WHITE PAPER

Basics of Power Amplifier and Front End Module Measurements

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The power amplifier (PA) is one of the most integral RF integrated circuits (RFICs) in the modern radio. Whether it is a discrete component or part of an integrated front end module (FEM), the PA can dramatically affect the performance of a wireless transmitter. For example, a cellular PA's power added efficiency (PAE) can significantly affect the battery life of a mobile device, and its linearity can affect the ability of a receiver to demodulate a transmitted signal.

Discrete Components versus Front End Modules

In the early days of technologies such as GSM and UMTS, it was common for mobile devices to use discrete amplifiers for each GSM and UMTS radio. However, the introduction of LTE and WLAN technology—along with the use of more radio spectrum bands—has driven a need for more integrated RF front end technology. Today, vendors are increasingly packaging more devices, including PAs, low noise amplifiers (LNAs), diplexers, and switches, into a single component. As a result, today's RF test engineers are commonly tasked with testing highly integrated FEMs, shown in Figure 1, rather than a discrete PA. Although the required measurements for a FEM are generally the same as the measurements for discrete components, testing integrated FEMs typically requires additional steps to configure the device under test (DUT).

WLAN FRONT END MODULE

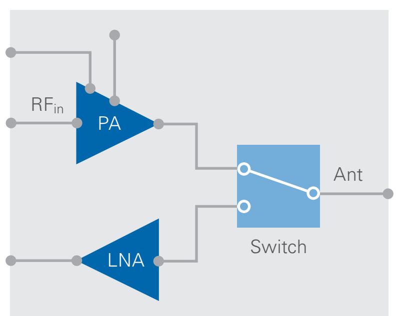


Figure 1. FEMs often integrate PAs and LANs into the same component.

When characterizing the performance of an RF PA, engineers can use a wide range of measurements and test techniques to understand the gain, linearity, and efficiency of the device. In practice, the specific measurements required to characterize a device depends on the intended application for the amplifier. For example, although metrics such as gain and efficiency are important for all PAs, devices intended for the transmission of wireless communications require standard-specific measurements. The error vector magnitude (EVM), one of the most important metrics of a PA, is designed to transmit modulated signals, and the adjacent channel leakage ratio (ACLR) is one of the most important measurements for UMTS or LTE PAs.

Gain and Output Power

Two important characteristics of an RF PA are gain and output power. Gain describes the relationship between the input and output power of the device. In general, a PA exhibits relatively constant gain across a wide range of input power levels with the gain dropping as the output power approaches the device's saturation region. This effect is known as gain compression.

INPUT VERSUS OUTPUT POWER

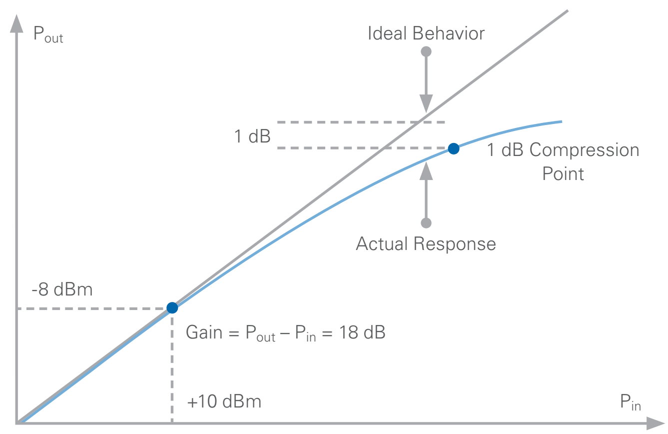


Figure 2. Input versus Output Power in a Typical PA

One of the most common methods to characterize a PA's usable maximum output power is with the 1 dB compression point metric. The 1 dB compression point, shown in Figure 2, describes the operating point at which a PA delivers gain that is exactly 1 dB less than the gain it would otherwise deliver in its linear operating region. For example, if a PA delivers 18 dB of gain in its linear region of operation, the 1 dB compression point is defined at the output power at which the PA delivers exactly 17 dB of gain.

When testing the 1 dB compression point, you can use either a power-calibrated vector network analyzer (VNA) or a combination of an RF signal generator and an RF signal analyzer. The RF signal generator and signal analyzer combination provides the fastest method to measure the 1 dB compression point, and you can use either a continuous wave (CW) signal generator or a vector signal generator (VSG) to perform this measurement.

You can measure gain as a function of input power by sweeping the power level of the signal generator and measuring the output power of the PA with the RF signal analyzer. One optimization technique to consider for production test is configuring the VSG to generate a ramp waveform instead of a series of CW tones at different power levels, as shown in Figure 3. By acquiring the ramp signal with the vector signal analyzer (VSA), you can easily correlate the input power to output power to determine gain versus input power. This ramp signal method is substantially faster than configuring the signal generator for multiple steps, and can save valuable test time.

