver processing functions detect the signal in dB

The 33 dB term comes from 10 $\log(4\pi)^3$ and the 204 dBW/Hz from Johnson noise at room temperature.

The dB term for RCS σ_{dBsm} is determined in dBsm or dB relative to a one meter sphere (sphere with cross section of a square meter), which is the standard target for radar cross section measurements.

2. 0 Radar Block Diagram and the Radar Range Equation

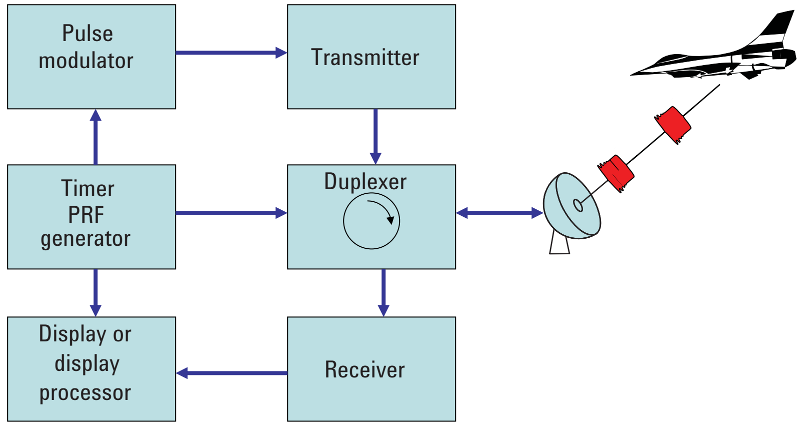


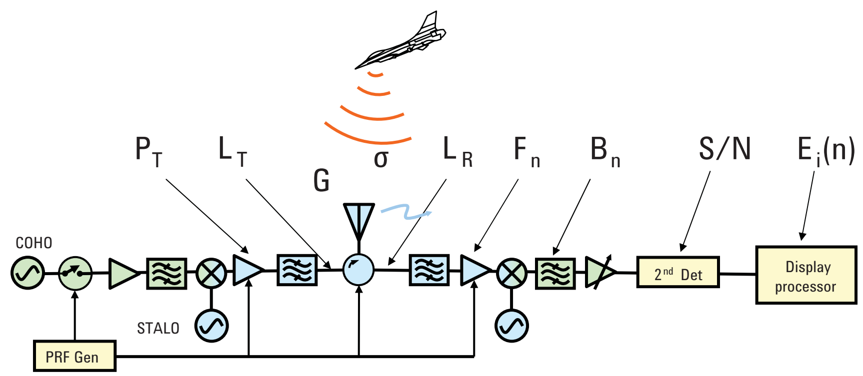
Figure 4. Basic radar block diagram

Figure 4 shows a basic radar block diagram. The diagram could be much more complicated; however this diagram shows all of the essential blocks of the radar system. The diagram shows the master timer or PRF generator as the central block of the system. The PRF generator will time synchronize all components of the radar system shown in Figure 4 by the connections to the pulse modulator, duplexer or transmit/receive switch, and the display processor. In addition, connections to the receiver would provide gating for front-end protection or timed gain control such as sensitivity time control (STC).

For the purpose of this application note we will focus on the transmitter, receiver, duplexer, and antenna sections of this diagram. As these blocks are expanded, the parameters of the range equation will be allocated to each block or component.

Relating the Range Equation to the Elements of the Radar Design

In Figure 5 we have expanded the transmitter and receiver blocks of our block diagram to identify some typical components. This block diagram could vary quite extensively depending on the type of transmitter employed. For example, if our transmitter used a magnetron power oscillator as its output stage the diagram would be greatly simplified. Also shown is the simplified radar equation identifying the major blocks in our diagram that have the major significance for each parameters.



$$40\text{Log}(R) = P_T + 2G + 20\,\text{Log}_{10}\lambda + \sigma + E_i(n) + 204\text{dBW/Hz} - 10\text{Log}(B_n) - F_n - (S/N) - L_T - L_R - 33\text{dB}$$

Figure 5. Relating the radar range equation to the basic transmit and receive design.

Power, Spectrum and Related Measurements

Generally, a radar transmitter is the most costly component of the system with the highest power consumption, most stringent cooling requirements, and greatest influence on system performance.

There are many different terms used when talking about power as shown in Figure 6. Average power is the power that is integrated over the complete time waveform (on time and off time) of the radar. If the pulse width and PRF are not constant, the integration time must be long enough to represent all possible variations in pulse parameters. Most typical RF and microwave power meters are average power meters and respond to the heating energy of the signal. Peak power is the maximum instantaneous power. Pulse power is the integrated or average power for one complete pulse.

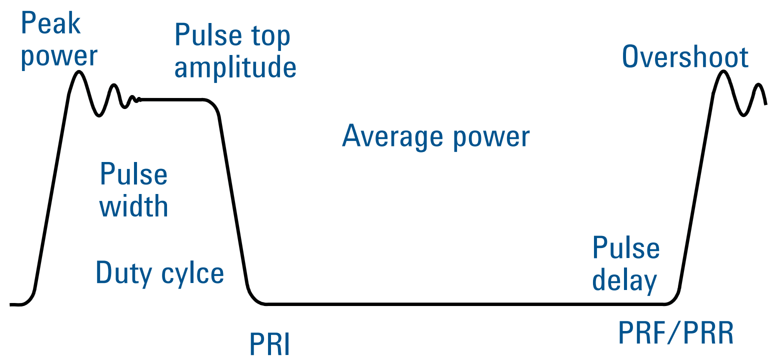


Figure 6. Pulse parameters

Other parameters including duty cycle, pulse width, PRF, rise and fall times shown in Figure 6 are useful for characterizing the power of the radar signal.

From a radar equation standpoint, the power term corresponds to the power of the transmit pulse. If the integration term is excluded, the equation applies to a single pulse. Therefore, it can be useful to examine the peak and pulse power on an individual pulse basis. This technique is becoming more important for modern radar systems where pulse width and PRF are dynamically adjusted and pulse profile may be contoured to improve system performance. It is also becoming easier to do with the use of modern test equipment.

However, average power measurements are a common method for characterizing the power of a radar signal. It is simple to measure and requires only low cost instruments. If pulse characteristics, such as the duty cycle of the radar signal, are known, the pulse power can be derived or estimated based on average power. Note, however, that this derived result does not provide information about droop, or any peak excursions that may occur due to ringing or overshoot. The result would be nearly equivalent to the pulse top amplitude, and in the case of a perfectly square pulse, it would be equivalent to the true peak power or pulse power.

Along with measuring power the spectrum shape is critical to verify that a radar system is operating efficiently. An unsymmetrical or incorrect spectral shape indicates radar that is operating less than optimally. For example, the radar may be wasting power by transmitting or splattering power at unwanted frequencies causing out-of-band interference. For some radar systems, pulse shaping is used to reduce the level of the spectral sidelobes, to improve the efficiency and life of radar components, and to reduce the bandwidth.

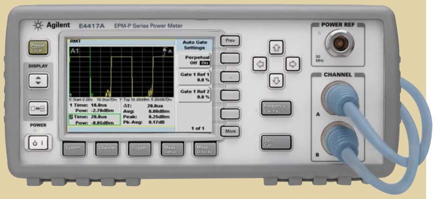
There are several options for measuring radar power, pulse characteristics and spectrum including the use of a power meter, spectrum analyzer, or vector signal analyzer. Each instrument has its advantages and limitations. The best instrument is determined by the measurement objectives and the restraints of the radar and test instrument. This section will discuss making measurements with each of these instruments.

Maximum instrument input level

One of the first things to be considered is the magnitude of RF power that could be encountered. Parameters such as frequency, antenna match (SWR), pulse width (PW), pulse repetition time (PRT), and duty cycle will affect power measurements and selection of measurement hardware.

RF and microwave instruments are limited in both the amount of average power and the amount of peak power that can be input without damaging the instrument. For typical radar systems, with pulse powers of approximately 1 MW, a directional coupler is required to sample the transmitter power and provide a safe drive level to the test instrument.

P-Series Power Meters



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Measuring pulse power with a power meter

The most common and lowest cost method of measuring pulse power is using a power meter. The right power meter can provide a number of measurements including average power, peak power, duty cycle, and even power statistics. One of the first things to consider when measuring with a power meter is the power sensor.

The power sensor

The basic idea behind a power sensor is to convert high frequency power to a DC or low frequency signal that the power meter can then measure and relate to a certain RF or microwave power level. The three main types of sensors are thermistors, thermocouples, and diode detectors. There are benefits and limitations associated with each type of sensor. We will briefly go into the theory of each type and talk about the advantages and limitations associated with each sensor.

Thermister based sensors

Thermister based sensors implement a balanced Wheatstone bridge. When RF power is applied to a thermister in the bridge it warms and its resistance decreases. This change in resistance unbalances the bridge and produces a differential input to the amplifier. Being a feedback loop, the amplifier decreases the DC bias to the bridge enough to bring the bridge back into balance. The difference in DC power can then be measured with the power meter and related to the RF power incident on the thermistor.

The downside with these types of sensors is their sensitivity to temperature changes. To help with this problem a second thermistor may be added to detect and correct for the ambient temperature.

Themocouple-based sensors

Thermocouple sensors are based on the fact that metal generates a voltage due to temperature differences between a hot and cold junction and that different metal will create different voltages. Thermocouple sensors exploit this behavior by detecting and correlating these voltage changes, which are a result of changes in temperature caused by incident RF power on the thermocouple element. Since the change of voltage is small, many junctions are used connected in series, which is referred to as a thermopile.

Both thermisters and thermocouple-based sensors can be used to measure average power but can not be used to directly measure peak power.

Diode-based sensors

Unlike thermistors and thermocouples, a diode does not measure the heat content of a signal but rectifies the signal instead. The matching resistor (approximately 50 ohms) is the termination for the RF signal. RF voltage is rectified and converted to DC voltage at the diode. A bypass capacitor is used as a low-pass filter to remove any RF signal getting through the diode. A major attribute of the diode sensor is sensitivity, permitting power measurements as low as –70 dBm (100 pW).

One should consider, however, if these measurements are independent of signal content. If we expand the diode equation into a power series, we find that the rectified output voltage is a function of the square of the input signal voltage up to a power level of about –20 dBm. This performance yields a rectified output that is proportional to the RF signal power regardless of signal content.

As the power level increases above –20 dBm, the rectification process becomes more and more linear and the output voltage transitions to a function of the input voltage (rather than the square of the input voltage). For complex signals, the output is then dependent upon the phase relationships among the various components of the input signal. In other words, the output is related to the peak power of the signal rather than the heating power of the signal contained within the video bandwidth (VBW) of the sensor. For pulsed signals, this is very important.

Many of today's average power measurements require a dynamic range greater than 50 dB. Agilent's approach to creating a wide dynamic range, average power sensor to meet this need is to incorporate diode stacks in place of single diodes to extend square law operation to higher power levels at the expense of sensitivity. A series connection of m diodes results in a sensitivity degradation of 10*log(m) dB and an extension upwards in power of the square law region maximum power of 20*log(m) dB, yielding a net improvement in square law dynamic range of 10*log(m) dB compared to a single diode detector. The E-series E9300 power sensors are implemented with a two diode stack pair for the low power path (–60 to –10 dBm), a resistive divider attenuator and a five diode stack pair for the high power path (–10 to +20 dBm).

Measuring power with an average power meter

An average power meter can be used to report average power and pulse power if the duty cycle of the signal is known. There are some advantages to using this method but there are also a number of points that must be taken into consideration. When an average power meter reports a pulse or peak power result, it does so by deriving the result from the average power and a known duty cycle. The result is accurate for an ideal or close to ideal pulse signal but it does not reflect aberrations due to a non-square pulse shape and will not detect peak excursions that may result from ringing or overshoot. The main advantage of average power meters is that they are the lowest cost solution. Both the power meters and sensors are less expensive than the corresponding peak power meters and sensors. They also generally have the ability to measure over a wider dynamic range, frequency range, and bandwidth, and they can measure a signal no matter how fast the rise time or narrow the pulse width.

Figure 7 shows an example of measuring pulse and average power using an average power meter and a known duty cycle. The example uses a simple 10 us pulse with period of 40 us. The pulse signal is set with a power level of approx 0 dBm. The average power result is –6.79 dBm. Since the duty cycle is known (10 us divided by 40 us, or 25%) it can be entered into the power meter to obtain the result of the pulse power, which is measure at –0.77 dBm. The equation and calculation used to come to this result is shown in Figure 7.

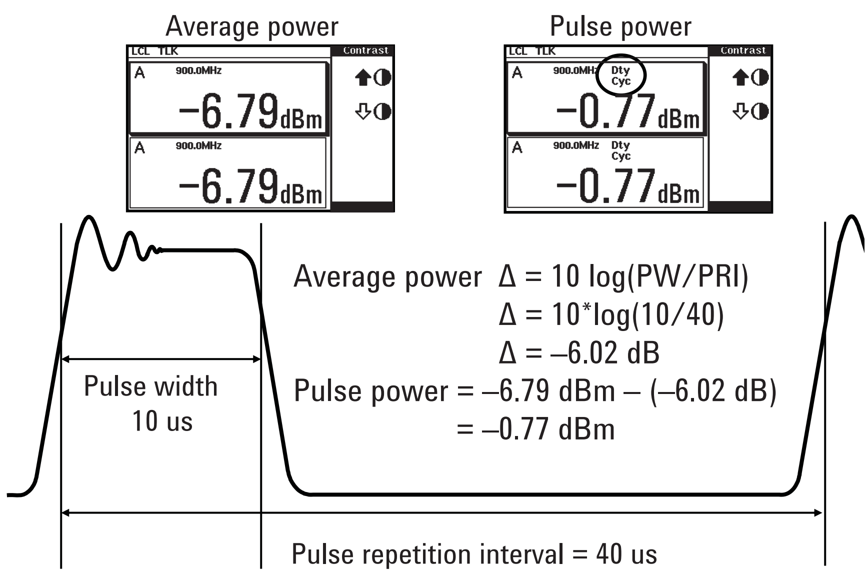


Figure 7. Using an average power meter to derive pulse power.

In reality, as was previously mentioned, the pulse may not be purely rectangular since there is an associated rise and fall time, as well as possible overshoot on the signal. The combination of these effects creates an error in the calculated result.

Measuring power with a peak power meter

The peak power meter with sensor has the advantage in that it is able to measure peak power and pulse power directly. This is particularly useful for shaped or modulated pulses for which deriving pulse power from average power may be inadequate.

Figure 8 shows an example of measuring peak and pulse power using the Agilent P-Series peak power meter. A convenient feature of the meter is that it has a trace display that allows a person to view the envelope of the pulse signal that is being measured. The meter operates by continuously sampling the signal with a 100 MS/s digitizer, buffering the data, and calculating the result. This gives the meter measurement flexibility including flexible triggering, time gating with multiple gates and the ability to take single shot measurements.

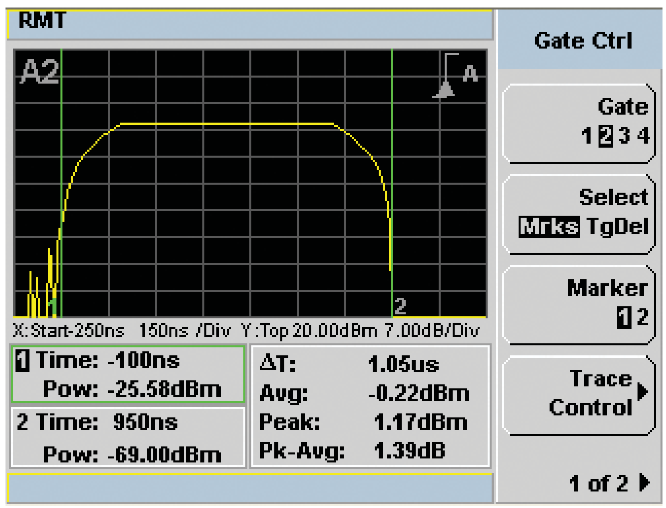


Figure 8. Using the P-Series peak power meter to measure peak power, gated pulse power, and peak-to-average ratio. Due to the shape of this pulse the peak power is 1.39 dB higher than the pulse power.

In Figure 9, a time gate is set up to measure over just one pulse. The meter is able to simultaneously report the peak power, average power, and peak-to-average power result within the time gate. In this case, the reported average power result of –0.09 dBm is equal to the pulse power since the time gate is set to specifically measure one pulse. The peak power is slightly higher at 0.24 dBm, a likely result due to some overshoot. The peak-to-average ratio of 0.32 dB is the difference between the two numbers.

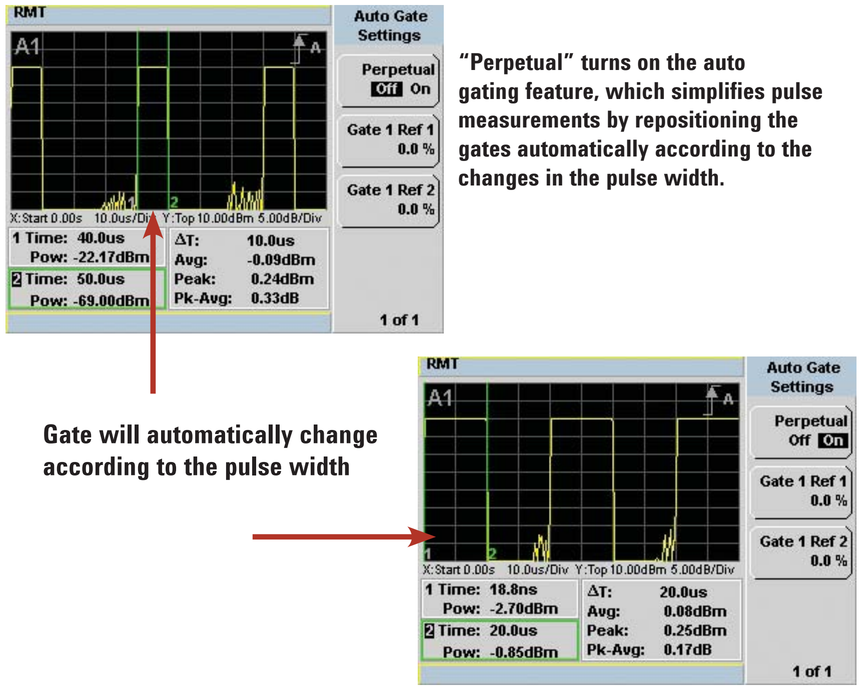


Figure 9. The P-Series power meter has the ability to directly measure the pulse power by automatically positioning the gate on the pulse.

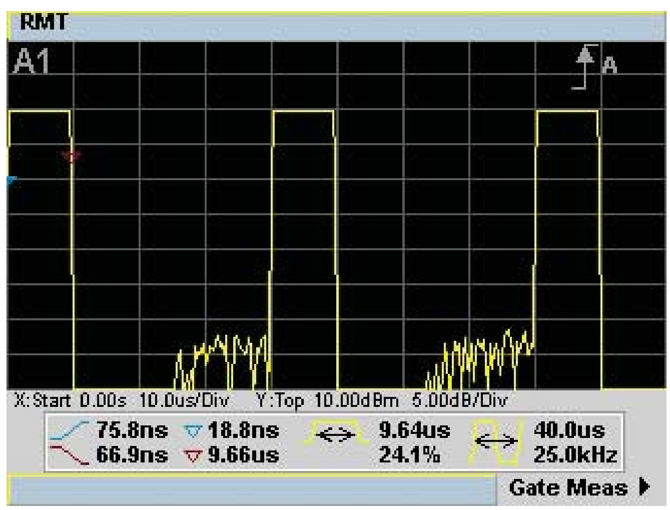


Figure 10. The P-Series peak power meter will automatically measure pulse characteristics including rise time, fall time, pulse width, and pulse period.