MEASURING THE COMPRESSION POINT WITH A RAMP SIGNAL

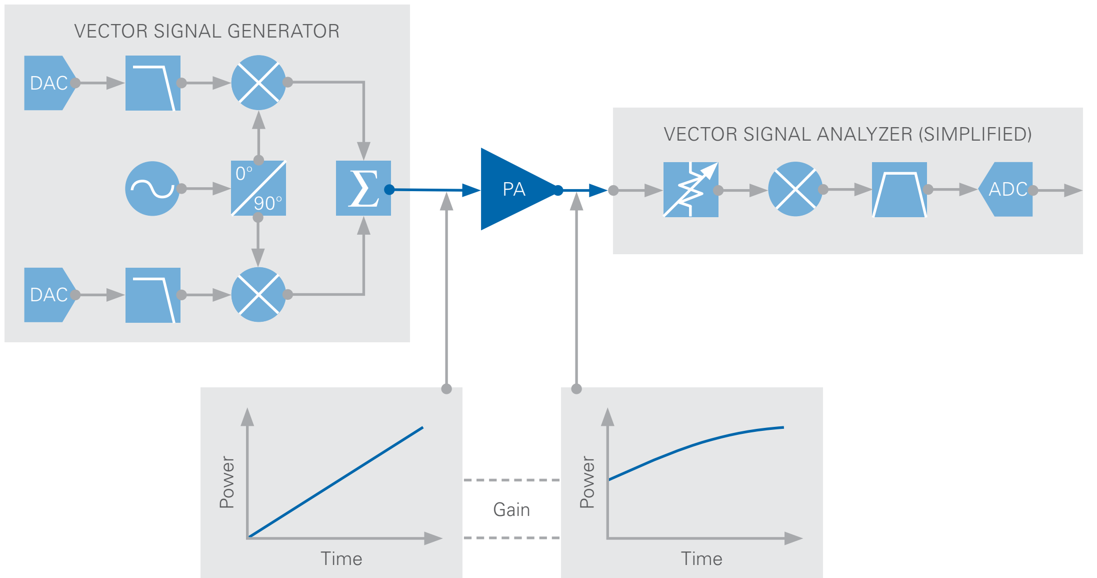
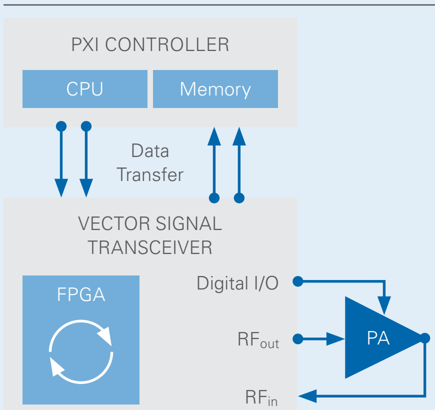


Figure 3. You can measure the 1 dB compression point faster by stimulating a PA with a ramp signal.

Fast Power Level Servoing Using the NI Vector Signal Transceiver

A unique technology of the NI PA test solution is FPGA-based power level servoing using the NI Vector Signal Transceiver (VST). Power level servoing is traditionally a time-consuming process. However, you can achieve the fastest possible power level convergence by performing the control loop entirely on the instrument FPGA. If you decouple the power level servoing algorithm from the embedded controller and perform it on an FPGA, the test software can exploit dramatic measurement parallelism. This results in significant reductions in test time and test cost. For more information about fast power measurements using the NI VST, visit FPGA Servoing for Power Amplifier Test.

PXI SYSTEM



One important technique to improve the power accuracy of gain and power measurements is to use a small attenuator, or pad, between the instrumentation and the PA under test. The power uncertainty due to mismatch error is significantly reduced by using an in-line fixed attenuator on both the PA input and output, as shown in Figure 4.

IMPROVING MISMATCH UNCERTAINTY

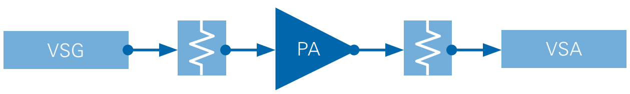
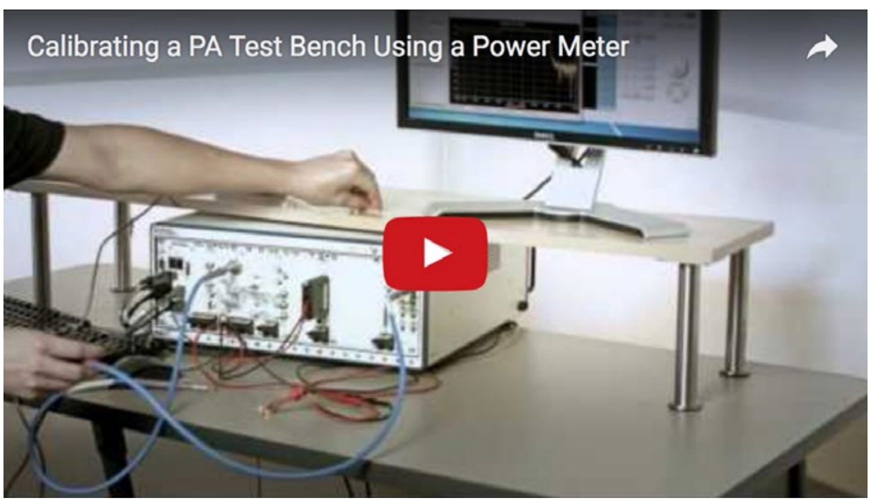


Figure 4. Attenuators between the instrument and the PA improve mismatch uncertainty.

Calibrating Power Measurements with a Power Meter

You can measure the output power of a PA using either a power meter or a VSA. Historically, the power meter was the most accurate method of measuring power with absolute power measurements , providing accuracies to within ±0.2 dB. However, today's modern VSAs are equipped with tools such as an onboard calibration standard that greatly improve their accuracy when measuring absolute power. Using only onboard calibration, a VSA, such as the NI PXIe-5668R, can measure power to within ±0.4 dB, and can achieve even better power accuracy when referenced to a calibration standard, such as a power meter.



Although power meters can generally measure RF power more accurately than a VSA, there are several advantages to measuring the DUT's output power and gain with a VSA. In addition to the simplicity of doing many measurements with a single instrument, VSAs can generally measure power faster than power meters, and, because of this, many engineers rely on the VSA to measure power along with the 1 dB compression point in automated RF test applications.

An important step when measuring power and gain is to calibrate the system setup using a power meter. To complete this calibration step, you first connect a power meter to the reference plane at the input of the DUT, as shown in Figure 5. Using the power meter, you measure the output power of the signal generator plus any attenuators and cabling over a range of frequencies. Once you complete this step, you have characterized the signal generator to within the power accuracy of the power meter.