The P-Series power meter also has other features that are convenient for measuring radar signals. Figure 10 illustrates the ability of the meter to automatically measure pulse characteristics such as pulse width, pulse period, rise time, and fall time. The ability of the meter to automatically adjust the time gate to the width of the pulse using the perpetual setting in the meter is shown in Figure 9. This makes measuring the pulse power simple without requiring prior knowledge of the pulse width. This is especially convenient for radars that have dynamic pulse widths and PRI.

Peak power meters do have their limitations. The frequency coverage is one limitation. For example, the P-Series peak sensor maximum range is 40 GHz compared to 110 GHz for the average power sensor. Peak power meters with sensors typically also have limitations on their power range. The range of the P-Series power sensor is approximately –35 dBm to +20 dBm compared to the E9300 average power sensors that can range from –60 dBm to +20 dBm. Peak power meters are also limited in rise times, pulse widths and the modulation bandwidth of the signal they can measure. These limitations can be partially controlled by the video bandwidth setting.

Power meter video bandwidth

In simplest terms, the video bandwidth (VBW) of the power meter gives an indication of how fast it can track signal variations of the peak power envelope and is an indication of the modulation bandwidth that can be accurately measured.

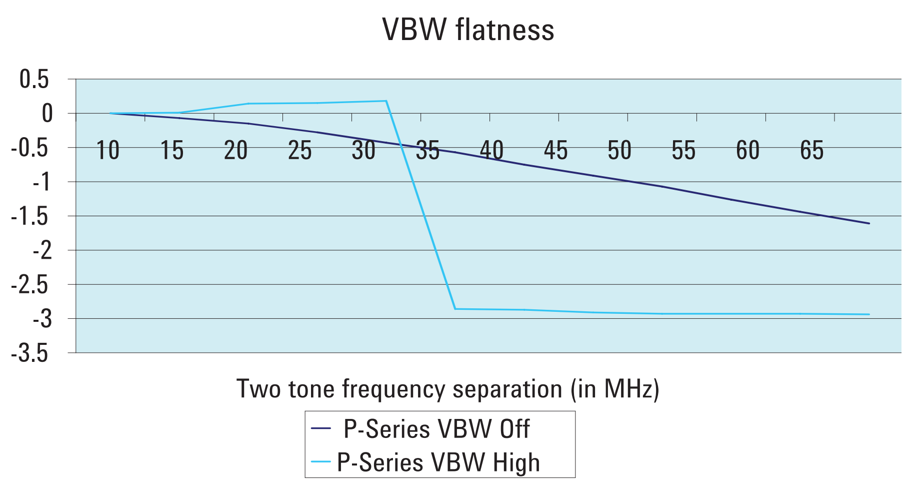
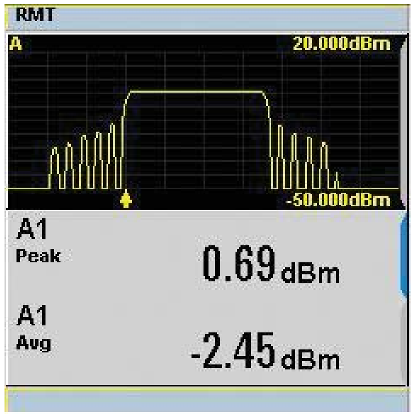
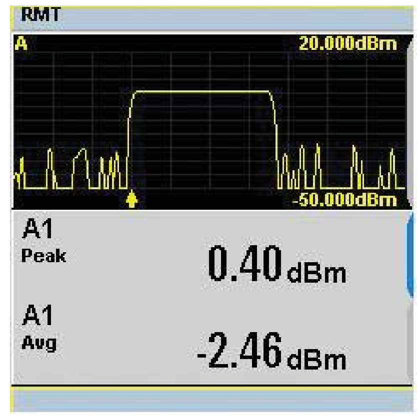


Figure 11. This chart shows the measured flatness of the P-Series power meter with the 30 MHz VBW filter turned on and off. Generally, it is recommended to turn off the VBW when measuring radar to maximize the bandwidth.



Video BW on results in ringing due to filter shape



Video BW OFF no ringing

Figure 12. Effects of VBW on a power meter measurements of a radar with fast rising pulse edge.

When measuring pulsed radar however, it is generally recommended that the video bandwidth filter be turned off. This maximizes the bandwidth of the meter, as shown in Figure 11, and avoids ringing that can occur due to the interaction of the sharp roll off of the video bandwidth filter and the rising edge of the radar pulse, as shown in Figure 12. The price paid for not using the video bandwidth filter is degradation in the meter flatness that the video bandwidth helps correct, as shown in Figure 11. However, for a pulsed radar signal whose spectrum falls off as sin(x)/x the advantage of this correction is small and therefore best results are usually achieved without using the filter. If the full bandwidth of the radar is within the maximum VBW setting then turning the video bandwidth filter on can improve the accuracy and measurement range.

Note that in the case of FM modulated (chirped) radars the modulation bandwidth can be distinguished from the RF bandwidth. Chirps may have a very wide RF bandwidth due to frequency change, but since the modulation does not affect the amplitude of the signal, it will not be limited by the video bandwidth constraint of the power meter.

Overall the power meter will have a limit on the rise time it can measure, and the minimum pulse width for which it can achieve a full peak response. For the reasons mentioned above the fastest rise time is achieved with the video bandwidth filter turned off. As an example, the rise time of the P-Series peak power meter and sensor is about 13 ns. The minimum pulse width for which it is able to accurately measure the pulses amplitude top is 50 ns.

PSA Series High-Performance Spectrum Analyzers

Maximum performance and flexibility for measuring radar

- FFT-mode for high-resolution narrow-band measurements
- Swept-mode for high-speed wideband measurements & maximum dynamic range
- Built-in measurement functions for channel power, burst power, & occupied bandwidth
- Vector signal analysis with 89601A software
- 80 MHz instantaneous bandwidth, 14 bit resolution & 512 Mbyte of memory
- Noise figure and phase noise measurement options

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Measuring pulse power and spectrum with a spectrum analyzer

The primary advantage of a spectrum analyzer is that it is able to measure frequency content of the radar in addition to power. This is important because the incorrect spectrum can indicate a number of problems that result in wasted power and the emission of unintentional signals. In general, an improper spectral shape indicates radar that is operating less than optimally. For example, Figure 13 shows a spectrum of a radar signal before and after adjustment to the cross-field amplifier used for the radar transmitter. The symmetry of the spectrum indicates optimally performing radar.

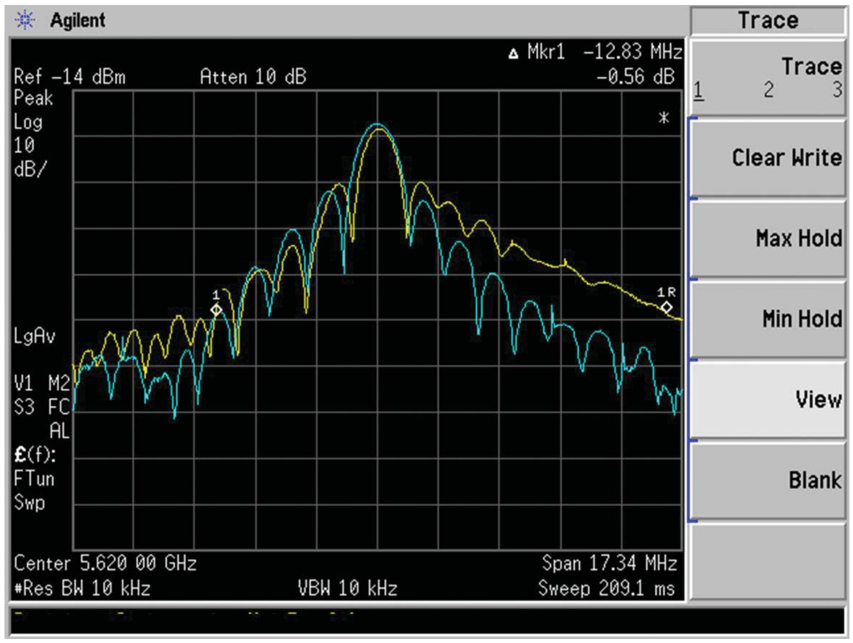


Figure 13. A spectrum analyzer is useful for examining the shape and symmetry of the radar spectrum. This example shows the spectrum trace before and after a timing adjustment in the magnetron.

Measuring pulsed radar with a spectrum analyzer is complicated by the different modes of operation that occur, which depend on the resolution bandwidth (RBW) setting of the spectrum analyzer. These variations exist when measuring any type of pulsed signal but tend to be more noteworthy when measuring low duty cycle pulses as is common with radar signals. Further, different types of spectrum analyzers, namely swept versus fast Fourier transform (FFT), can behave differently when measuring pulsed signals.

This section will start with a quick review of the basic spectral shape of a simple pulsed RF signal. Next, it will examine measuring radar with both a swept-based spectrum analyzer and FFT-based spectrum analyzer (including different measurement modes) and then conclude with a survey of the different built in measurement functions included in many of today's analyzers.

Pulse spectrum review

In the time-domain, multiplication of a continuous wave signal by a pulsed waveform results in a pulsed carrier. The spectrum of a pulsed signal forms a characteristic sinc function shape with main lobe and sidelobes. From a mathematical standpoint this can be understood by taking the Fourier transform of a rectangular waveform and then translating it to the frequency of the carrier.

As can be seen in Figure 14, the pulse width and PRF of the signal determine the characteristics of the basic pulsed spectrum. As the pulse width narrows the width of the spectrum and sidelobes broadens. The PRF of the pulsed RF waveform determines the spacing between each spectral component. Viewing the spectrum of the pulse can therefore provide meaningful information about the signal's pulse width, period, and duty cycle. For a basic pulsed RF signal, the duty cycle can then be used to calculate the peak pulse power from the average power and vice versa. A detailed discussion and derivation of the pulse spectrum can be found in the *Agilent Application Note 150-2, Spectrum Analysis...Pulsed RF,* literature number 5952-1039.

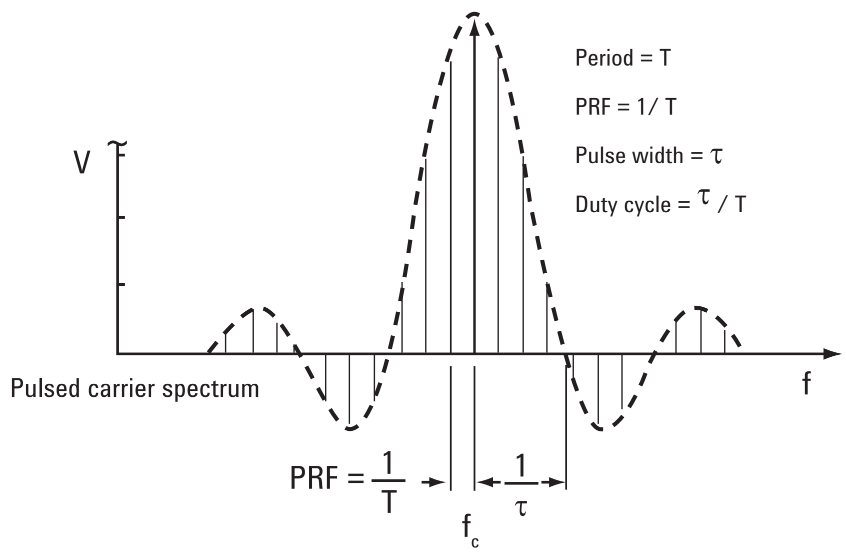


Figure 14. Pulsed spectrum

Pulsed RF measurements with a swept spectrum analyzer

Conventional spectrum analyzers are based on analog super-heterodyne swept architectures. Most modern spectrum analyzers such as the Agilent PSA Series high-performance spectrum analyzer or MXA midrange signal analyzer employ a digital implementation of a swept architecture that has many benefits in speed and accuracy over their analog counter parts. Other spectrum analyzers may work by calculating FFT. Still others, including the PSA and MXA, employ both techniques. Each has its advantages. For example, swept analyzers usually have the best dynamic range while FFT analyzers are likely faster for computing in-channel measurements. Other differences, as they relate to measuring pulsed radar signals, will be outlined below. One advantage of a swept architecture is that most RF designers are familiar with its operation. That familiarly results in an intuitive understanding of signals from their swept spectrum measurement that is lost in a snapshot FFT spectrum.

For swept spectrum analyzers there are three primary modes of operation. Line spectrum, pulse spectrum, and zero span.

Line spectrum mode

To accurately measure and view each spectral component, the resolution bandwidth filter chosen for the spectrum analyzer must have enough resolution to resolve each spectral component. The general rule is that the RBW < 0.3*PRF, as shown in Figure 15. When this condition is met the measurement is often referred to as measuring in the line-spectrum mode, the true spectrum of the signal.

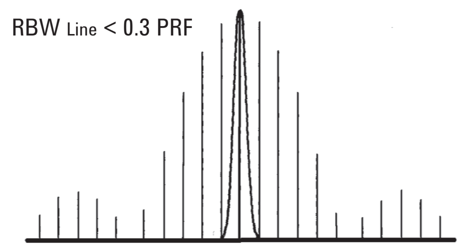


Figure 15. RBW setting on spectrum analyzer must be set less than PRF to resolve spectral components.

When viewing the spectrum, the power cannot be measured simply since the level of the signal is spread into its spectral components. However, the total power, meaning average power, can be measured using the band power marker functions or a built-in channel power measurement function of the analyzer. More information on these functions is given below. For a simple pulsed RF signal, however, the peak and average power can be extracted from the spectrum view. This is done by calculating what is known as the line spectrum desensitization factor. The peak power of the displayed spectrum is related to the peak power of the signal, assuming a near ideal pulse, by a factor of 20*log(duty cycle). The peak power can be determined by placing a marker on the central or highest power line in the measured spectrum and then adding, in dB, 20*log(duty cycle). The average power can then be determined from the peak power by subtracting the duty cycle, in log form, 10*log(duty cycle), as shown in Figure 16.

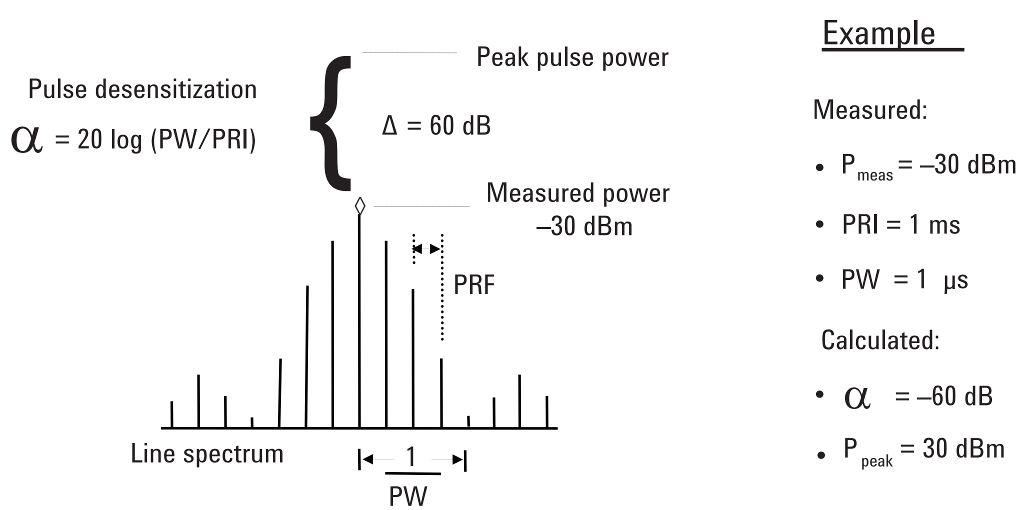


Figure 16. Calculating peak power from the pulsed RF spectrum (line mode) using the pulse desensitization factor.

If needed the duty cycle can be determined from the spectrum since the distance between the spectral lines is equal to the PRF and the distance between the nulls of the spectrums sinc function shape is equal to the inverse of the pulse width, as shown in Figure 16.

Note that the line spectrum mode is valid for all spectrum analyzers no matter if they are conventional analog swept, like the Agilent 8566B; swept using digital implementation, like the Agilent PSA; or use FFTs and have an equivalent RBW setting.

Pulse spectrum mode

The pulse spectrum mode applies specifically to spectrum analyzers that employ swept architectures. It is what occurs when the RBW setting on the spectrum analyzer is too wide to resolve the individual spectral components of a pulsed RF signal but not wide enough to contain the majority of the spectral power. Under this condition, the spectral components within the RBW filter at any one instance are added and displayed. If the sweep time is similar (or somewhat longer) to the pulse period, these lines, known as a PRF lines, will be displayed across the screen and have a similar sinc function shape as the line spectrum view. Note, however, that these lines are not spectral lines. Their location, though on frequency axis, has no specific frequency domain meaning and they will move around each sweep. A PRF line appears each time a pulse occurs. The space between each line is the space or distance in time that the analyzer sweeps between each pulse when no power is present at the spectrum analyzer input. Thus, in reality the spacing between the lines is related directly to the period of the signal. If the sweep time happens to be fast compared to the pulse period, the PRF lines will be spaced too far apart for a good view of the signal. In this case, the common practice is to increase the sweep time so that the PRF lines occur more closely together on the display and can be viewed.

Many spectrum analyzers include a marker feature that reads out the time difference between marker occurrences instead of frequency. Using this feature the pulse period can be measured directly, simply by computing a delta time marker between to adjacent PRF lines.

Two methods can be used to determine the operating mode of the analyzer: pulse-spectrum or line-spectrum. First, change the RBW. The amplitude of the displayed signal will not change if the analyzer is operating in the line-spectrum mode. If the analyzer is operating in the pulse-spectrum mode, the displayed amplitude will change because it is a function of RBW. Second, change the sweep time. The lines representing spectral components of the signal and will not change with sweep time in the line-spectrum mode. In pulse-spectrum mode, the spacing between PRF lines will change as a function of sweep time.

If the sweep time is set to be much longer than the pulse period, eventually the distance the sweep moves between pulses will become small relative to the resolution of the display. The resulting display with peak detector will show an outline of the pulse spectrum envelope, as shown in Figure 17. This display can be used to conveniently measure the pulse width by using minimum peak search function and measuring the spacing between the nulls as shown in Figure 17. This can be done because the side lobes are the same for the line-spectrum (true spectrum) and pulse-spectrum modes.

$$\alpha_{\rho} = 20 \cdot \mathrm{Log}\,(\tau \cdot B_{imp}) = 20 \cdot \mathrm{Log}\,(\tau \cdot K \cdot \mathrm{RBW})$$

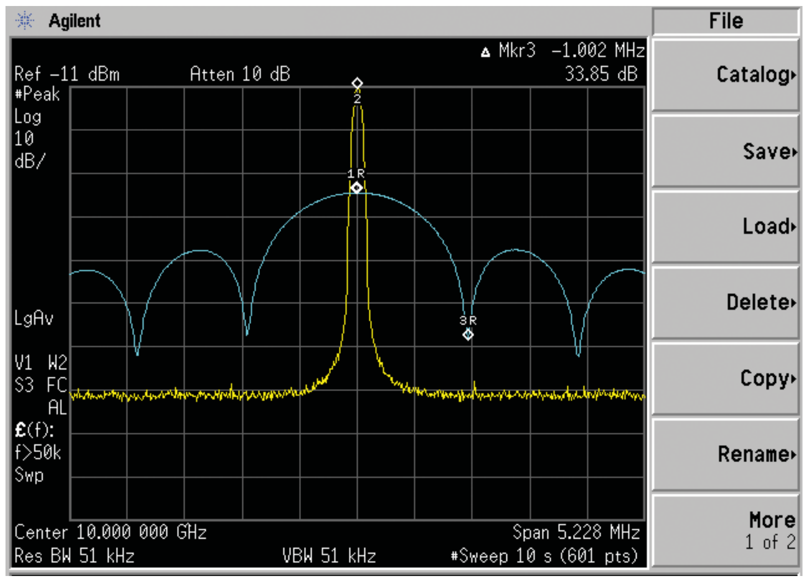


Figure 17. Pulse spectrum mode view: pulse modulation turned on and off.