SYSTEM CALIBRATION

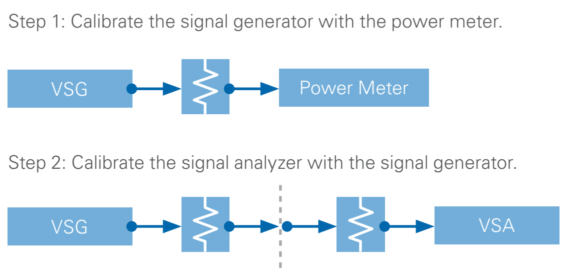


Figure 5. System calibration is a two-step process that uses a power meter to calibrate both the signal generator and analyzer.

After you calibrate the signal generator setup, you then connect the signal analyzer setup, which includes both instruments and any cables, attenuators, and so on, directly to the signal generator setup. Using the calibrated response of the signal generator, and assuming the measurements made with the power meter are correct, you can then determine the measurement offset of the signal analyzer setup. By executing these calibration steps, you can more accurately measure output power and gain by referencing the result to the power meter.

Measuring Gain with a Vector Network Analyzer

Although the most common, and often fastest, way to measure PA gain in automated testing scenarios is with a VSG and VSA, you can measure PA gain with a VNA as well. To measure PA gain using a two-port VNA, connect port 1 of the VNA to the PA input and connect port 2 of the VNA to the PA output. Then measure the S21 response, which equals the PA gain.

One of the key considerations when measuring PA gain with a VNA is to ensure that the output power of the PA does not saturate or damage the VNA receiver. In this scenario, the exact amount of external attenuation can significantly affect the accuracy of the S21 measurement. Although many VNAs have a maximum safe input power level that is typically on the order of 1 W (+30 dBm), measurement accuracy typically degrades when operating the instrument close to the maximum power level, especially because VNAs typically have a much narrower programmable attenuator range than a VSA does.

An accurate PA measurement using a VNA requires careful attention to the power levels present at the input of port 2. As a general rule of thumb, ensure that the source power of the PA and the input power of the VNA port 2 are relatively similar. Thus, if you expect the PA to produce 20 dB of gain, you should connect a 20 dB attenuator between the PA output and the VNA port 2, as shown in Figure 6.

AVOIDING PORT SATURATION

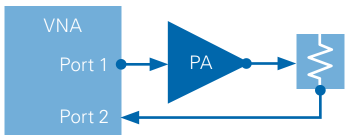


Figure 6. Use an attenuator when measuring PA gain with a VNA to avoid saturating Port 2.

One important nuance of using an attenuator between the output of the PA and the VNA port 2 are the implications on the calibration reference plane. Whether you calibrate the VNA using the short-open-load-thru (SOLT) method or with an automatic calibration kit, you should aim to establish a reference plane as close to the DUT as possible.

In the case of using an external attenuator, you should calibrate the measurement system with the attenuator and any associated cables and fixturing in the path, as shown in Figure 7. As a result of calibrating the measurement system with the attenuator in the signal path, subsequent VNA S21 measurements display the gain directly. For more information about VNA calibration, visit Introduction to Network Analyzer Measurements available on ni.com.

UNDERSTANDING THE REFERENCE PLANE

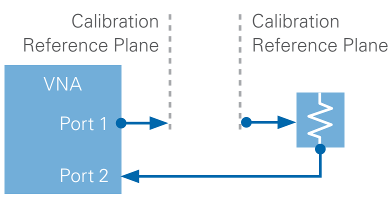


Figure 7. The VNA calibration reference plane must extend beyond the external attenuator.

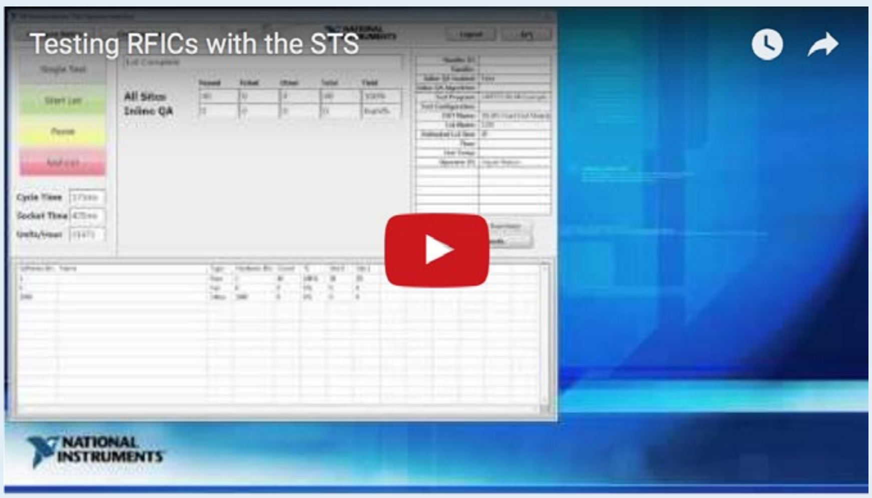
Return Loss and Reverse Isolation

Although measurements such as gain do not technically require a VNA to perform the measurement, return loss and isolation measurements do require full network analysis. The instrumentation setup for return loss and reverse isolation can vary depending on whether you are attempting to characterize the small signal or large signal behavior of the PA. The small signal is the signal within the linear region of operation, and the large signal is the signal in the nonlinear region of operation. When measuring the small signal behavior, you can accurately measure S11 (input return loss) and S22 (output return loss) using a VNA.

In some instances, measuring the output return loss might require slight modifications to the test configuration, shown in Figure 8. The required attenuation used between the PA output and the VNA port 2 can be relatively high, especially for high-gain PAs. In this scenario, the combination of high PA gain and relatively low return loss results in an extremely low-powered reflection signal, measured by port 2 of the VNA. As a result, accurate S22 measurements on a high-gain PA often require the use of an attenuator that produces less loss than the amplifier produces gain. In these instances, it is actually common to use one

Fast S-Parameter Measurements in Production Test with STS

The NI Semiconductor Test System (STS) is a fully automated production test system that applies an innovative approach to S-parameter measurements in production test. This system combines port modules with the NI Vector Signal Transceiver (VST). In addition to switching and pre-selection, the port modules contain directional couplers that effectively turn the VST into a VNA. As a result, you can quickly perform S-parameter measurements in a production test environment without the cost of additional instrumentation. S-parameter measurements are calibrated using the multiport calibration module, which enables you to calibrate up to 48 RF ports automatically. For more information on the NI STS, visit ni.com/semiconductor-test-system.



attenuator value for S11, S12, and S21 measurements and another attenuator value for S22 measurements.

MEASURING S-PARAMETERS

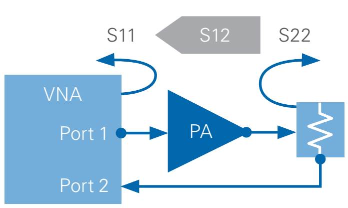


Figure 8. You can measure reverse isolation and return loss with a VNA.

When testing PAs in large signal conditions, the test configuration is substantially more complex. In large signal conditions, a substantial portion of the output energy is converted to harmonics and not captured by a traditional VNA. As a result, complete characterization of the large-signal performance of a PA requires a large-signal network analyzer (LSNA), or load-pull test bench, as illustrated in Figure 9. Because S12 and S21 measurements are more difficult in large-signal conditions, one approach is to measure S21 performance as a function of input and/or output impedance. In this scenario, a programmable tuner is placed at either the input or the output of the DUT.

BASIC LOAD-PULL TEST CONFIGURATION

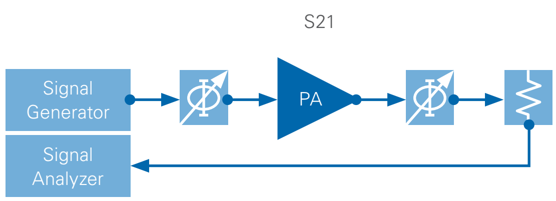


Figure 9. Block Diagram of Basic Load-Pull Test Configuration

Although this method does not enable you to measure the input impedance (S11) or output impedance (S22) directly, it does enable you to estimate the input/output impedances that result in the best PA performance or efficiency through trial and error. Note that the classic configuration involves using a CW signal generator to source power and a power meter to measure it. Today, it is now possible to measure large-signal performance using modulated signals sourced and measured with VSGs and VSAs.

Noise Figure

Although gain and output power are some of the most critical measurements of power amplifiers, noise figure remains the most crucial measurements of low noise amplifiers (LNAs). Because the LNA is typically the first component in a receiver, the noise figure and gain of the LNA drives the noise figure of the receiver.

There are a number of ways to quantitatively describe both noise figure and noise factor. One of the earliest definitions is the one proposed by Harold Friis in the 1940s. In Friis's definition, noise factor (the linear equivalent of noise figure) is the degradation of a particular signal's signal-to-noise ratio (SNR) as it passes through a particular component. Noise factor and noise figure are inherently unitless ratios, and while noise factor expresses this ratio in liner terms, noise figure expresses this ratio in logarithmic terms.

$$F = \frac{SNR_{in}}{SNR_{out}} \quad (1)$$

Equation 1. Noise Factor as a Function of SNR

Using Equation 1, if a signal had an SNR of 100 dB at the input of an LNA with a noise figure of 5 dB, the SNR at the output would be 100 – 5 dB = 95 dB. As Figure 10 illustrates, a "black box" component with a noise figure of X_{dB} would degrade the SNR by X_{dB}.

INTRINSIC NOISE POWER OVER THERMAL NOISE

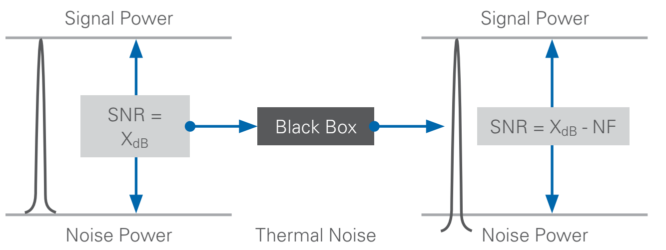


Figure 10. Noise figure is the addition of a component's intrinsic noise power on top of thermal noise.

Another view of noise figure is that it describes the noise power, in dB, that a particular passive or active component adds on top of room temperature's thermal noise of -174 dBm/Hz. This definition closely mirrors the widely accepted IEEE definition of noise factor, which is defined in Equation 2.

$$F = \frac{N_{added} + kT_o BG}{kT_o BG} \quad (2)$$

where k represents Boltzman's constant
T_0 represents room temperature
B represents bandwidth
G is the gain of the DUT

Equation 2. Formal Definition of Noise Factor

In Equation 2, kTo is simplified to the thermal noise at room temperature, or -174 dBm/Hz. Thus, noise factor is a component's noise power added on top of signal power.

For example, in a scenario where an antenna is connected to an LNA, the noise power at the input of the LNA is -174 dBm/Hz. At the LNA's output, the noise power is -174 dBm/Hz plus the noise figure of the LNA. In this scenario, a noise figure of 5 dB would yield an output noise power of -169 dBm/Hz. Note that in this case, you can simply add 5 dB to -174 dBm/Hz because we are describing noise figure in logarithmic terms.