Though the technique is used less often today, both the peak power and the average power can be determined from the pulse spectrum. For some corner cases where very narrow pulses are being used its possible that this technique may be the best viable option.

The maximum displayed level of the signal in the pulse spectrum mode is related to the impulse response of the spectrum analyzer's RBW filter. This will vary depending on the model and make. For example, the Agilent PSA spectrum analyzer's narrower RBWs are eight-element Gaussian filters with impulse bandwidth of 1.48 times the RBW. (B_{imp} = 1.48*RBW for RBW < 4 MHz) Using this information the peak power can be determined by calculating a desensitization factor and adding it to the measurement. The pulse mode desensitization factor is equal to 20*log(pulse width * B_{imp}). An example is shown in Figure 17.

Zero span measurements

In addition to making frequency-domain measurements, the spectrum analyzer provides the zero-span mode for time-domain measurements. In the zero-span mode, the spectrum analyzer becomes a fixed-tuned receiver with a time-domain display similar to an oscilloscope, except it displays the pulse envelope as shown in Figure 18. The modern spectrum analyzer also has various trigger modes to provide a stable display. Trigger delay controls allows you to position the pulse envelope for convenient measurements of pulse width, peak power, on-off ratio, and rise time.

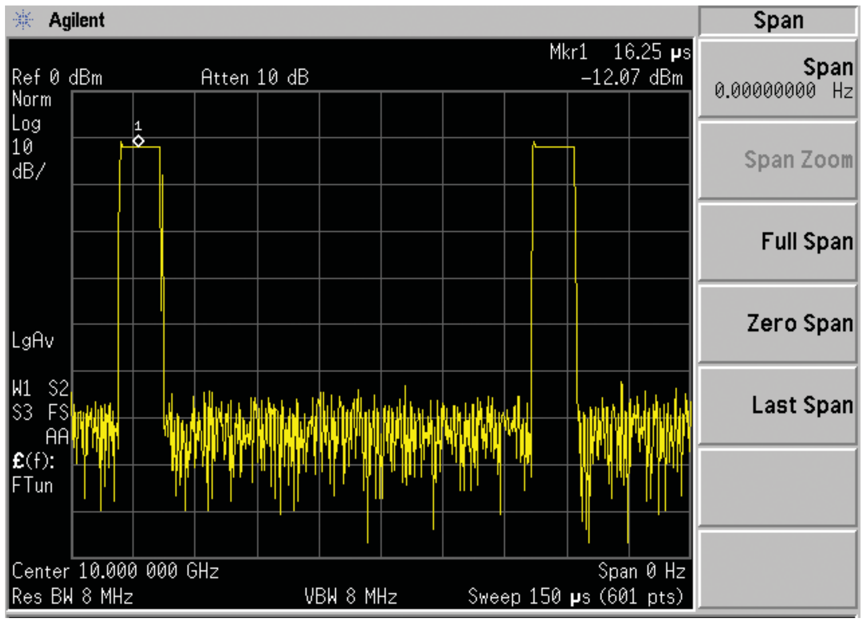


Figure 18. This example of zero span mode gives a time domain view of the envelope of the radar pulse. It is a convenient way of measuring pulse characteristics when the bandwidth of the radar signal is much less the RBW of the spectrum analyzer.

Many important pulse parameters such as rise time and amplitude droop that are impossible to measure in the frequency-domain are easily measured in zero-span. However, for a valid zero-span measurement, the RBW must be set such that all or at least a majority of the signal power is contained within the RBW. More specifically, to accurately measure the maximum power of a pulse the analyzer filter must be able to settle between measurements. The following condition must be met:

Pulse width > settling time of analyzer ≅ 2/RBW

(settling time = 2.56/RBW for the PSA and MXA signal analyzers)

To accurately measure the rise and fall time the analyzers settling time must be faster than the signal under test. A general rule is:

Rise time of pulse >> rise time of analyzer ≅ 0.7/RBW

In the case of the PSA spectrum analyzer the maximum resolution bandwidth is 8 MHz in the standard spectrum analysis mode. If a wider bandwidth is needed, two options exist for the PSA. The first is an optional fast rise time log video output. With this option a user can use an oscilloscope to analyze the video output signal from the PSA and measure rise times as fast as 15 ns.

The other option is to use the PSA vector signal analysis (VSA) mode and the 89601A VSA software. In the vector signal analysis mode, the PSA has an equivalent video bandwidth of 80 MHz and the ability to measure the time-domain envelope of the signal similar to zero span. Using this mode of measurement, the PSA can measure rise times of approximately 25 ns.

Pulsed RF measurements with spectrum analyzers that compute FFT

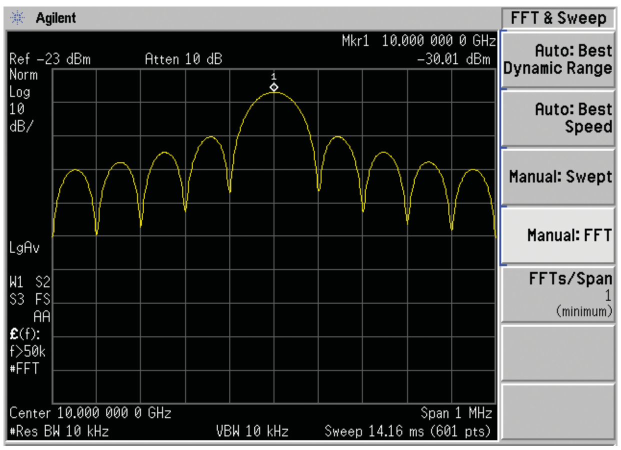
As mentioned above, some spectrum analyzers use FFT to compute the spectrum similar to the way that a vector signal analyzer does. One example of such an analyzer is the Agilent CSA spectrum analyzer. Analyzers that use FFT techniques have advantages and disadvantages when compared to swept analyzers. Swept analyzers have advantages in sensitivity and measuring over wide spans. FFT analyzers can be faster for measuring radars with bandwidths less than the analyzer's maximum FFT analysis bandwidth. FFT-based spectrum analyzers can also perform VSA measurements (if the software is implemented) since phase information is maintained. However, for reasons covered below, FFT analyzers tend to be inadequate for measuring wideband radar or radar with low duty cycles. Some spectrum analyzers like the PSA and MXA include both swept and FFT modes and automatically switch modes based on settings. The analyzer can be set to automatically optimize for speed or dynamic range or forced to remain in either FFT or swept mode.

FFT-based spectrum analyzers typically have a user interface designed to have a similar look and feel as a traditional swept analyzer. A casual user may not even recognize that they are using FFT techniques rather than traditional spectrum sweeps. However, the differences become apparent when measuring pulsed RF signals.

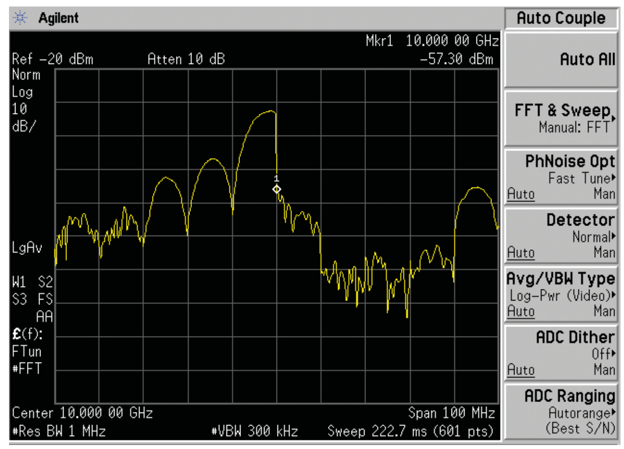
Like a vector signal analyzer, the FFT-based spectrum analyzer can be fast at measuring signals whose entire bandwidth is contained within a single FFT (i.e. contained within its analysis bandwidth). With this condition met the FFT spectrum analyzer is essentially equivalent to a VSA though typically without as many measurement functions and displays. See more information about how a VSA works in the VSA section below.

When the span of interest is wider than the analysis or FFT bandwidth of the analyzer, the FFT-based spectrum analyzer calculates the spectrum by taking multiple FFTs at different frequencies and concatenating the results. Sometimes this technique is referred to stitching since the analyzer computes the spectrum one section at a time, step tuning to a different frequency each section, and patching them together. Depending on the speed of the analyzer one may be able to see each section of the spectrum appear as it is computed as opposed to a smooth sweeping action of a swept analyzer.

If the condition for line-spectrum mode is met (PRF < 0.3 RBW) then there is no difference in the result for the traditional swept or FFT analyzer. However, if these conditions are not met, the FFT-based analyzers will behave quite differently than the swept analyzers. In this case, the FFT analyzer will not display PRF lines as occurs in the swept analyzer. Rather the data displayed will depend on the probability of intercept between the FFT acquisitions and the pulses. Because FFT analyzers require time to retune and capture when measuring over wide spans it may have difficulty intercepting low duty cycle pulses resulting in a blank or a sporadic measurement result. When the speed of the analyzer and duty cycle of the pulse are such that the signal is intercepted then the result may appear as disjointed segments of the spectrum envelope, as shown in Figure 19.



FFT analyzer will display pulse envelope rather than PRF lines as measured on a swept analyzer.



The measurement result may be sporadic when measuring wide band radar signals because the FFT analyzer is required to retune between FFT calculations.

Figure 19. A spectrum analyzer that computes FFTs will behave differently than a swept analyzer when the RBW is greater than PRF. Instead of PRF lines the spectrum analyzer will display the pulse envelope.

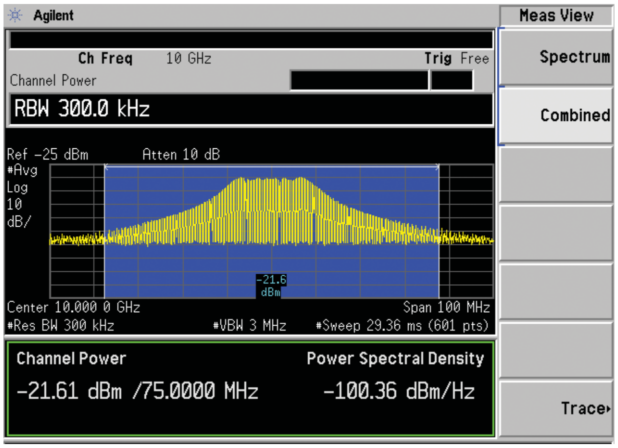
In the best case, the complete RF spectrum envelope will be displayed rather than PRF lines. Some spectrum analyzers using FFT may include settings such as FFT dwell time, FFT length, or triggering to improve the ability to capture and measure the pulses. These settings may be tied to a sweep time control to mimic the user interface of a traditional analyzer but in actuality it's the FFT length, dwell time, or number of averages that is being modified since the analyzer is not sweeping in a traditional sense. Generally, FFT-based analyzers are not optimum for measuring radar signals whose bandwidth requirements are beyond the FFT analysis bandwidth of the analyzer.

Built-in measurements

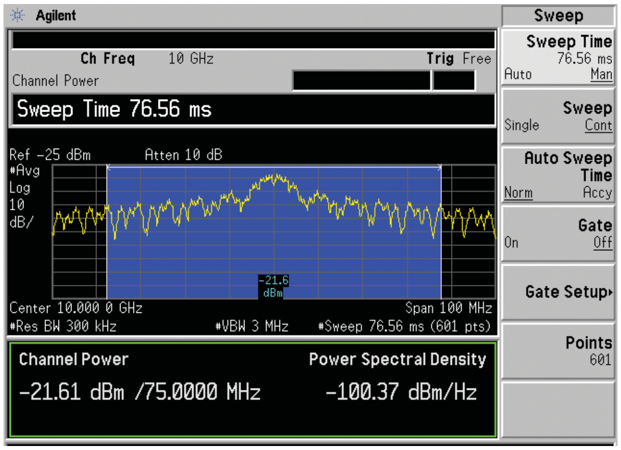
Today's modern spectrum analyzers include many built in functions and capabilities that can simplify and enhance radar measurements. Several of these features are highlighted here.

Channel power

The channel power function is designed to measure the average power across a given frequency band. It is a common measurement that is frequently used to measure many different types of signals. Spectrum analyzers use different techniques for making channel power measurements. The most common and typically the most accurate way is using the integration bandwidth method. The analyzer essentially integrates the power as it sweeps across the given integration bandwidth. Typically the measurement uses the analyzers averaging detector. For the best accuracy the RBW should be small compared to the integration bandwidth. However, it does not matter if the conditions for line spectrum or pulse spectrum are met. Examples of channel power measurements of a radar signal are shown in Figure 20.



Chirped radar



8 bit Barker code

Figure 20. Channel power measurement of radar signals performed on a spectrum analyzer. Results are equivalent to average power.

The channel power measurement can be especially useful for modulated, chirped, or pulses that are more complex and vary their PRF or pulse width. The spectrums of these signals are more complex and as a result, the power can not be as easily derived from the spectrum as was explained above for a simple RF pulse. Figure 20 shows a channel power measurement of chirped radar and pulse-coded radar. Note that the spectral shape is no longer a simple sinc function shape as the modulation on the pulse dominates in shaping the spectrum.

Occupied bandwidth

The occupied bandwidth (OWB) measurement automatically calculates the bandwidth in which a specified percentage of the power is contained. Often the OBW of the signal is determined based on the bandwidth for which 99% of the signals power is contained, as shown in Figure 21.

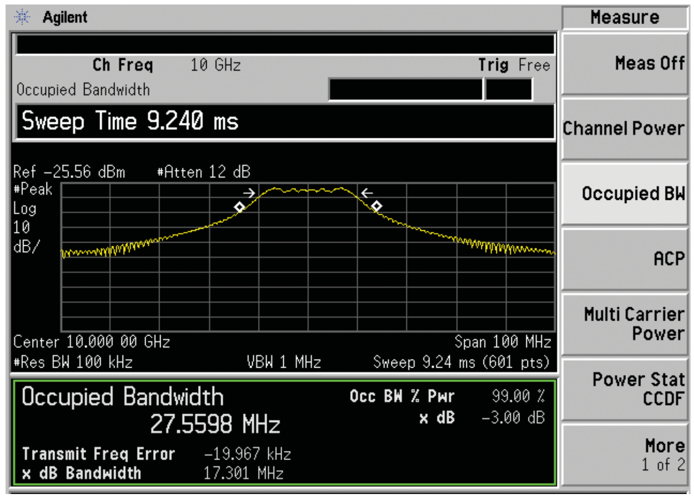


Figure 21. This is an example of an occupied bandwidth measurement on the PSA spectrum analyzer. Bandwidth for which 99.00% of the signal power is contained is automatically measured and reported.

Burst power

The burst power measurement is an automated zero span measurement. Rather than integrating the power in the frequency-domain as is done in the channel power measurement, the burst power measurement integrates the power across a defined time slot or gate and is essentially equivalent to the gated power measurements discussed earlier on the power meter. Often this measurement uses a burst power trigger, provided on some spectrum analyzers like the PSA, that automatically finds and triggers on the burst (or pulse), as shown in Figure 22. This measurement can be very useful for radar as it is a direct measurement of the pulse power. Its limitations, however, are the same as they are for zero span: the resolution bandwidth filter must be wide relative to the occupied bandwidth of the signal.

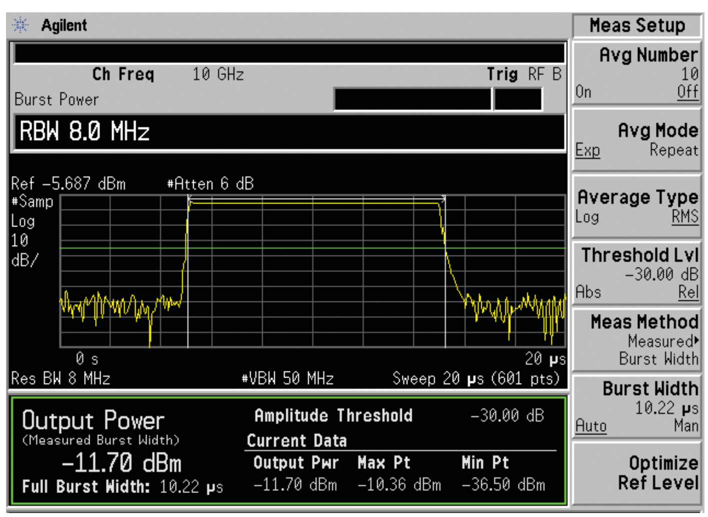


Figure 22. This is an example of burst power measurement on PSA spectrum analyzer. Burst power and pulse width are automatically measured in zero span mode.

Measuring with a vector signal analyzer

Unlike a spectrum analyzer, a vector signal analyzer captures the phase and magnitude information of the measured signal and uses this information to perform more advanced analysis. Vector signal analyzers are typically very flexible and can display results in the time-, frequency-, and the modulation-domain. (For detailed information on how a vector signal analyzer operates see *Agilent Application Note 150-15, Vector Signal Analyzer Basics,* literature number 5989-1121EN.) [6]

A vector signal analyzer does not sweep across a wide frequency range like a spectrum analyzer. Most vector signal analyzers operate by tuning to a specific frequency, conditioning the signal, down converting, digitizing, and processing the signal. Some vector signal analyzers skip the analog down conversion stage and directly digitize the baseband, IF or even RF signal after conditioning.

The primary limitation of a vector signal analyzer is its analysis bandwidth (sometimes referred to as information bandwidth or FFT bandwidth). To properly analyze a signal virtually all of the signal power must be contained within the analysis bandwidth of the instrument. The analysis bandwidth of a vector signal analyzer is usually dictated by the analog-to-digital converter (ADC) sampling rate and Nyquist law. The dynamic range limit of the analyzer is usually limited by the bit level of the ADC though the effective bits may vary noticeably between instruments with the same bit depth depending on the quality of the ADC, sophistication of dithering techniques, image correction, and over-sampling techniques.

There are a variety of vector signal analysis solutions available with varying performance restraints and trade-offs in bandwidth, sensitivity, memory, and frequency range. (See Agilent product overview: *Hardware Measurement Platforms for the Agilent 89600 Series Vector Signal Analysis Software,* Literature number 5989-1753EN). Many of today's modern instruments include dual functionality and operate as both a spectrum analyzer and a vector signal analyzer, as is the case with the Agilent PSA Series and MXA, or as an oscilloscope and vector signal analyzer as with the Agilent VSA80000 Series. Agilent VSA software can also be extended to work with logic analyzers to analyze signals in digital form or to work in software simulation environments such as Agilent ADS or MATLAB®.

In the section below, how the vector signal analyzer measures the power spectrum of a pulsed RF signals is explained, and then other useful VSA measurements, such as time-domain characteristics, time-gated measurements, gap-free or live measurements, and spectrogram, are discussed. Examples of measuring the chirp linearity and pulse coding of more complex radar waveforms are also given. A more in-depth discussion of advanced VSA measurements for radar can also be found in Agilent 2004 Aerospace Defense Symposium paper, *Advanced Pulse Stability, Clutter Cancellation Ratio, and Impairment Testing Using a Vector Signal Analyzer as a Flexible Ideal Receiver.* (Included on the Agilent A/D Symposium CD, literature number 5989-6075EN.) [7]

Measuring the radar spectrum with a vector signal analyzer

A vector signal analyzer calculates the spectrum by performing an FFT. The analyzer will calculate the FFT of whatever portion of the signal is within its measurement time window. Therefore, a best practice to achieve good spectrum results of a pulsed RF signals is to include multiple pulses within the measurement window so that the FFT calculation captures the repetitive characteristics of the signal, as shown in Figure 23. However, one of the attractive capabilities of a vector signal analyzer is that it can perform single shot analysis. This can be very useful for radar since this gives the analyzer the capability of analyzing a single pulse or transient events, as in Figure 24. To be clear, however, the nature of the FFT calculation should be understood to fully appreciate the results that the analyzer displays.