Noise Unit Conversion

Before we describe noise figure measurements in detail, it is useful to first define some units and terms commonly used to describe noise measurements. Some of the most common metrics include noise figure, noise factor, and noise temperature.

Noise figure (NF) describes the noise power a component adds on top of thermal noise in dB, and *noise factor (F)* describes the noise power a component adds on top of thermal noise in linear terms. You can convert NF to F and vice versa using Equations 3 and 4.

$$NF = 10\log_{10}(F) \quad (3)$$

$$F = 10^{\frac{NF_{dB}}{10}} \quad (4)$$

Equations 3 and 4. Conversion of Noise Factor to Noise Figure and Vice Versa

A related expression of noise power is noise temperature. Because noise power is directly proportional to the temperature of the device in Kelvin, the noise temperature (T_e) is the equivalent temperature of a device that produces a certain amount of noise power. It is important to recognize that the equivalent noise temperature of a device is a theoretical value that merely describes the theoretical temperature at which a passive device produces a particular noise power level. You can relate noise temperature to noise factor using Equations 5 and 6.

$$T_e = T_0\,(F-1) \quad (5)$$

Equation 5. Noise Temperature as a Function of Noise Factor

$$F = \frac{T_e}{T_0} + 1 \quad (6)$$

Equation 6. Noise Factor as a Function of Noise Temperature and Vice Versa

In Equations 5 and 6, T_0 is an expression that commonly refers to room temperature, or 290 K. Given these equations, a component with a noise factor of 4, or a noise figure of 6.02 dB, would have an equivalent temperature of 290 K (4 – 1) = 870 K. Given this calculation, the inherent thermal noise of a component that has been heated to 870 K is exactly 6.02 dB higher than a component at room temperature, which has a temperature of 290 K. Thus, having an equivalent temperature of 870 dB is the same as having a noise factor of 4 and a noise figure of 6.02 dB.

The Friis formula for the noise factor of a cascaded RF system is a final key formula that is useful for noise factor measurements. This equation is important because when measuring the noise figure of a component, you must consider that any measurement includes both the noise contribution of the DUT and the noise contribution of the instrument itself. When using the Friis formula, consider the cascaded RF system shown in Figure 11.

CASCADED RF SYSTEM

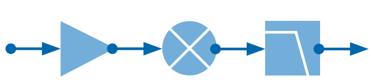


Figure 11. Each component can be described by both its gain and noise figure.

Using the Friis formula, shown in Equation 7, you can calculate the noise factor (F) of the system.

$$F_{System} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots + \frac{F_n - 1}{G_1 G_2 G_{3 \ldots} G_{n\text{-}1}}$$

Equation 7. The Friis Formula for Noise Factor of a Cascaded System

Note that the Friis formula requires that both noise and gain are expressed in linear terms and not logarithmic terms. Also, note that when the first component of the system has a high gain, like an LNA, the noise figure of the system is dominated by the first component. Thus, for typical noise figure measurements you can generally omit all but the first two terms of Equation 7 and use the simplified version in Equation 8.

$$F_{12} = F_1 + \frac{F_2 - 1}{G_1}$$

Equation 8. Noise Factor of a Cascaded Two-Stage System

Likewise, you can relate the noise temperature of the cascade using a similar relationship. By substituting the equation for noise temperature from noise factor, you can derive that the noise temperature of the first component in a cascaded system is equal to the noise figure of the system minus the noise contribution from the second element, as shown in Equation 9.

$$T_{12} = T_1 + \frac{T_2}{G_1}$$

Equation 9. Noise Temperature of a Cascaded Two-Stage System

Noise Figure Measurements

Although there are several methods for measuring noise figure, the two most common methods are the cold source method, also known as the gain method, and the Y-factor method. The basic principle of the gain method is to terminate the input of the DUT and then measure the output noise of the DUT using a signal analyzer, as shown in Figure 12. In this scenario, output noise power is the intrinsic noise of the DUT that is amplified by the gain of the DUT.

COLD SOURCE NOISE FIGURE MEASUREMENT TECHNIQUE

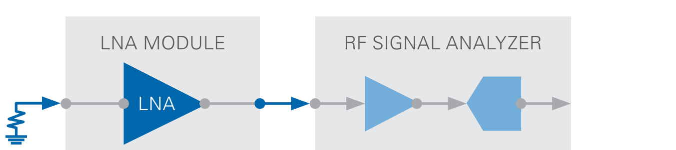


Figure 12. Terminate the input of the DUT when using the cold source method.

The cold source method is generally most effective in high-gain LNAs since signal analyzers can measure noise power more accurately for signals that are significantly above their inherent noise floors. One of the drawbacks of the cold source method is that it is most susceptible to voltage standing wave ratio (VSWR) uncertainty. In addition, traditional methods to improve VSWR, such as the use of external attenuators, degrade the instrument's ability to measure low-power signals. As a result, the cold source measurement technique is more accurate when you can compensate for VSWR. In fact, you can occasionally use a network analyzer to measure noise figure, assuming that its noise floor is low enough, because it is able to reduce uncertainty due to VSWR.

Y-Factor Method Using a Calibrated Noise Source

A second, and perhaps more common, noise figure measurement is the Y-factor method. This method involves introducing a calibrated noise source to an LNA, or a PA, and measuring noise power both when the noise source is turned on and when the noise source is turned off. The premise of the Y-factor method is actually straightforward if you consider that both the DUT and the signal analyzer are part of a two-part cascaded RF system, as shown in Figure 13.

Y-FACTOR NOISE FIGURE MEASUREMENT TECHNIQUE

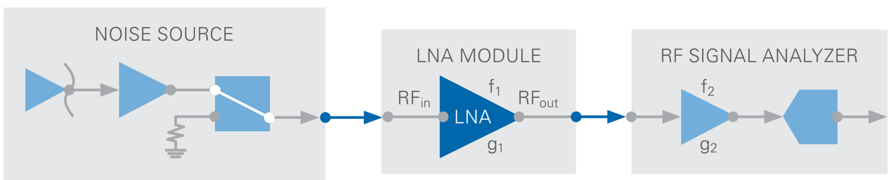


Figure 13. Connecting an LNA to a signal analyzer produces a cascaded RF system.

With a noise source, typically either an LNA or a demodulator, connected to the input of the DUT, you can model the test setup as a two-stage system. In this case, the noise figure of the system includes the noise figure of the first component, an LNA, plus the noise contribution of the RF signal analyzer. The Y-factor approach is designed to measure the noise factor of the DUT (F1) by first solving for both the noise factor of the system (F12) and the gain of the DUT (G1). Thus, the process of measuring the noise figure of an RF component using the Y-factor method involves the following two steps:

1. Measure the noise figure of the signal analyzer.
2. Measure the noise figure of the system with the DUT in place.

One of the essential elements of the Y-factor test setup is the calibrated noise source. A calibrated noise source is extremely useful when measuring noise figure because it is able to provide a noise-like signal into the DUT at a relatively low power level, with a calibrated on/off ratio.

Noise sources have two settings, on and off, and their characteristic specification is their excess noise ratio (ENR). ENR is defined by Equation 10, where Ts^{ON} and Ts^{OFF} represent the equivalent temperature, noise power, in each setting. For practical measurement purposes, you can generally assume that $Ts^{OFF} = T_0 = 290$ K. The ENR of a noise source is typically printed on the source directly or offered through a specification document, and typical ENR values range from 5 dB to 30 dB depending on the application.

$$ENR_{dB} = 10 \log \left(\frac{T_s^{ON} - T_s^{OFF}}{T_0} \right)$$

Equation 10. ENR is essentially the power ratio of noise on to noise off.

Step 1: Characterize the Noise Figure of the Signal Analyzer

The first step in measuring the noise figure using the Y-factor method is to measure the noise figure of the signal analyzer without the DUT connected. Note that typical noise sources require a 28 VDC supply that is usually provided via the 28 VDC port of the RF signal analyzer, as shown in Figure 14.

CALIBRATION STEP FOR Y-FACTOR TECHNIQUE

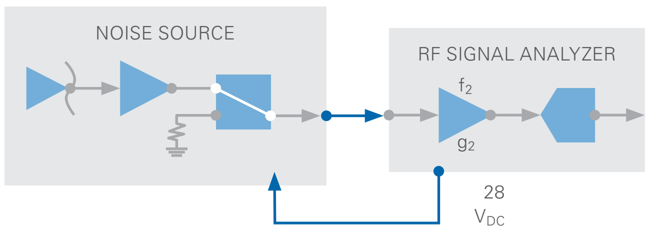


Figure 14. Connect the noise source directly to the signal analyzer to measure the signal analyzer's inherent noise figure.

With the setup shown in Figure 14, the Y-factor is the measured ratio of the noise source's output noise power when turned on (N_{on}) versus when it is turned off (N_{off}). Thus, measuring the Y-factor consists of two power measurements, N_{on} and N_{off}. Note that the ratio of N_{on} and N_{off} must be expressed in linear terms, with noise power in watts. This ratio is illustrated in Equation 11.

$$Y = \frac{N_{on}}{N_{off}}$$

Equation 11. The Y-factor is the ratio of N_{on} to N_{off}.

You can measure N_{on} and N_{off} with an RF signal analyzer using a channel power measurement. Because the accuracy of a noise figure measurement with an RF signal analyzer depends on the noise figure of the instrument itself, it is important to minimize the instrument's noise figure by taking the following steps:

1. Turn on the instrument's pre-amplifier (if available).
2. Set the reference level as low as possible, typically to less than -50 dBm.
3. Manually set the instrument's attenuation to 0 dB.

Note that for high-gain DUTs, the VSWR benefit of having the instrument's attenuation set higher than 0 dB may outweigh the noise-floor improvement of removing attenuation. Although by using the Y-factor method some uncertainty due to VSWR is theoretically removed, a small amount of error due to VSWR is present given that the signal analyzer sees different mismatch during the calibration step and the measurement step.

With the settings above, you can measure the RF signal analyzer's noise power using the power-in-band measurement. The power-in-band measurement provides a more accurate approach to measuring noise power than merely measuring noise floor with a marker. If you are measuring power in dBm, simply convert dBm to watts using Equation 12.

$$P_w = 10^{\frac{P_{dBm}-30}{10}}$$

Equation 12. Power in Watts as a Function of dBm

Because the power-in-band measurement integrates the noise power over a large number of frequency bins, the bandwidth of the measurement significantly affects the measured power result. For example, -90 dBm in a bandwidth of 1 MHz is equivalent to a measurement of -100 dBm in a bandwidth of 100 kHz. For that reason, it is often useful to express noise power in terms of dBm/Hz, as shown in Equation 13.

$$P_{dBm/Hz} = P_{dBm} - 10 \text{ x } \log \text{(Measurement Bandwidth)}$$

Equation 13. Converting Measured Power to dBm/Hz

Note that although expressing noise power in dBm/Hz is useful because of the insight it provides about the noise floor of the signal analyzer, the measurement bandwidth typically does not actually affect the Y-factor ratio unless it is wider than the bandwidth of the noise signal itself. Assuming you use the same measurement bandwidth to measure both N_{on} and N_{off}, the bandwidth units cancel each other out. The general rule of thumb is to ensure your measurement bandwidth is narrower than the output bandwidth of the noise source and equal to or narrower than the bandwidth of the signal the DUT is designed to amplify. Once you have determined the Y-factor based on the power measurements described previously, noise figure is merely a function of ENR and Y-factor, as illustrated in Equation 14.