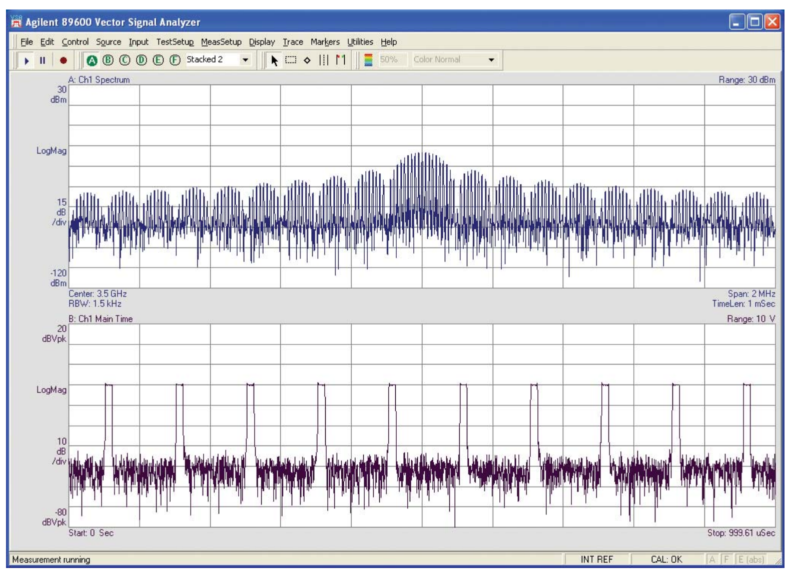


Figure 23. Vector signal analyzer measurement of radar pulse showing spectrum and time domain view.

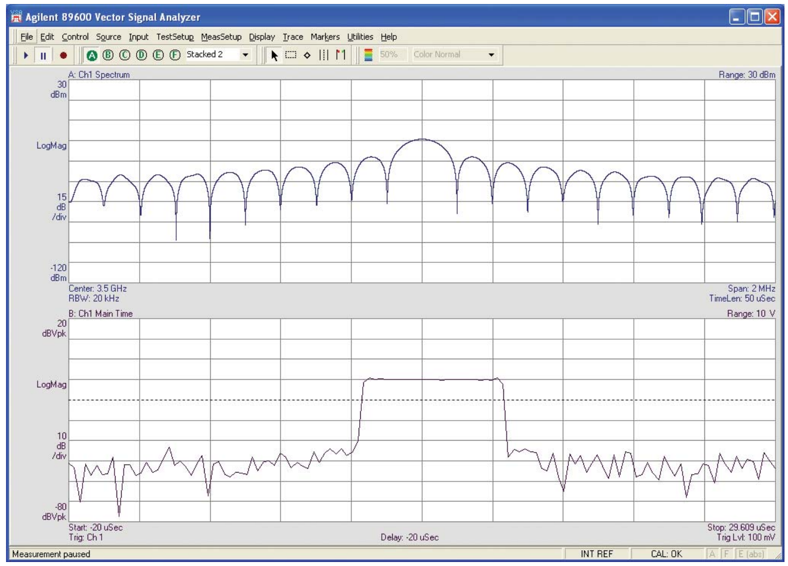


Figure 24. A vector signal analyzer has the ability to provide spectrum analysis of a single pulse.

Inherent in an FFT is the assumption that the time-domain signal used in the calculation repeats. If you capture a single time event and use it to calculate an FFT the resulting spectrum result is really the spectrum of a repeating version of the event, as shown in Figure 25.

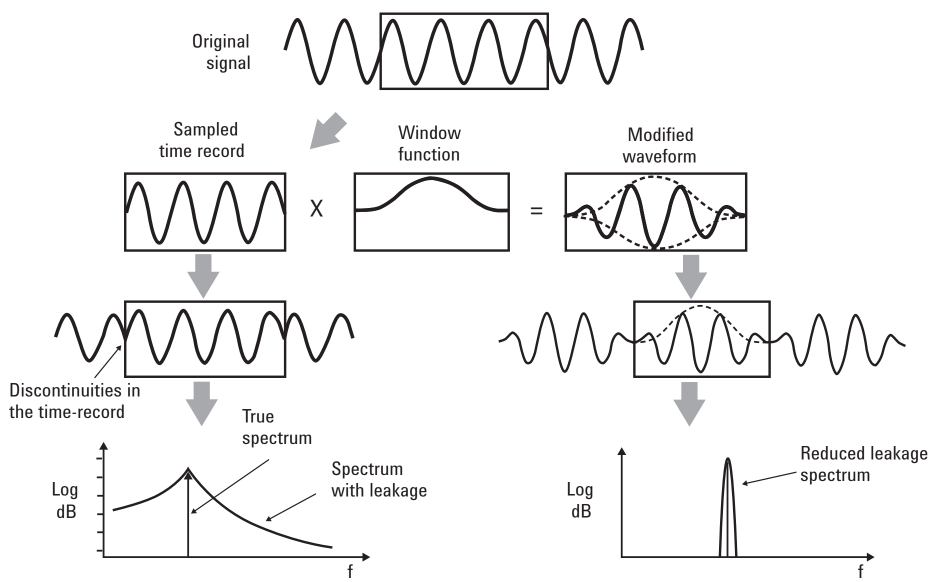


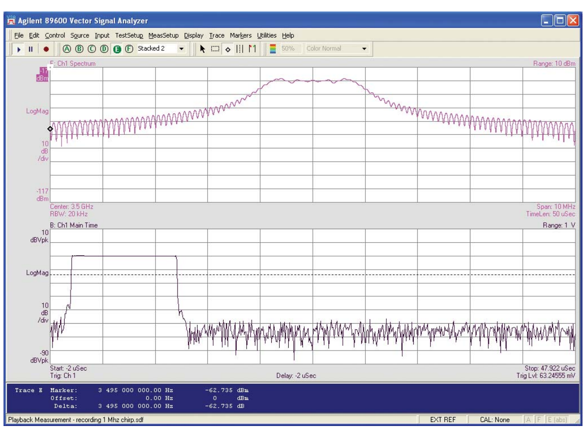
Figure 25. FFT analysis done in a spectrum analyzer or vector signal analyzer assumes a repeating signal. To minimize effects of discontinuities in the signal a windowing function is used.

Vector signal analyzer windowing function

With a vector signal analyzer each measurement update captures a portion of the time-domain signal and uses it to compute the FFT. A inherent problem is that the successive time captures may start and end at different voltage-level points in the signal. Since the FFT assumes that the signal repeats, a discontinuity (sudden change in voltage) results. This discontinuity will show up as leakage in the spectrum. To mitigate this problem, vector signal analyzers utilize a windowing function that operates on the waveform to minimize the discontinuity. (See Agilent *Application Note 150-15, Vector Signal Analysis Basics*, literature number 5989-1121EN for detailed explanation.) [6] Even though this windowing function changes the shape of the time-domain waveform, its shape (typically Gaussian or Hanning window) is designed to have minimal effect on the spectral result. However, it can have some impact on the accuracy of the spectrum. Different types of windowing functions are typically made available in order to trade off performance objectives in frequency accuracy, amplitude accuracy, or sensitivity.

Pulsed or bursted signals have the benefit that the start and stop of the measurement window can usually be set to occur in between the pulses when no signal is present thus avoiding discontinuities. Most radar signals, as it turns out, have an added advantage in that they tend to be self windowing due to the symmetry of the signals. The best results for radar signals are therefore achieved by using the uniform window function that does not alter the time-domain waveform.

Figure 26 shows an example of a chirp radar spectrum with two different windowing functions. The Gaussian window function in this case actually distorts the spectrum since the pulse is occurring near the beginning of the measurement time window where the windowing function has the most effect on the signal. The same measurement using a uniform window gives a much better result.



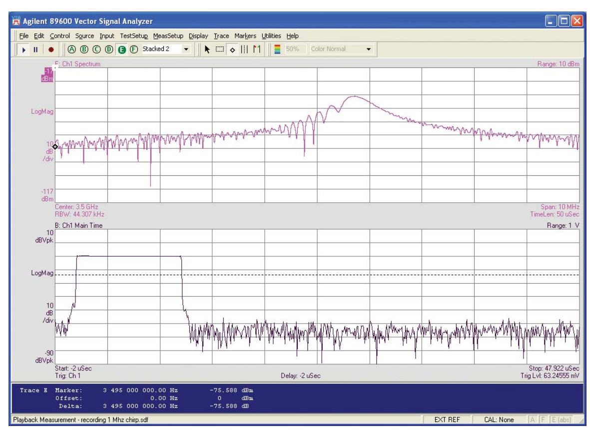


Figure 26. This compares a radar measurement made on a VSA using different windowing functions, uniform (left) and Gaussian (right). The Gaussian windowing function distorts the spectrum since the pulse occurs near the start of the measurement time window. Since pulsed radar signals are self-windowing the uniform window function should be used for best accuracy.

Agilent 89601A Vector Signal Analysis Software

Flexible signal analysis of frequency, time and phase

- Supports multiple platforms including PSA & MXA
- Ultra-wide bandwidth analysis with 80000-series oscilloscope
- Gap-free analysis with record playback feature
- Spectrogram

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Measuring power and pulse characteristics with a vector signal analyzer
Similar to a spectrum analyzer the average power and peak power can be determined from the spectrum result. However, the right measurement conditions must be met and understood to ensure an accurate measurement. The most straightforward way to ensure accurate results is to set the analyzer to measure in the line spectrum mode with the RBW less than 0.3*PRF. This condition ensures that the analyzer resolves each spectral component. Note that since the vector signal analyzers RBW setting is tied to the measurement time, setting the RBW to meet line spectrum mode will automatically result in a longer measurement time essentially having the same effect as including several pulses in the FFT calculation as shown in Figure 23. The average power and peak power for a simple pulsed RF signal can be determined using the same method used in the line spectrum mode on the spectrum analyzer. This is not normally done however since with a vector signal analyzer other measurement windows can be used to make these measurements directly in the time-domain view.

Band power (similar to channel power) and OBW functions are also typically included with the vector signal analyzer and have the same capability as the equivalent functions on a spectrum analyzer. Additional measurements such as complementary cumulative distribution function (CCDF), spectrogram, and time-gated spectrum are also available. To achieve accurate results the spectrum results need to be calculated from a representative time-domain sample of the waveform. If only a small sample of the signal relative to the period is being used to calculate the spectrum, the displayed level of the spectrum may not reflect the true power level or correct spectral characteristics of the signal. This can be resolved by either increasing the FFT measurement time to include several pulse periods or setting the RBW to meet the line spectrum mode.

Measuring time-domain characteristics with a vector signal analyzer

When measuring radar signals with a vector signal analyzer it is intuitive to measure signal characteristics like average power, peak power, pulse power, duty cycle, pulse width, pulse period, and pulse shape in the time-domain.

With the 89601A VSA software time-domain and frequency-domain measurements can be displayed simultaneously. This makes it easy to correlate the time-domain and frequency-domain views of the signal.

A variety of time-domain display options typically exist for a vector signal analyzer. Figure 27 shows the display of a radar chirp using the PSA Series spectrum analyzer with 89601A VSA software. Time-domain displays and markers can be used to readily measure pulse power, peak power, pulse width, pulse period, rise time, frequency versus time or group delay, pulse to pulse phase continuity, or pulse droop.

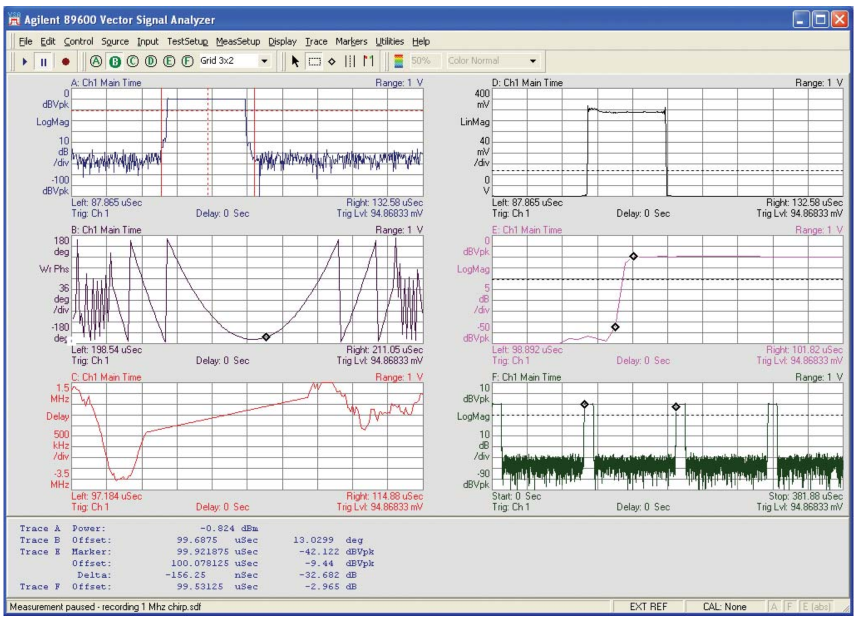


Figure 27. Time domain measurements on chirp radar using a vector signal analyzer.

Time-gated analysis of pulsed RF signal on vector signal analyzer
Many vector signal analyzers have the ability to perform time-gated analysis. The technique that these analyzers use is FFT time gating where only the samples collected during a defined timeslot is used in the FFT calculation. The unique advantage to the vector signal analyzer is that it is able to perform single shot time-gated analysis. This is useful to examine the spectral characteristics of an individual pulse, portion of a pulse, or transient. The time-gated measurements on the 89601A VSA software are convenient, intuitive, and performed simply by dragging and dropping the gate onto the time-domain display of the waveform. Figure 28 shows the time-gated spectrum of a simple RF pulsed radar. Remember that the spectrum displayed is in reality the spectrum of a repeating version of the waveform capture in the time record gate.

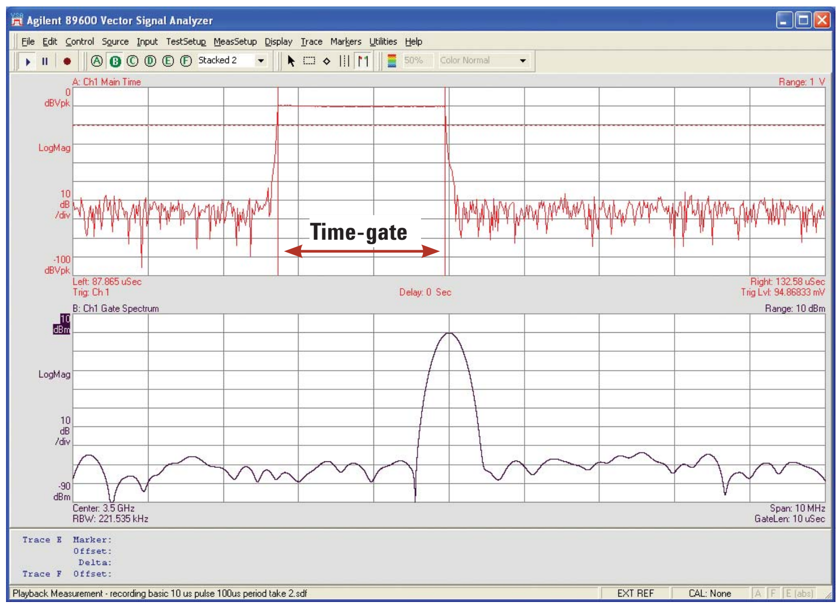


Figure 28. This is a time-gated measurement of simple RF pulse showing CW spectrum of signal when pulse is on.

Spectrogram

Spectrograms are made up of a sequence of ordinary spectrum measurements, where each spectrum measurement is compressed to a height of 1 pixel row on the display and the amplitude values of the spectra are encoded as color. This produces a display of spectrum versus time, containing hundreds or even thousands of spectrum measurements.

The spectrogram allows easy visual recognition of the major signal characteristics. Figure 29 shows an example of measuring a frequency agile signal with the Agilent VSA80000. The measurement is useful in characterizing the frequency behavior of signals over time.

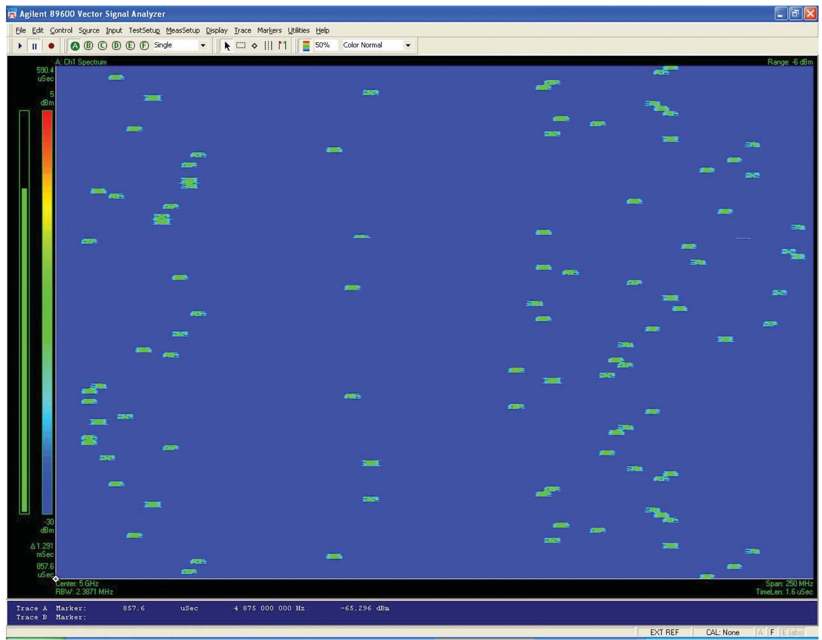


Figure 29. Spectrogram showing frequency agile signal over 1 GHz span, made with Agilent's VSA80000 ultrawideband vector signal analyzer.

Gap-free or live measurements

Another useful feature of a vector signal analyzer is that it is capable of performing gap-free analysis. This is sometimes referred to as a live measurement since all the data within the analysis bandwidth of the instrument and time record is being captured and analyzed.

A gap free or live measurement is similar to a real-time measurement. Sometimes the term "real-time" is used loosely to describe a live measurement. A true definition of the term real-time would mean that all data within the analyzer information bandwidth is acquired, processed and displayed continuously without gaps. By this definition, real-time operation of FFT-based spectrum analyzers and vector signal analyzers is limited to some low bandwidth. (Note that the U.S. government imposes export restrictions on real-time capabilities over 500 kHz.) For example, the Agilent 89400 vector signal analyzer specifies a real time bandwidth of 78 kHz. However, it is possible to perform gap-free analysis of wider bandwidth signals, such as radar, by recording and then post processing the input signal.

A vector signal analyzer with the capabilities like the 89601A VSA software performs gap-free analysis by taking advantage of the record-playback feature. By recording the signal into memory it can be played back at a slower rate and analyzed without skipping any data. Figure 30 shows a gap-free spectrogram measurement view of a chirped radar signal. The spectrogram shows the linearly changing frequency of the chirp and the pulse widths of the signal. In this example the FFTs are set to overlap to show more detail of the signal spectral content.