$$NF_{SA} = ENR_{dB} - 10 \log_{10} (Y - 1)$$

Equation 14. Noise Figure as a Function of ENR and Y-Factor

Alternatively, you can also solve for the noise figure and noise factor by expressing noise in terms of noise temperature. Assuming that T_0 = 290 K (room temperature) when the noise source is off, the noise temperature in the on state of the noise source is a function of ENR. Using Equations 15 and 16, you can first solve for the noise temperature of the noise source based on its ENR, and then use that value in conjunction with the measured Y-factor to determine the noise temperature of the signal analyzer.

$$T_{source}^{ON} = \mathrm{T}_0 \, (10^{\frac{ENR}{10}})10 + T_{source}^{OFF}$$

$$T_{SA} = \frac{T_{source}^{ON} - (Y_{SA} \, x \, 290)}{Y_{SA} - 1}$$

Equations 15 and 16. You can use the Y-factor to determine the noise temperature of the signal analyzer.

Step 2: Insert the DUT

Once you've solved for the noise figure/factor/temperature of the RF signal analyzer by connecting the noise source directly to the signal analyzer, the next step is to measure the noise figure of the system with the DUT in place. To do so, connect the output of the noise source to the input of the DUT, as shown in Figure 15.

MEASUREMENT STEP FOR Y-FACTOR TECHNIQUE

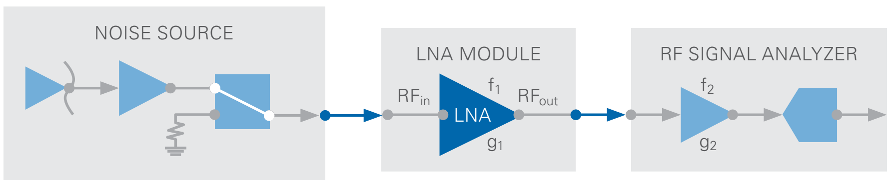


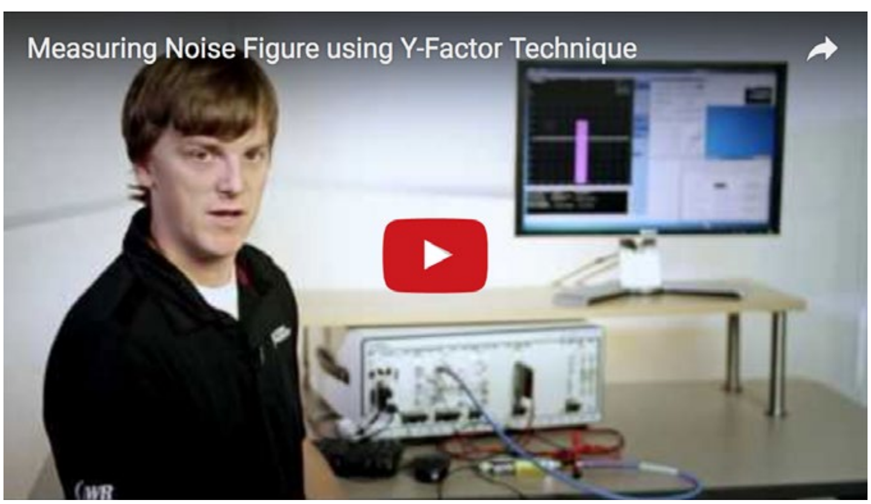
Figure 15. Insert the DUT to measure the noise figure of the RF system.

With the DUT inserted between the noise source and the signal analyzer, terms such as F_{12}, G_{12}, and T_{12} refer to the noise factor, gain, and temperature, respectively, of the entire system. Similar to the calibration step, you must next calculate the Y-factor of the entire system. In this step, you measure the Y-factor of the system or cascade, with the end result of calculating Y_{12}.

$$Y_{12} = \frac{N_{12}^{ON}}{N_{12}^{OFF}}$$

Equation 17. The Y-factor of the system is the ratio of measured noise with the DUT inserted.

Again, you can calculate either the noise figure or the noise temperature of the system using either Equation 18 or 19, respectively.



$$NF_{12} = ENR_{dB} - 10 \log_{10} (Y_{12} - 1)$$

Equation 18. Calculating Noise Figure in dB

$$T_{12} = \frac{T_{source}^{ON} - (Y_{12}\ x\ 290)}{Y_{12} - 1}$$

Equation 19. Calculating Noise Temperature in Kelvin

Once the noise figure (NF_{12}) or the noise temperature (T_{12}) of the entire system is known, it is possible to determine the noise figure of the DUT by applying the Friis formula.

Step 3: Calculate the Noise Figure

Once you have measured the noise figure, or noise factor, of both the signal analyzer alone and the measurement system with the DUT in place, you are almost ready to solve for the noise figure of the DUT. The last remaining step before you can do this is to calculate the gain of the DUT, as shown in Equation 20.

$$G_{DUT} = \frac{N_{12}^{ON} - N_{12}^{OFF}}{N_2^{ON} - N_2^{OFF}}$$

Equation 20. Calculating the Gain Based on All Four Noise Power Measurements

With the system noise figure (F_{12}) and DUT gain (G_1) both known, you can solve for the noise figure of the DUT using the Friis formula, as shown in Equation 21. Note that the Friis formula expresses noise factor in linear terms, so you must convert any units of gain or noise figure to linear terms.

$$F_1 = F_{12} - \frac{F_2 - 1}{G_1}$$

Equation 21. Calculating the Noise Factor of the DUT Using the Measured Results

Alternatively, if you've kept all the measurements in terms of noise temperature, you can solve for the noise temperature of the DUT using Equation 22.

$$T_1 = T_{12} - \frac{T_2}{G_1}$$

Equation 22. Calculating the Noise Temperature of the DUT Using the Measured Results

Again, the gain must be expressed in linear terms. Once the equivalent noise temperature of the DUT (T_1) is known, you can convert it to noise figure using Equation 23.

$$F_1 = 1 - \frac{T_1}{T_0}$$

Equation 23. Converting Noise Temperature to Noise Factor Assuming T_0 = 290 K

The Y-factor method for measuring noise figure is an accurate method to measure the noise figure of an LNA or even a PA. Although it requires some mental exercise to think in terms of noise figure, noise factor, and noise temperature, you can easily measure noise figure accurately with basic knowledge.