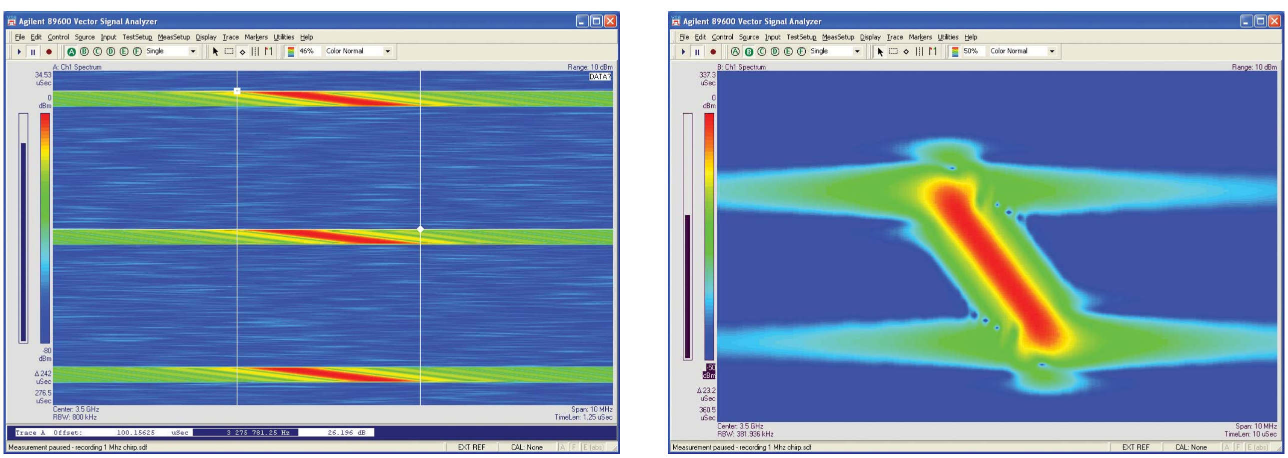


Figure 30. Spectrogram showing chirped radar signal. The left view shows the spectrogram of three consecutive pulses. The right view shows detailed view of a single pulse.

Power statistics — complementary cumulative distribution function (CCDF)
The CCDF measurement is used to measure the statistical power characteristics of a signal. It calculates the peak-to-average ratio (peak-to-average is equivalent to the duty cycle for basic pulses) and plots the power on a graph that shows power in dB above average power on the *x* axis and percentage of time on the *y* axis. With the measurement result, the percentage of time the signal is at a specified power level above the average power can be determined. The CCDF measurement is especially useful for determining the power characteristics of shaped pulses or the peak-to-average (duty cycle) of radar signals that vary their PRF.

Figure 31 shows an example of the CCDF measurement for simple pulse and for a shaped pulse. The peak-to-average ratio of the simple pulse is equivalent to the duty cycle. The shaped pulse shows a more gradual transition in power levels.

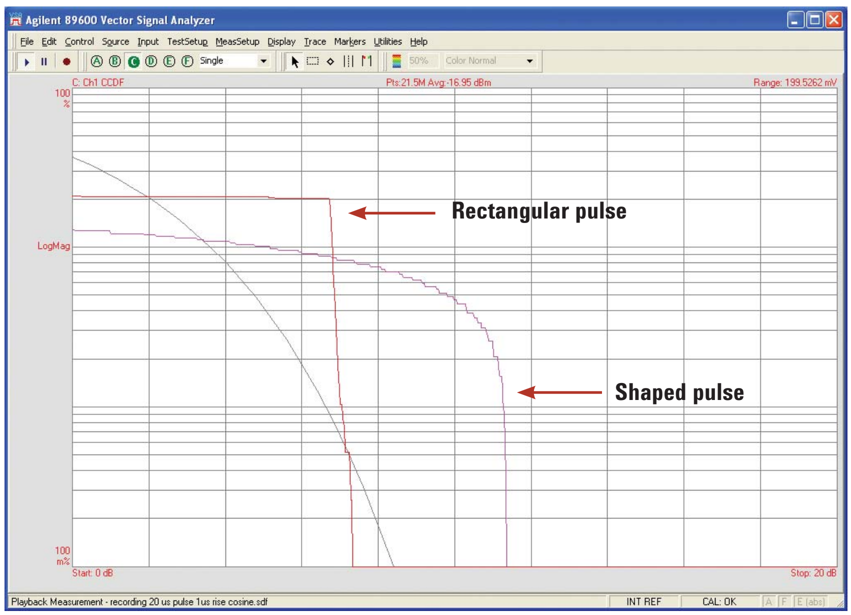


Figure 31. CCDF measurement of a square pulse radar signal compared to a raised cosine shaped pulse radar signal. A CCDF plot describes the power statistics of a signal by plotting the percentage of time (Y axis) the signal spends at x dB above the average power (X axis).

Chirp linearity

Since the vector signal analyzer is able to analyze both phase and magnitude of a signal it is a useful tool for measuring modulation on a pulse such as chirp. A common measurement to make on a chirp radar signal is to examine the group delay. Since group delay is directly proportional to the frequency change, the result is an analysis of the chirps frequency modulation. An example of group delay measurement of a linear chirp is shown in the bottom left view of Figure 27 on page 38.

Pulse-coded modulation

The ability to view phase as a function of time is a useful feature for analyzing the modulation on a pulse for a coded radar pulse. Typically this coding is modulated onto the pulse by modulating the phase. Figure 32 shows an example of a coded pulse measurement. Notice the 180° phase shift that can be seen in the phase versus time display.

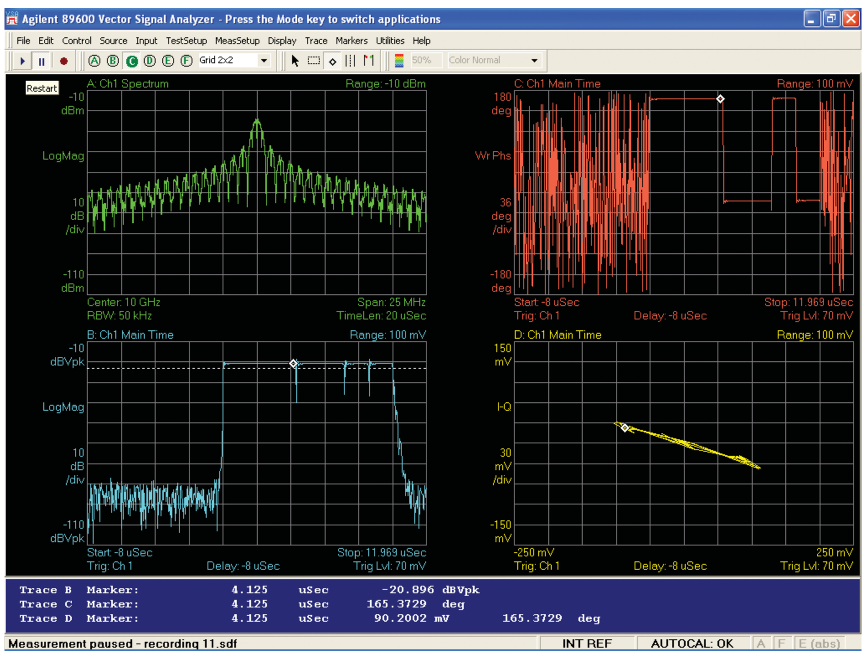


Figure 32. A vector signal analyzer has the ability to examine phase variations over time. This is a useful feature for examining a Barker coded pulse with code embedded in the phase transitions.

Component and Subassembly Test

Maximizing radar performance requires thorough analysis and the optimization of each subassembly and component in the radar system.

The effects of losses of the signal caused by system components in the transmit path is directly characterized by the L_T and L_R terms in the radar range equation. Power can be expensive and losses directly subtract off the effective power of the radar. A 1 dB loss has the same impact as a 1 dB reduction in power. When transmitting 1 megawatt of power a one dB increase to accommodate for a loss can be expensive. The more losses can minimized, the better. Loss measurements are especially important for components such as filters, duplexers, and circulators that are located after the transmit power amplifier and before the low noise amplifier of the receiver.

Though not shown directly in the radar range equation the phase and amplitude flatness and group delay of components also affect the performance of the radar. These impairments have an indirect impact on the range by limiting the radar's ability to optimally compress the signal with a matched filter. In addition, these impairments may limit the radar's ability to extract other information from the radar returns such as Doppler information.

A common way to characterize component effects including loss, flatness and group delay is utilizing S-parameter measurements. S-parameters describe the incident and reflected response of a component in complex terms and paint a comprehensive picture of the linear effects that the component is having on the signal. These measurements are typically made with a network analyzer using a CW stimulus.

With radar however, it is often not sufficient to make the measurements in a conventional way using continuous wave signals. This is due to several reasons.

One reason is that the performance of components may differ when tested under pulsed conditions compared to using continuous wave stimulus due to different bias conditions, ringing effects cause by fast rising edges, or different operating temperatures. Another reason may simply be the fact that a device may not be designed to handle the power dissipation associated with a continuous stimulus.

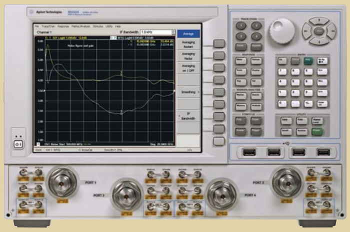
In this section of the paper, the various approaches to measuring components and subassemblies will be presented.

Direct measurement of power loss with a power meter

A simple way to measure losses is to use a power meter. Power meters like the Agilent P-Series, EPM-P, or EPM Series have dual port model options to allow simultaneous measurements provided measurement ports are available before and after the element being tested. By making a measurement before and after the device, the power difference can be determined.

The accuracy of measuring losses with a power meter is typically less than what can be achieved with a calibrated vector network analyzer; however, the advantage is that the measurement can be made under normal operating conditions.

PNA-X Network Analyzer



Fast accurate passive and active device characterization

- Pulsed S-parameter test
- Single connection active device test including noise figure
- Error-corrected frequency converter test
- 2-, 4- & N-port configurations
- Single-ended and balanced measurements with true balanced stimulus
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This may be important for measuring in the transmit path since power levels may be much higher than can be provided by a signal generator or network analyzer used in a stimulus response test setup. Another point that should be noted is that the measurements results from a power meter may not directly correlate with the results from using other instruments such as network analyzers since power meters are broadband detectors and will measure fundamental and harmonic power as opposed to the tuned receiver measurements of a network analyzer that isolates the fundamental signal component.

Measuring with a network analyzer

Network analyzer measurement modes

Before discussing the specific measurements made with a network analyzer it is necessary to first discuss the different modes of measuring pulsed RF signals.

Network analyzers can measure pulsed RF signals using either a wideband (or synchronous acquisition) detection mode or a narrowband (asynchronous acquisition) detection mode. A summary of these modes is given here, (A more detailed discussion can be found in *Agilent Application Note 1408-12, Pulsed-RF S-Parameter Measurements Using Wideband and Narrowband Detection*, literature number 5989-4839EN.) [8]

Wideband detection mode

Wideband detection can be used when the majority of the pulsed-RF spectrum is within the bandwidth of the receiver. It can be thought of as being analogous to the zero span mode of the spectrum analyzer, as shown in Figure 33. In this case, the pulsed-RF signal will be demodulated in the instrument, producing baseband pulses, again similar to zero span in the spectrum analyzer. This detection can be accomplished with analog circuitry or with digital-signal processing (DSP) techniques. With wideband detection, the analyzer is synchronized with the pulse stream, and data acquisition only occurs when the pulse is in the on state. This means that a pulse trigger that is synchronized to the PRF must be present, and for this reason, this technique is also called synchronous acquisition mode.

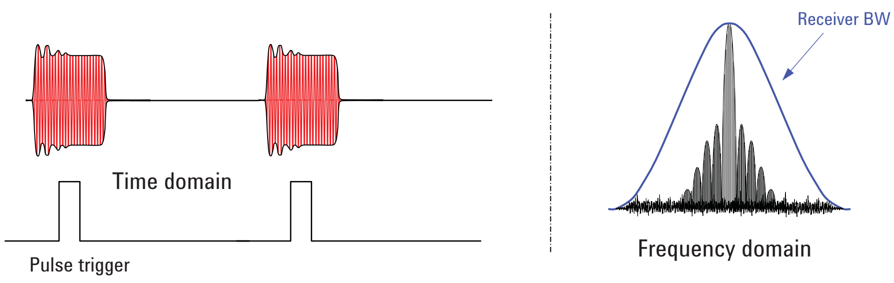


Figure 33. Wideband detection mode (synchronous detection mode) of a network analyzer requires the majority of the pulse power be contained within the receive bandwidth of the network analyzer. Wideband mode may have the best dynamic range for signals with low duty cycle but will be limited in measurable pulse widths due to bandwidth constraints.

The advantage of the wideband mode is that there is no loss in dynamic range when the pulses have a low duty cycle (long time between pulses). The measurement might take longer, but since the analyzer is always sampling when the pulse is on, the S/N ratio is essentially constant versus duty cycle. The disadvantage of this technique is that there is a lower limit to measurable pulse widths. As the pulse width gets smaller, the spectral energy spreads out — once enough of the energy is outside the bandwidth of the receiver, the instrument cannot detect the pulses properly. Another way to think about it in the time-domain is that when the pulses are shorter than the rise time of the receiver, they cannot be detected. In the case of the Agilent PNA-X network analyzer the bandwidth limit is 5 MHz, which corresponds to a minimum pulse width of approximately 250 ns.

Narrowband detection mode

In narrowband detection mode the receiver bandwidth of the network analyzer is set so that all of the signal power is filtered out except for the central spectral component, as shown in Figure 34. Narrowband detection mode is analogous to the line-spectrum mode of the spectrum analyzer except the analyzer stays tuned to one specific frequency. The center spectral component of the signal represents the frequency of the RF carrier. After filtering, the pulsed RF signal appears as a sinusoid or CW signal. With narrowband detection, the analyzer samples are not synchronized with the incoming pulses (so no pulse trigger is required), thus the technique is also called asynchronous acquisition mode.

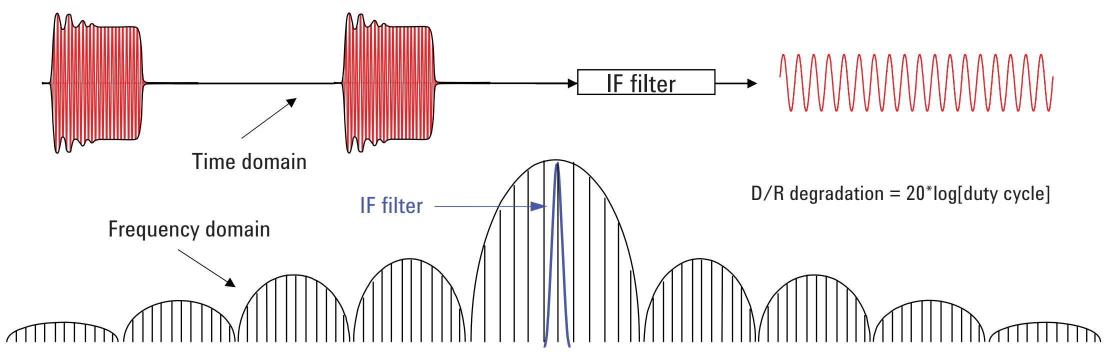


Figure 34. The narrowband detection mode (asynchronous detection mode) of a network analyzer uses a narrow filter to extract only the central spectral component. The narrowband mode has no pulse width limitations.

Agilent has developed a novel way of achieving narrowband detection using wider IF bandwidths than normal and a unique spectral-nulling technique. This technique lets the user trade dynamic range for speed, with the result almost always yielding faster measurements than those obtained without the technique.

The advantage to narrowband detection is that there is no lower pulse-width limit, since no matter how broad the pulse spectrum is most of it is filtered away anyway, leaving only the central spectral component. The disadvantage to narrowband detection is that measurement dynamic range is a function of duty cycle. As the duty cycle of the pulses gets smaller (longer time between pulses), the average power of the pulses gets smaller, resulting in less S/N ratio. In this way, measurement dynamic range decreases as duty cycle decreases. This phenomenon is often called pulse desensitization. The degradation in dynamic range (in dB) can be expressed as 20*log(duty cycle). Some of this degradation can be overcome with sophisticated signal processing. The PNA-X for example has a 40 dB improvement in dynamic range compared to a PNA when using a stimulus with a .001% duty cycle.

Network analyzer measurements

There are a number of measurements and display functions the network analyzer can make to describe the behavior of a component or subassembly. These include insertion loss, group delay, S-parameters, and variations of these same parameters over time. Figure 35 shows an example of an insertion loss measurement made in both the narrowband detection mode and the wideband detection mode on the PNA-X network analyzer. For a pulsed RF signal with pulse width of 1 us and PRI of 100 us the wideband detection mode makes the measurement 17 times faster. Because of the speed advantage the wideband mode is generally preferred when possible. However, as previously explained the narrowband mode may be the only measurement option for narrow pulses with wide bandwidth.

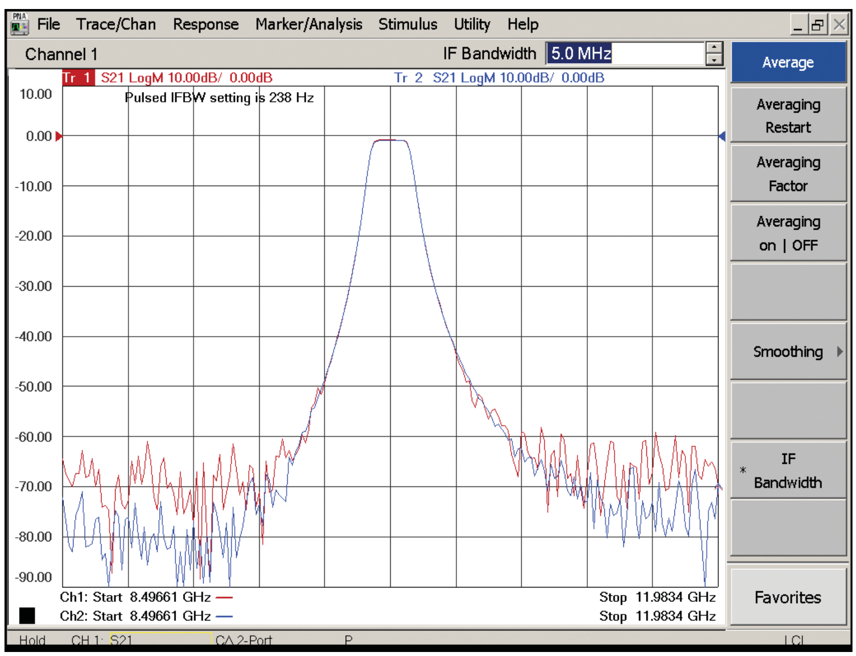


Figure 35. The display shows a comparison of a pulsed insertion loss filter measurement using the wideband mode versus the narrowband mode in the PNA-X network analyzer. In this example, the narrowband mode achieves the best dynamic range; however, which mode is best will depend on the duty cycle.

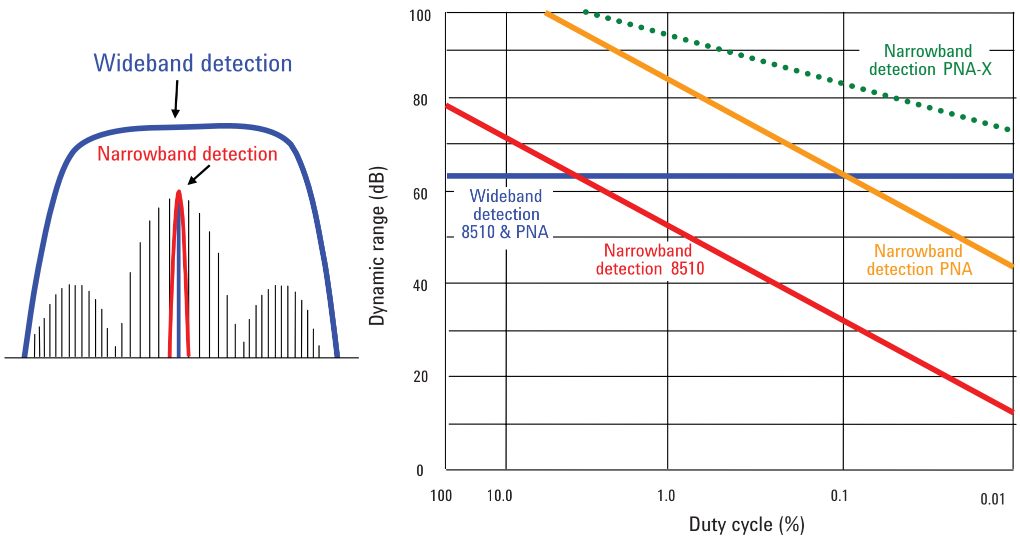


Figure 36. The effects of duty cycle on dynamic range are shown by comparing network analyzer wideband detection and narrowband detection modes.

In terms of dynamic range the advantage depends on the instrument and will be a function of the duty cycle. As can be seen from Figure 36, the PNA-X has exceptional dynamic range performance for the narrowband mode based on the advanced processing techniques it employs.

Network analyzer pulse response measurement types

Another consideration when making network analyzer measurements is to determine the type of measurement to be made. Detail about how each type is implemented can be found in *Agilent Application Note 1408-12, Pulsed-RF S-Parameter Measurements Using Wideband and Narrowband Detection,* literature number 5989-4839EN. [8]

Average-pulse

Average-pulse measurements are performed by not applying any receiver triggering or gating. This means that the receiver measures and integrates all the energy from the device-under-test (DUT) during the pulse duration. In effect, when using narrowband mode, the gate widths are set equal to or greater than the pulse width, as shown in Figure 37. Using this method the average insertion loss or group delay result over the duration of the RF pulse can be plotted against frequency.

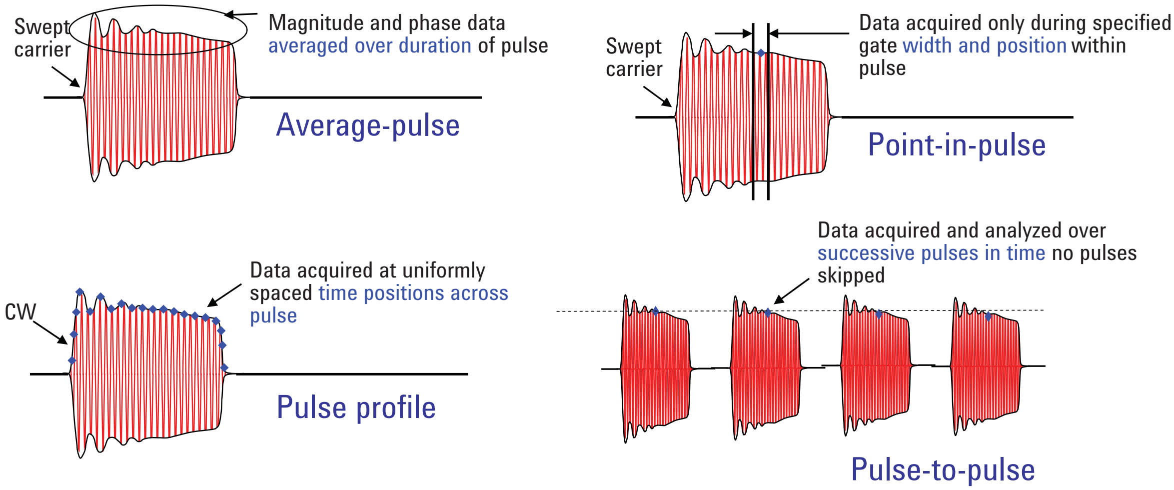


Figure 37. A network analyzer like the Agilent PNA has four different pulse sampling options.

Point-in-pulse measurements

Point-in-pulse measurements provide the user the ability to measure the output of the DUT at any point in time during the pulse by applying a time delay between when the source/bias is pulsed and when the receivers start taking data, as shown in the upper right of Figure 37. A time gate width for which the pulsed energy is allowed to pass to the receivers can also be specified providing a variable receiver integration window. Using this method the insertion loss or group delay result for the given period of time within the RF pulse can be plotted against frequency.

Pulse-profile measurements

Pulse-profiling is similar to point-in-pulse except that the measurement information is displayed in the time-domain, at a CW frequency, where the time axis represents a point-in-pulse measurement with variable time delay (i.e. from a starting delay to a stop delay). This can be thought of as walking the point-in-pulse measurement across the width of the pulse. With the microwave PNA the minimum receiver gate width is approximately 20 ns resulting in excellent resolution for pulse-profiling analysis, as shown in the lower left example of Figure 37. The results of a pulse-profile measurement are displayed as function of time rather than frequency. A pulse profile can be used to analyze phenomena such as pulse droop that may be caused by a component.

Pulse-to-pulse measurements

Pulse-to-pulse measurements are used to characterize how a pulse stream changes versus time due to variations in the performance of the DUT. For example, thermal effects in an amplifier can cause gain reduction and phase shifts. The measurement results are displayed as either magnitude or phase versus time with each data point representing a consecutive pulse. The measurement point remains fixed in time with respect to a pulse trigger. The lower right example in Figure 37 shows a representation of a pulse stream decreasing in magnitude due to gain reduction in a high-power amplifier as it heats up. The pulses do not necessarily need to be repetitive as long as an appropriate pulse trigger is available. For example the measurement can be made with a diversified PRI. However, a pulse-to-pulse measurement is only available in wideband mode.

Antenna Measurements

The performance of the antenna is critical for any radar. The antenna gain is a key variable in the radar range equation and therefore directly influences the range.

Antenna gain is defined as the maximum power relative to the power from an isotropic antenna. It is sometimes referred to as directivity. Gain is usually defined in logarithmic terms, dBi (dB relative to an isotropic antenna) and is expressed by:

$$G_{dBi} = 4\pi\eta A/\lambda^2$$

Where

- η = antenna efficiency
- A = antenna area
- λ = carrier wavelength

In addition to gain, the polarization of the antennas is also an important consideration. The polarization of the transmit and receive antennas must match each other in order to efficiently transfer a signal. Types of polarization include elliptical (most common), linear or vertical, and circular.

The radar antenna is designed to form a steerable beam. The beam is not perfect and will have a width that is defined as the angle between the 3 dB points, as shown in Figure 38. The width is not necessarily the same in both the horizontal and vertical directions. For example, a tracking antenna may have a pencil beam that is the same both horizontally and vertically. However, a search antenna may have a beam that is narrow in the horizontal axis, but more fan-like in the vertical direction. The beam width is important to consider along with the gain of the antenna since the two are linked. As the beam is narrowed the gain increases since power is more focused.

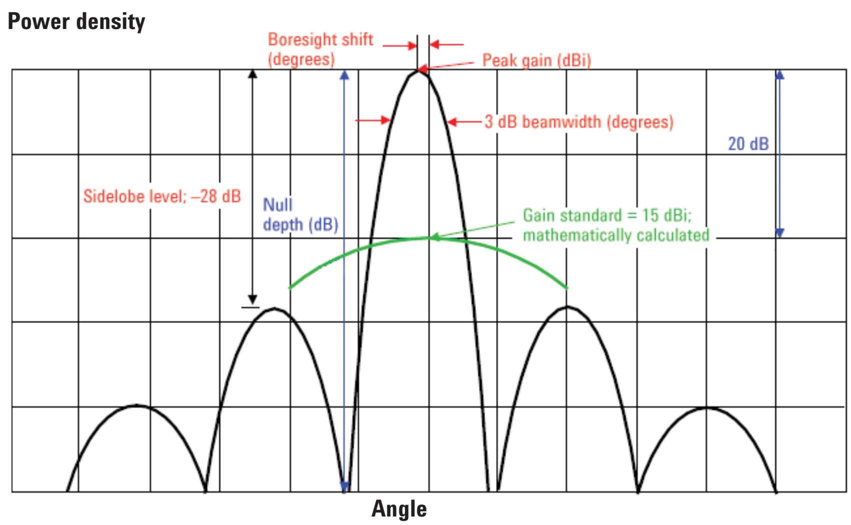


Figure 38. Many importance characteristics of the antenna can be determined from the antenna scan pattern.

Conventional radar relies on mechanical mechanisms to steer the beam. Modern radar may use electronically controlled antenna arrays, which can greatly increase the speed and accuracy of the beam steering. A boresight measurement (measure of the intended angle versus actual measured angle of the peak power) helps calibrate the direction of the beam. The accuracy in steering the antenna will determine the accuracy with which the direction of the target can be identified.

Sidelobes are undesirable artifacts of beam forming that transmit energy in unwanted directions, as shown in Figure 38. Sidelobes are generally small but can be measured against their theoretical limits for a given antenna design. It is desirable that the sidelobes are small to avoid false returns due to objects near the antenna.

In addition to gain, polarization, beam width, boresight, and sidelobes other measurements commonly made on an antenna include frequency response and impedance.

Far-field versus near-field antenna test

There are two different types of test configurations that can be used to test antennas, far field and near field, as shown in Figure 39. Each has advantages and disadvantages.

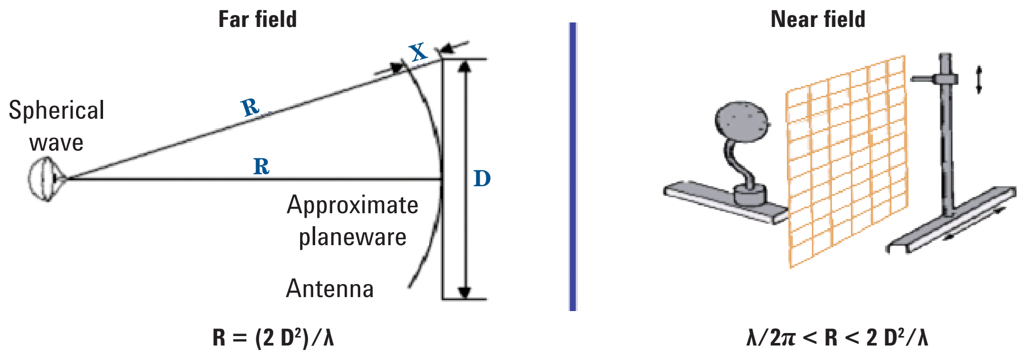


Figure 39. Antenna measurements can be made using either a far-field or near-field approach. Far-field testing is simpler and faster but requires a large area. Near-field testing requires sophisticated calculations, may take longer to perform and requires less space.

Far-field antennas usually operate with a long distance between the source and receive antenna. Antennas radiate a spherical wave-front, but at a great distance away from each other, the spherical wave-front becomes almost planar across the aperture of the receive antenna. Antennas have to be separated to simulate a planar wave-front to reduce receiving errors. The far-field criteria that is generally accepted is $R > 2D^2/\lambda$, which allows 22.5 degrees of phase variation across the aperture of the antenna-under-test (AUT).

Near-field antennas usually operate at much shorter distances between the source and receiving antenna. Very near the antenna plane the field is reactive in nature and falls off more rapidly than the radiating near-field region. Near-field measurements are made in the radiating near-field region of $\lambda/2\pi < R < 2\ D^2/\lambda$. Near-field measurements involve large amounts of data collection and transformation analysis to derive the far-field result.

Far-field test configuration

Far-field test configuration footprints generally run from 10-1000 meters and are the primary drawback to far-field test setups. The advantage is that the testing requires less calculation and can be faster.

With a far-field antenna measurement, the radiated energy is measured real-time as the AUT is rotated in azimuth and elevation coordinates. The resulting data is a measure of amplitude and/or phase as a function of angular position. The rotation of the antenna is usually accomplished by a mechanical antenna positioner, which determines the exact position in the coordinate system and typically restricts movement to a single axis at a time.

An example far-field test configuration using the Agilent PNA is shown in Figure 40. The configuration utilizes an 85320A/B external mixers and 85309A LO/IF distribution unit to provide the first down conversion so that the mixers can be located near the antennas. The first down conversion is to an IF frequency of 8.333 MHz, which is the second IF frequency of the PNA.

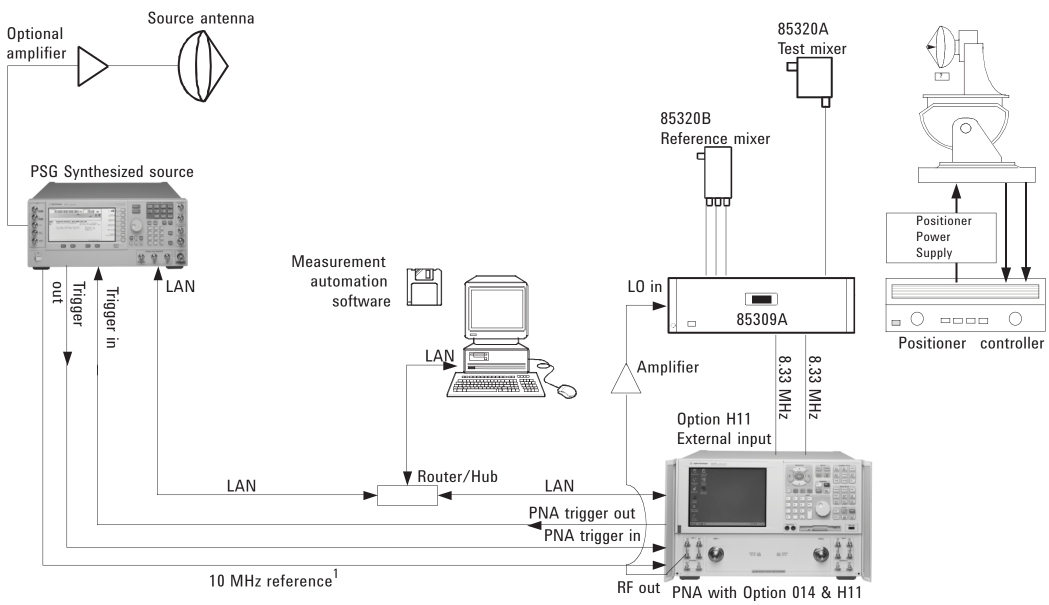


Figure 40. In this example of a far-field antenna test configuration, external mixers are used so that they can be placed close to the reference and test antenna. Directly inputting the IF signal into the PNA second mixing stage also improves measurement sensitivity.

An important aspect of maximizing the receiver sensitivity is to minimize losses to the signal under test. To help with this the PNA provides a way to bypass the PNA coupler and first IF converter stage (Option H11). Doing this improves sensitivity by as much as 20 dB.

For a larger far-field antenna range, controlling a remote microwave source across a significant distance is always a concern. This configuration utilizes a PSG microwave signal generator (source), utilizing TTL handshake triggers between the PNA/PNA-X and the PSG source. Low-cost fiber optic transducers are one method that could be used to provide long-distance TTL transmission signals across a far-field antenna range.

Another example test configuration that can be used for smaller far-field ranges is shown in Figure 41. This simpler configuration can be used when the range is small enough so that cable losses do not impact the measurement.

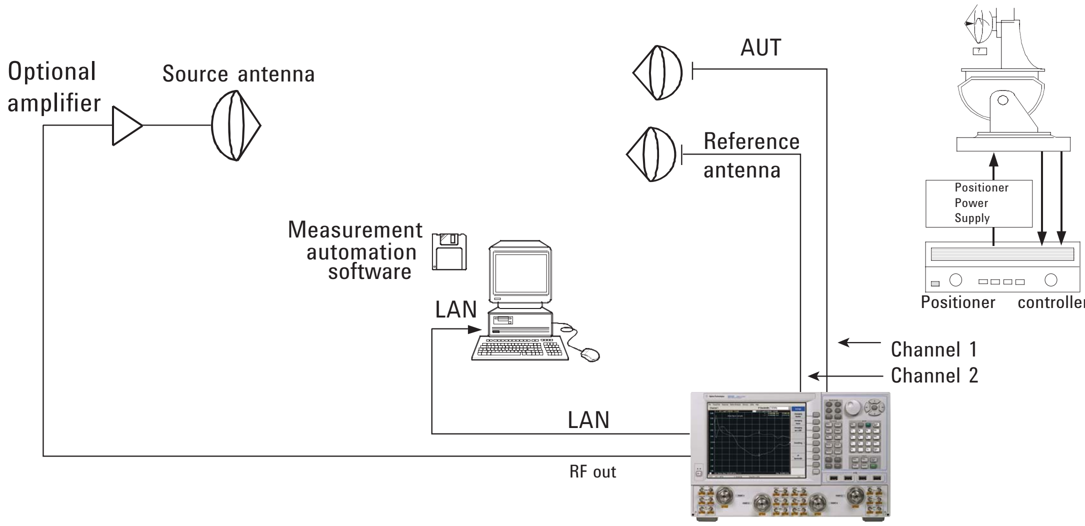


Figure 41. This example of a simple test configuration with a network analyzer is used for smaller far-field test setups, but may not be possible for large ranges due to cable losses.

There are several advantages to using a small-range configuration versus a large-range configuration if possible. With a small range configuration the network analyzer is able to minimize cost, space and complexity by providing both the source and all the required receiving channels. Significant speed and cost improvements can generally be accomplished with this configuration. Additionally, the PNA/PNA-X network analyzer can measure up to four independent inputs simultaneously, thus providing a highly integrated, cost effective solution.

The process for analyzing the antenna pattern once the test process is setup involves making many gain and phase measurements relative to the known reference antenna while varying the angular position of the antenna. This can be a laborious process and therefore is typically done with antenna measurement software that is able to control the position of the antenna and synchronize the measurements with the network analyzer and reference source.

Near-field test configuration

Near-field techniques measure the far-field patterns of an antenna, by means of a different technique. A near-field technique moves a probe across the aperture of the antenna, and measures amplitude and phase points on a sample grid spaced every half wavelength, as shown in Figure 42. Different scan patterns can be used depending on the nature of the antenna. The energy radiated in the near-field region is analytically converted to the far-field result using a Fourier transform computational technique that produces standard far-field antenna patterns.

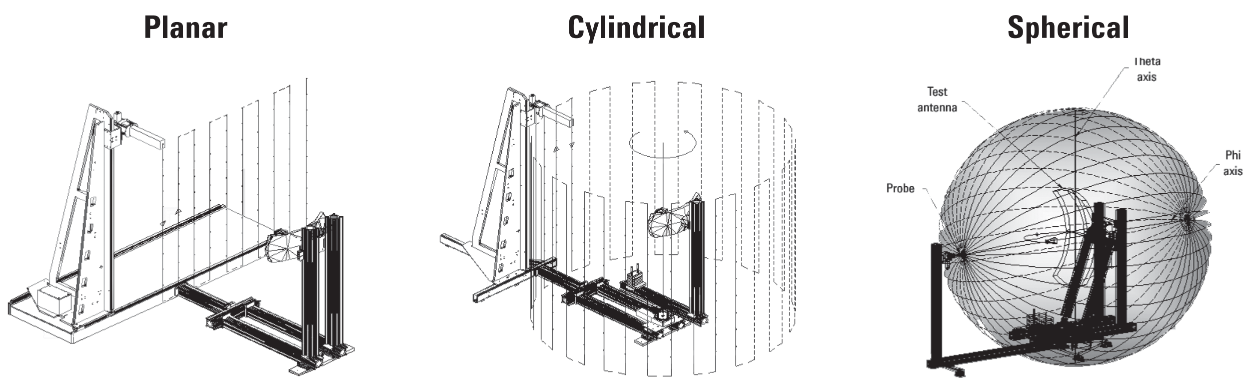


Figure 42. A near-field test uses different scan patterns depending on the type and purpose of the antenna.

Near-field antenna ranges have many advantages over far-field ranges; less space is required, the antenna is protected from weather, there is better security for the antenna and test frequencies, and for very large antennas a near-field system is usually significantly lower in cost. The main disadvantage is the complexity and time required to measure and process large amounts of data.

Figure 43 illustrates a basic near-field antenna measurement configuration utilizing a PNA family network analyzer. It is similar to the small range configuration. The network analyzer operates both as the source and the receiver, while an external software application controls the data collection timing of the network analyzer and the movement of the positioner controller. Additionally, the external application controls the switching of the polarization of the AUT.

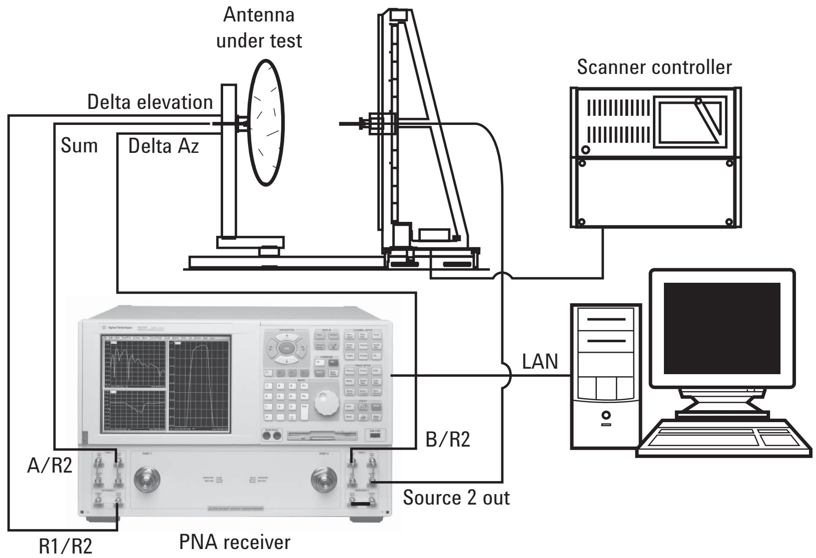


Figure 43. This is an example of a near-field antenna test configuration. Far-field antenna pattern results are extrapolated from the near-field measurements using transformational analysis.

To minimize the test time the frequency can be multiplexed during each data scan. However, this can result in a misalignment of the rectangular near-field grid between forward and reverse data scan directions resulting in errors in the far-field pattern result. This error can be eliminated by always collecting data measurements in the same scan direction; however, this doubles the test time. Alternatively, the frequencies can be scanned in reverse order on the reverse scan. Using this reverse sweep in conjunction with correct triggering between the forward and reverse passes ensures that each frequency set is spatially aligned on the rectangular near-field grid. This technique requires an RF source that supports reverse frequency list mode of operation. The PNA/PNA-X network analyzer includes reverse sweep and edge triggering capability specifically designed for antenna measurements.

Example antenna measurement results

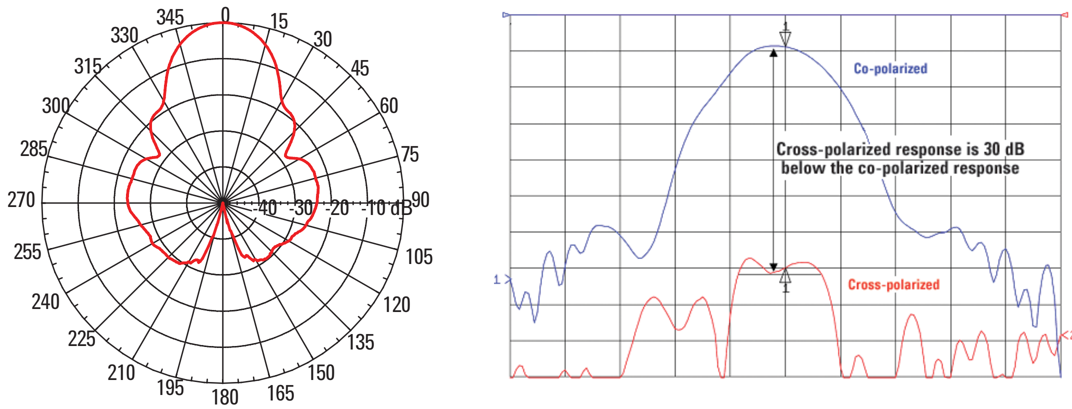


Figure 44. Example a far field antenna scan pattern and and of a cross polarization measurement. The gain can be measured and compared using different polarization orientations of the antenna.

The left image in Figure 44 is an example of a far field antenna scan pattern for an X band radar for the horizontal plane. Measurement results of the principle planes are often used to characterize the antennas performance. The results could be obtained using either near- or far-field techniques. A number of antenna characteristics can be read from the results. These include the gain, beam width, and sidelobe level.

Another example of an antenna scan measuring the cross polarization is shown in the right image of Figure 44. The cross polarization is essentially the difference in the result with opposite polarization. In this case a difference of 30 dB is observed which is a good level of polarization purity. A well designed linear antenna should only respond to the polarization for which it was designed.

Radar Cross Section

The radar cross section (RCS) of a target is a measure of a target's reflectivity in a given direction. The main contributors to RCS are:

- Specular scattering – localized scattering dependent on the surface material/texture
- Diffraction scattering – incident signal scattering at target edges and discontinuities
- Multiple bounce – reflections among target elements at offset angles

As shown by the radar range equation on page 6 the radar cross section (σ) has a direct affect on the range of the radar. Although the radar designer cannot control the cross section of the target, the objective in modeling RCS is to develop simulation tools capable of predicting the behavior of radar receivers in a realistic environment.

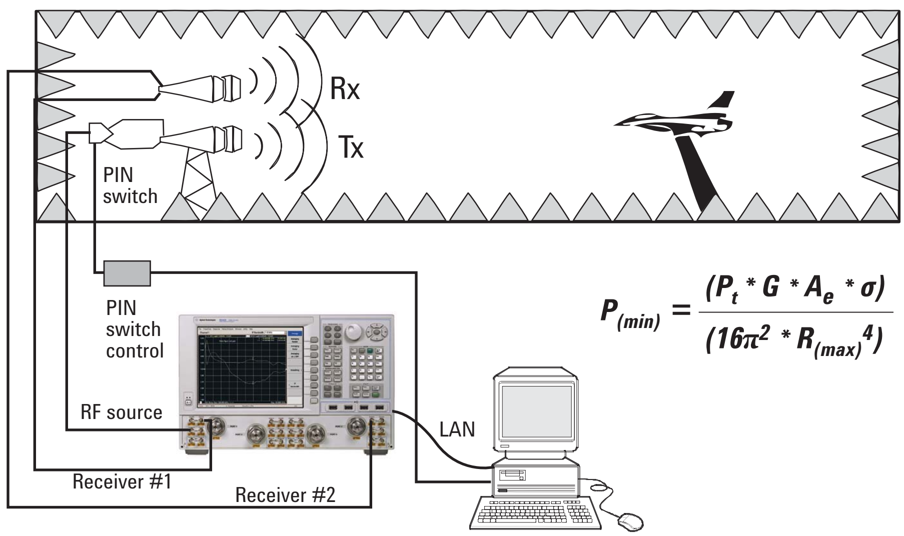


Figure 45. This example radar cross section measurement setup uses a network analyzer and an anechoic chamber.

Figure 45 show a basic range set-up for transmission and reception of co-polarized and cross-polarized signals. In this configuration, the network analyzer measures both polarizations simultaneously through independent measurement channels as well as providing the source signal to the transmitting antenna.

As technology has improved, there have been large innovative improvements in the understanding of how to minimize the reflected energy of an object. This has led to a point where the actual returned signal levels are extremely small and require extremely sensitive measurement instrumentation to acquire these signals. The received signal is going to be very small due to the energy being transmitted and reflected (1/R4 term) and the object reflection term σ which is optimized for the smallest return possible. Additionally, large distances due to the object size and desire to have the impinging waves planner add to reducing the returned signal level.

As a result, it is critical to have instrumentation with very good sensitivity. To achieve the best sensitivity instruments like the Agilent PNA network analyzer are mixer based receivers, as opposed to using sampler based converters, since mixers provide the best sensitivity.

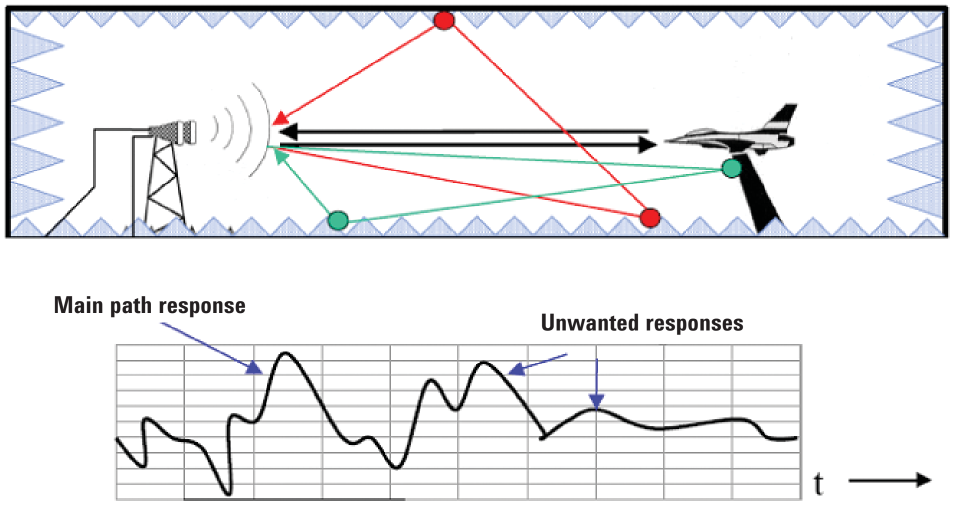


Figure 46. RCS signals are small and returns caused by the measurement chamber can interfere with the measurement results.

Because the signals are small, small reflections caused by elements in the range itself can contribute significant reflected energy. To solve this problem advanced network analyzers like the PNA/PNA-X provide a time gating feature that can remove the unwanted signals. This is achieved by computing an inverse fast Fourier transform (IFFT) on the measured frequency data, mathematically removing the unwanted signals, and then computing an FFT to restore the frequency result. Figure 46 illustrates this concept. An artifact of computing the IFFT on a finite sampling is that it will create repetitions of the fundamental signal in time called aliases. These artifacts can be worked around through a process of testing to create an alias free measurement span. The width of this span will depend partly on the number of data points the analyzer is able to measure and process. A typical number of data points for a network analyzer may be 1601, which is suitable for an alias free range required for many measurements. However, more may be needed. The PNA/PNA-X network analyzer provides 20001 points to ensure wide alias free spans for measurements for test setups that require them. More detail on the time gating process can be found in *Agilent Application Note 1287-12, Time-domain Analysis with a Network Analyzer*, literature number 5989-5723EN. [9]

Noise Figure

As accounted for in the derivation of the radar range equation, the threshold of the radar receiver is based on following four factors: noise figure (NF), KT (Boltzmann's constant * temperature), the noise bandwidth of the system, and the S/N ratio. KT is the familiar –204 dBW/Hz, effectively a constant with little opportunity for improvement. The system bandwidth is dictated by the radar design, and the S/N cannot be improved once the signal gets to the receiver. Thus, receiver noise figure becomes the term of great interest for receiver optimization.

When considering NF it is helpful to review Friss's formula. First, to avoid common confusion, noise factor (F) is a ratio of the S/N at the input versus S/N at the output of a device. Noise factor is therefore a unit-less ratio, whereas noise figure (NF) is 10 log of the noise factor and is expressed in decibels.

Friss's formula describes how the noise factor of successive components in a system adds up to determine the total noise factor of the system.

$$F_{Total} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots$$

Where F_x is the noise factor for each successive element.

What is telling about the formula is that the first component, usually an amplifier, in the chain typically has the greatest impact on the overall NF since the effects of the other elements are reduced by the gain of the previous element.

At first glance one might be inclined to believe that noise figure is an area where great improvement in system performance could be achieved at little cost. Modern, low-noise amplifiers can deliver very low-noise figures. Properly engineered into the receiver architecture, the system noise figure penalty can be minimal. Therefore, it would seem that it might be more economical to reduce receiver noise figure by 3 dB rather than to increase transmitter power by the same amount. However, the reality is not quite that simple. The receiver designer must also be concerned with providing adequate gain, phase stability, amplitude stability, dynamic range, fast recovery from overload and jamming, and reliability. In addition, protection must be provided against overload or saturation and burnout from nearby transmitters.

Because of these considerations, many radar receivers do not employ low-noise RF amplifiers in the receiver front-end, but simply use a mixer as the first receiver stage. Regardless NF is a critical metric that must be optimized within other given restraints.

There are two main methods used today for measuring noise figure. The Y-factor or hot/cold source method and the direct noise or cold-source method. This section will discuss these two methods and the instruments used to make them.

Y-factor measurement technique

The most prevalent method is the Y-factor or hot/cold source technique. It relies on a noise source that is placed at the input of the DUT. The noise source generates excess noise compared to a room temperature termination. When the noise source is turned off, it represents a cold source termination (or the same termination as would be present with a passive termination at room temperature). When the noise source is turned on, it presents a hot termination. While not physically hot, the noise source's excess noise can be described by an equivalent temperature that would produce the same amount of noise from a hypothetical termination that was actually at the temperature. With these two termination states (hot and cold), two measurements of noise power are made at the output of the DUT. The ratio of these two powers is called the Y factor. Using the Y factor and the ENR (excess noise ratio) calibration data from the noise source the overall noise factor of the system can be determined by the following equation:

$$F_{sys} = \frac{ENR}{Y - 1} \quad \text{Where } ENR = \frac{T_h - T_c}{T_o}$$

Note that this equation results in the noise factor of the system, which includes the measurement instruments. In cases where the gain of the DUT is high, the noise factor of the system may be used to approximate the noise factor of the DUT. However, this may not be the case. To remove the noise factor effects of the measurement instrumentation requires calibration. This is basically done by having the measurement system measure its own noise factor and gain and then calculating out its own contribution using Friss's equation. (More detail on the Y-factor method can be found in *Agilent Application Note 57-1, Fundamentals of RF and Microwave Noise Figure Measurement,* literature number 5952-8255E.) [15]

The Y-factor method is the method most likely used by dedicated noise figure analyzers and by spectrum/signal analyzers with built in noise figure functions. Examples of instruments that use the Y-factor technique include the Agilent NFA Series noise figure analyzers and the Agilent PSA Series spectrum analyzer. These instruments simplify the measurement process by automating the process and measurement calculations. In the case of the Agilent SNS noise sources and the NFA Series noise figure analyzers the process of entering ENR calibration data done automatically by loading the values into the analyzer automatically from an electrically erasable programmable read-only memory (EEPROM) on the noise source.

Direct-noise or cold-source measurement technique

The second method is called the cold source or direct noise technique. With this method, only one noise power measurement is made at the output of the DUT, with the input of the amplifier terminated with a source impedance that is at room temperature. The cold source technique requires an independent measurement of the DUT's gain. Consequently this technique is well suited for use with vector network analyzers (VNA) since a VNA can make very accurate error-corrected measurements of gain. It does, however, also require a network analyzer that can make noise as well as CW measurements. An example of a VNA capable of making noise figure measurements is the Agilent PNA-X network analyzer.

Using a vector network analyzer to make noise figure measurements has several advantages. One example is that a network analyzer is able to make other important measurement during the same connection to the DUT. For example the PNA-X is able to measure NF, gain, IMD, and S-parameters. In addition, analyzers like the PNA-X are able to take advantage of advanced error correction techniques including the use of an ECal electronic calibration module as an impedance tuner to correct for imperfect system source match. This is especially useful in an automated test environment where switch matrices are typically used. These advanced features, combined with the very good sensitivity of the PNA-X provide the highest performing noise figure solution from Agilent.

Selecting the best noise figure measurement solution

The best noise figure solution will depend on the measurement objectives, relative NF and gain of the DUT being measured, and the cost restraints.

A spectrum analyzer based noise figure solution generally has the lowest incremental cost and provides the versatility to perform spectrum measurements such as transmit spectrum, IMD, spurious, etc. However, a spectrum analyzer will generally have greater measurement uncertainty than a dedicated NF analyzer. This is partly due to a higher noise figure of the analyzer. For high gain devices however, the noise figure of the measuring analyzer will have minimal effect on measurement uncertainty. As a result, spectrum analyzers based solutions may be the best solution for measuring high gain devices.

A dedicated noise figure analyzer such as the Agilent NFA is designed to have low noise figure and low measurement uncertainty. The NFA analyzer may provide the best solution when a high level of measurement accuracy is required without the extra performance and cost of a noise figure capable network analyzer. Ultimately the right solution may depend on which solution meets your specific performance requirements. More information on optimizing and determining the accuracy of Y-factor based solutions can be found in *Agilent Application Note 57-2, Noise Figure Measurement Accuracy: The Y-Factor Method,* literature number 5952-3706E. [16]

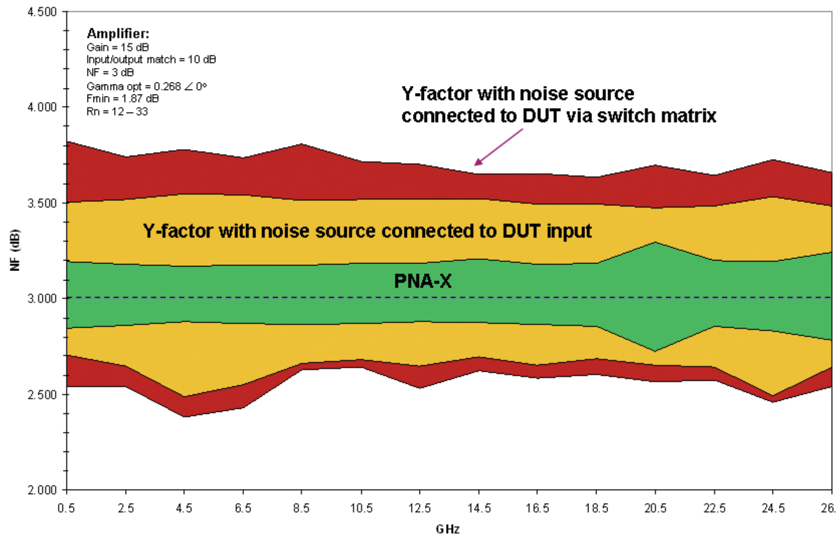


Figure 47. The PNA-X based noise figure solution provides the best measurement uncertainty due to its high sensitivity and ability to use vector error corrections to subtract out errors due to effects such as impedance mismatch.

The cold source based PNA-X network analyzer noise figure solution offers the highest level of performance for noise figure from Agilent for reasons previously stated. Figure 47 shows a typical comparison in accuracy using the traditional Y-factor method with a traditional noise figure solution compared to cold-source method using a PNA-X with source correction and vector calibrations. Results are show both with and without a switch included in the test system. An additional benefit of the PNA-X solution is that it also offers measurement capability such as S-parameters and IMD with a single connection to the DUT.

Phase Noise, AM Noise, and Spurs

Phase noise, AM noise, and spurs can have significant performance implications for radar. Phase and AM noise in the receiver decrease the S/N ratio. Noise on the transmit signal results in noise on the radar returns that can hide low level Doppler target signals in the presence of clutter. Spurs created by unwanted discrete PM or AM oscillations can result in false targets.

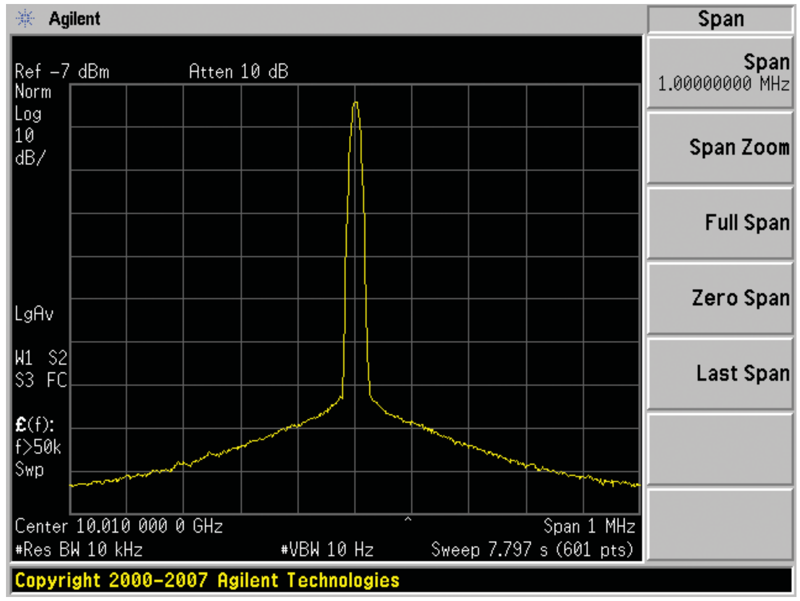


Figure 48. Phase noise results in sidebands that spread power into adjacent frequencies reducing the signal-to-noise ratio.

Phase noise is the results of random fluctuations in phase on a signal caused by time-domain instabilities. In the time-domain this effect is called jitter. In the frequency-domain this instability manifests itself as sidebands that spread power to adjacent frequencies, as shown in Figure 48. If the phase variations are random or random-like, the sidebands will slope downward from the signal. If, however, the phase variations of the signal is due to specific non random oscillations the result will be discrete spectral components or spurs. AM noise also spreads power into adjacent frequencies. Phase noise is more dominate at closer frequency offsets, AM noise may become more dominant at wide offsets.

In radar, good phase noise is critical for oscillators such as stable local oscillators (STALOs) and coherent oscillators (COHOs) as these signals are at the heart of the radar. Any phase impairments on these signals will be multiplied as they are transferred up to the higher frequency transmit and receive signals effectively reducing the S/N ratio.

Phase and AM noise effects may be particularly harmful for MTI (moving target indicator) radars. These radars work on the principle that a moving target return will be shifted in frequency as a result of Doppler effects. Typically the returns of the target are small when compared to the clutter that is returned from stationary objects such as the ground or a mountainside. Since the clutter and the target returns are at different frequencies the clutter is filtered off and the target's returns examined. However, the clutter canceling filters will not be able to filter out the noise on the clutter. Good noise performance on the transmit signal is therefore important for the operation of the radar.

Different measurement solutions are available for measuring phase noise. The appropriate solution will depend on cost and performance restraints. Phase noise measurements on CW signals within the radar including the STALOs and COHOs are critical, but it may also be necessary to measure the phase noise of pulsed signals or understand the phase noise contribution added by system components such as the power amplifier (residual or additive phase noise). This is especially true for Doppler radar where it is critical to understand the phase noise through the transmit path under normal operating conditions.

Instruments that may be used to measure phase noise include spectrum analyzers, signal source analyzers, or a dedicated phase noise test systems.

A spectrum analyzer based phase noise measurements typically offers the lowest cost option. A dedicated signal source analyzer like the Agilent SSA signal source analyzer provides high performance and efficiency for measuring oscillators and phase lock loops (PLLs). A phase noise test system is more complex but offer the greatest flexibility and performance. A phase noise test system may be the only available solution for making pulsed and residual phase noise measurements.

Making a direct spectrum measurement of phase noise with a spectrum analyzer

Phase noise measurements on a spectrum analyzer are made by direct analysis of the spectrum and examining the level of the phase noise sidebands. This process can be done manually with any spectrum analyzer simply by using marker functions and measuring the noise level at the desired offset frequency. Phase noise is typically measured in dBc/1 Hz and therefore should be normalized to 1 Hz based on the RBW setting. In addition, a noise correction term may also be needed depending on the detector and display mode used. (For more information see *Agilent Application Note 1303, Spectrum Analyzer Measurements and Noise,* literature number 5966-4008E.) [5] To simplify, most spectrum analyzers include a noise marker function that will do this automatically. When measuring the noise with a spectrum analyzer both phase and AM noise are actually included in the result, but since phase noise is usually dominant the measurement is frequently just referred to as phase noise.

Many spectrum analyzers also include automated single side band (SSB) phase noise measurement functions. Figure 49 shows an example of a phase noise measurement taken using the Agilent PSA spectrum analyzers with built-in phase noise function. The measurement function works for CW signals. The range of offsets and level of the phase noise that can be measured by a spectrum analyzer will depend on the RBW settings available and on the phase noise of the instrument itself. In the case of the PSA, it is able to measure offsets as close as 100 Hz. Its typical performance at a 10 kHz offset is –118 dBc/Hz.

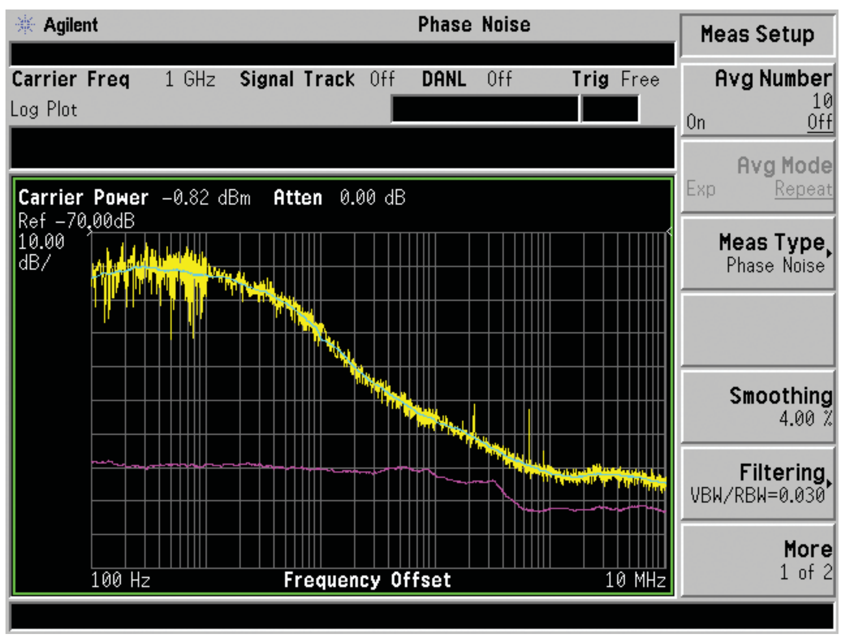
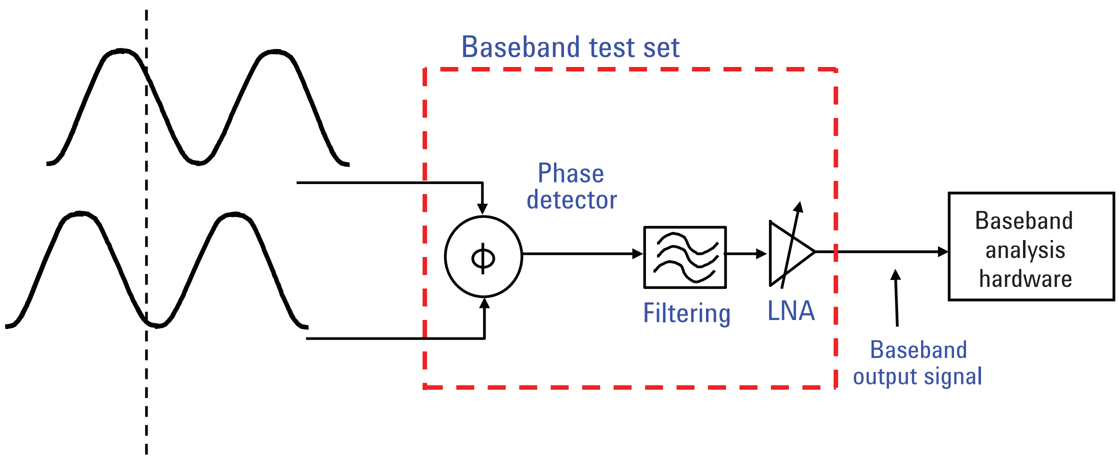


Figure 49. With the PSA spectrum analyzer phase noise measurement personality results are displayed as a function of offset frequency from the carrier.

Measuring phase noise using phase detector technique

To achieve the best sensitivity and accuracy, most dedicated phase noise and signal source analyzers are based on a phase detector, as shown in Figure 50. Most phase detectors are balanced mixers that require both an RF and reference signal. When the RF and reference signals are in quadrature with respect to each other and applied to the balance mixer, the IF output results in a measure of the instantaneous phase difference between the two signals. In most implementations, the quadrature phase relationship is maintained using a narrowband PLL. The instantaneous phase difference is represented by an instantaneous voltage change around 0 volts. Using the double balance mixer, in quadrature, suppresses amplitude noise while phase noise is being measured. The noise-voltage (IF output) is then amplified and spectrally processed to determine the noise and spurious signals as a function of offset frequency. Phase noise measurement systems also typically also include a separate AM detector to measure amplitude noise and amplitude spurious signals as a function of offset frequency. The AM detector may be diode-based or mixer-based.



$$\Delta V_{out} = K_{\phi} \Delta\phi_{in} \quad \text{for small differences in phase}$$

Figure 50. A basic block diagram of phase detector based phase noise measurement. The measurement compares the test signal to a phase locked reference signal with a phase detector. The variations in phase due to phase noise result in voltage output which is then processed to determine the phase noise result.

The measurement sensitivity is limited by the phase noise of the reference oscillator or any microwave down converter used in the measurement. New phase noise analyzers use advanced cross correlation techniques to improve the measurement performance beyond that of the reference oscillator and downconverter. This is especially useful in measuring voltage controlled oscillators (VCOs), which tend to have very low, far from carrier phase noise characteristics. For measurement systems that do not use cross correlation techniques a mathematical correlation can be done against three sources to extract actual noise performance of each source. (*Agilent Technologies, Users Guide Agilent Technologies E5500A/B Phase Noise Measurement System,* Part number E5500-90004, 2000).

E5052B Signal Source Analyzer

Characterize the performance — phase noise, transients for high frequency signal sources up to 110 GHz

- Worlds fastest measurement throughput
- Advanced cross-correlation technique for maximum performance
- Simultaneous transient measurements for frequency, phase, & power over time
- Real-time spectrum monitoring

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Measuring phase noise using a signal source analyzer

The Agilent SSA signal source analyzer is one such instrument that employs cross correlation techniques to dramatically improve the measurement performance. For convenience the analyzer includes its own reference oscillators. The SSA is designed to efficiently process cross correlations rapidly to obtain the maximum performance and efficiency. As a result the SSA offers a very high level of performance while keeping costs low.

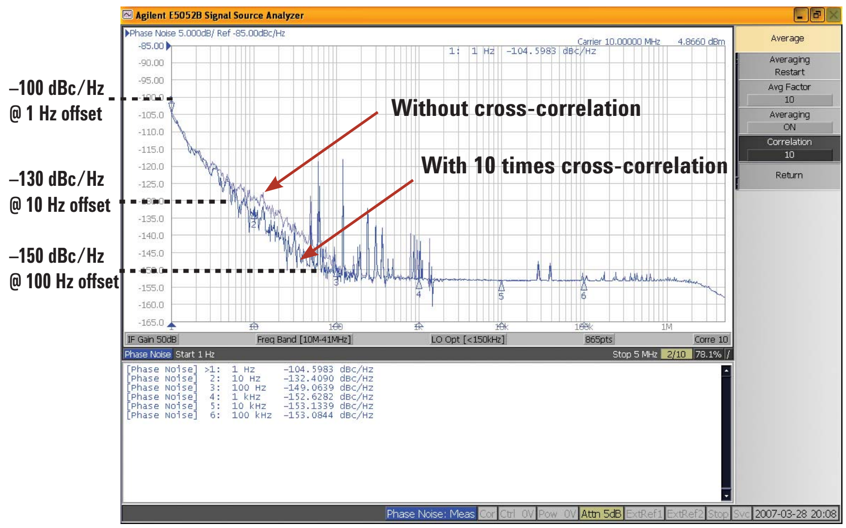

Figure 51. This phase noise measurement uses Agilent's signal source analyzer (SSA). The phase noise measurement is improved using cross correlation technique.

Figure 51 shows an example of a phase noise measurement using the SSA with different levels of cross correlation. The time tradeoff on the SSA analyzer for an improvement in 5 dB computing 10 correlations is approximately a 10 times longer measurement time.

In addition to phase noise, the DSP-based SSA also provides many other functions that can be useful for testing radar oscillators. Since it samples the signal and is DSP based, the analyzer includes the ability to analyze amplitude, frequency, and phase as a function of time. It also has the ability to perform transient analysis by triggering on frequency anomalies. The SSA provides two receiver channels: a wideband channel to monitor frequency change and a narrowband channel to capture the frequency versus time profile very precisely. The signal can be measured simultaneously in both channels while either the wideband channel or the narrowband channel monitors the signal for frequency changes. These frequency changes can then trigger and capture the transient event. An example is found in Figure 52, which shows the capture and analysis of a phase hit.

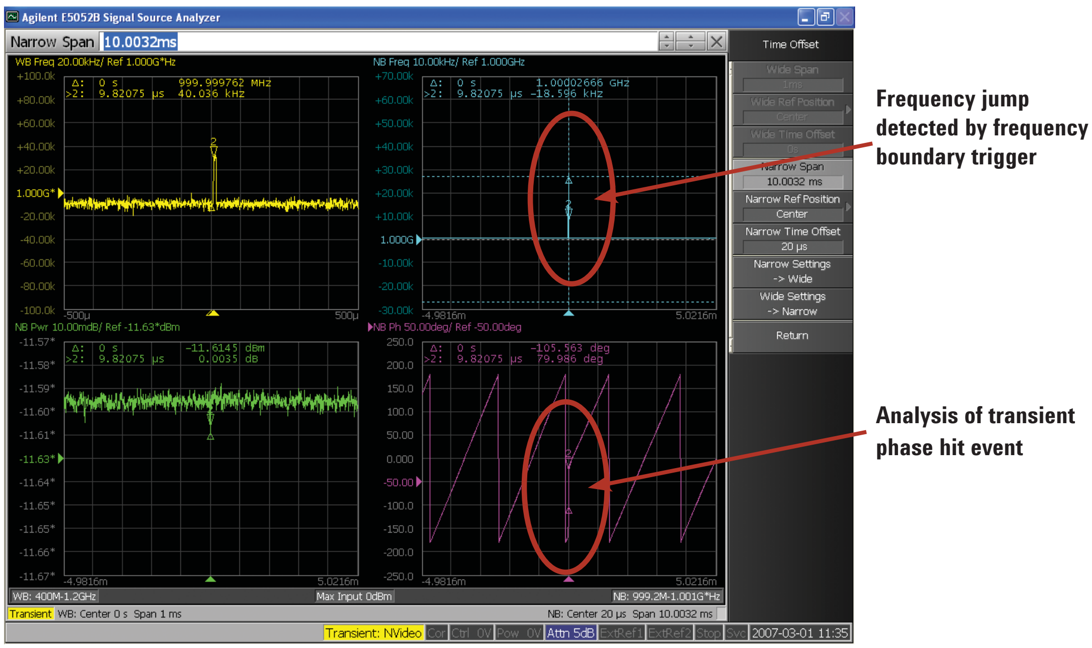

Figure 52. The SSA includes time domain analysis and advanced trigger functions useful for examining transient events. In this screen, the analyzer is able to use its frequency boundary trigger to capture and analyzer a phase hit.

Measuring phase noise with the E5500 phase noise test system

A phase noise test system like the Agilent E5500 phase noise test system offers the most flexibility and performance. Its capabilities are especially useful for radar because of its ability to perform pulsed absolute and residual (or additive) phase noise measurement. It is also able to measure over a wide range of frequency offsets from 0.01 Hz to 100 MHz. The E5500 is phase detector based like the SSA but requires a separate reference oscillator.

The main components of an E5500 system include a low noise down conversion module, external reference source, phase noise test set (detector, PLL), a digitizer based FFT analyzer and/or a swept spectrum analyzer, and PC software. Based on its modular system design different hardware components can be used depending on measurement requirements. The E5500 works well in ATE applications as it is fully programmable, and it can share common components such as the reference source and spectrum analyzer for other measurement uses. E5500 options include: absolute or residual measurements, CW or pulsed phase noise and spurious measurements, and AM noise measurements.

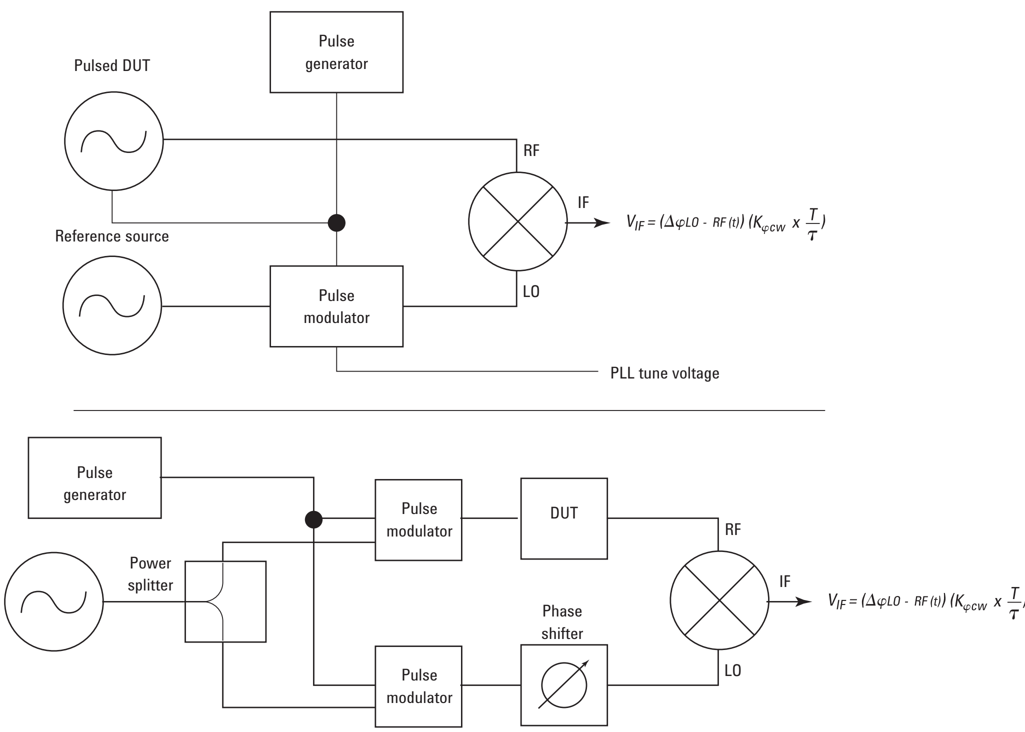

Figure 53. Test configuration for making absolute and residual pulsed phase noise measurements using the Agilent E5500 phase noise test system.

Pulsed and residual phase noise measurements are especially helpful for pulsed radar systems. These measurements are available with the E5500 system but not with a spectrum analyzer based system or SSA. Figure 53 shows a basic block diagram of the configuration used for pulsed absolute and residual phase noise measurements. Detail on how phase noise measurements are performed using the E5500 system can be found in *Agilent Application Note 1309, Pulsed Carrier Phase Noise Measurements*, literature number 5968-2081E. [18]

4.0 Summary

Since its inception over 50 years ago, radar has become ubiquitous and the breadth of application is still growing. At the same time, radar applications have grown more sophisticated as signal processing is used to enhance the returned signal and to extract information, such as target images, using post-processing. However, no matter how sophisticated the signal processing becomes, the performance of the radar is directly determined by the quality of the underlying radar transmitter and receiver.

Understanding radar measurements and how instrumentation responds to radar signals is crucial to designing high-performance and cost-effective radar solutions. This application note has sought to give an explanation of common radar measurements and the measurement solutions available today. Critical radar measurements include power, spectrum, pulse characteristics, antenna gain, target cross section, component gains and losses, noise figure, and phase noise—all of which can be shown, through the radar range equation, to directly influence the performance of the radar. These measurements should virtually always be made using quality test equipment designed to deal with the unique and demanding signal characteristics peculiar to radar.

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