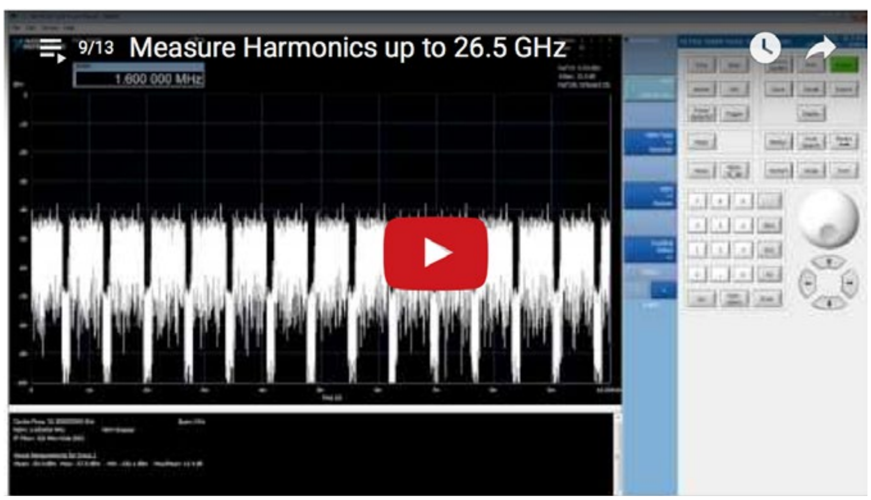
Harmonics

A second key attribute of active PAs and FEMs is their harmonic behavior. Harmonic behavior is caused by nonlinear operation and results in output power at frequencies that are a multiple of the transmit frequency. Because many wireless standards prescribe strict specifications for out-of-band emissions, engineers measure harmonics to assess whether a PA or FEM violates these emissions requirements.

The precise method of measuring harmonic power often varies according to the intended use of the PA. For devices such as a general purpose PA, harmonics measurements involve stimulating the DUT with a CW signal and measuring the power of the resulting tone at various frequencies. By contrast, when testing wireless handsets or base station PAs, harmonics measurements often require a modulated stimulus. In addition, measuring harmonic power often requires special attention to the bandwidth characteristics of the signal.

Harmonics Using a CW Stimulus

Harmonic measurements using a CW stimulus require a signal generator and a signal analyzer. On the stimulus side, you configure the signal generator to produce a CW tone at the desired output power and frequency. With the signal generator producing a stimulus, the signal analyzer measures the output power at multiples of the input frequency. Common harmonic measurements are the third and fifth harmonics, which are measured at 3X and 5X the stimulus frequency, respectively.



With an RF signal analyzer, you can use one of several measurement techniques to measure the output power of the harmonic. One straightforward approach is to tune the analyzer to the expected frequency of the harmonic and perform a peak search to find the harmonic. For example, when measuring the third harmonic of a PA when generating a 1 GHz tone, the third harmonic occurs at exactly 3 GHz.

A second method to measure harmonic power is to use the signal analyzer's zero span mode and perform the measurements in the time domain. Configured in zero span mode, the signal analyzer effectively performs a series of power-in-band measurements and displays the results as a function of time. With this mode, you can use some of the built-in averaging capabilities of the signal analyzer by measuring the power over a gated window in the time domain.

Harmonics Using a Modulated Stimulus

In practice, many PAs are designed to amplify modulated signals, and the harmonic performance of these PAs requires a modulated stimulus. Similar to the CW case, you introduce a stimulus at a known power to the input of the PA, generally at a power level close to the device's saturation point.

When measuring the output harmonic power, engineers often use a range of methods according to constraints such as measurement time and required accuracy. In practice, wireless standards such as 3GPP LTE and IEEE 802.11ac do not specifically prescribe a harmonics requirement. Rather, they specify maximum spurious emissions requirements over a range of frequencies. For example, 3GPP LTE dictates that an LTE transmitter may not emit power exceeding -30 dBm within a bandwidth of 1 MHz at frequencies over 1 GHz. In this case, validating that the PA does not cause a transmitter to exceed this limit requires engineers to measure the emissions in a 1 MHz bandwidth at the harmonic frequencies.

In practice, engineers use a range of methods to ensure that a PA does not violate the spurious emissions requirements. In an R&D or characterization lab, it is common for engineers to measure spurious emissions directly with a spectrum signal analyzer or a vector signal analyzer. However, in a manufacturing environment where test time is critical, it is more common to measure the power of the harmonic directly and use statistical correlation to predict whether or not the PA violates the spurious emissions requirement.

Measuring the harmonics of modulated signals requires careful attention to the measurement bandwidth. This is because the required measurement bandwidth of the harmonic changes as a function of the harmonic. For example, when testing the output harmonics of a PA that requires N MHz measurement bandwidth, the measurement bandwidth of the third harmonic must be 3 * N MHz, and the measurement bandwidth of the fifth harmonic must be 5 * N MHz. For example, Figure 16 illustrates that the bandwidth of a harmonic increases along with the order of the harmonic.

BANDWIDTH OF HARMONIC

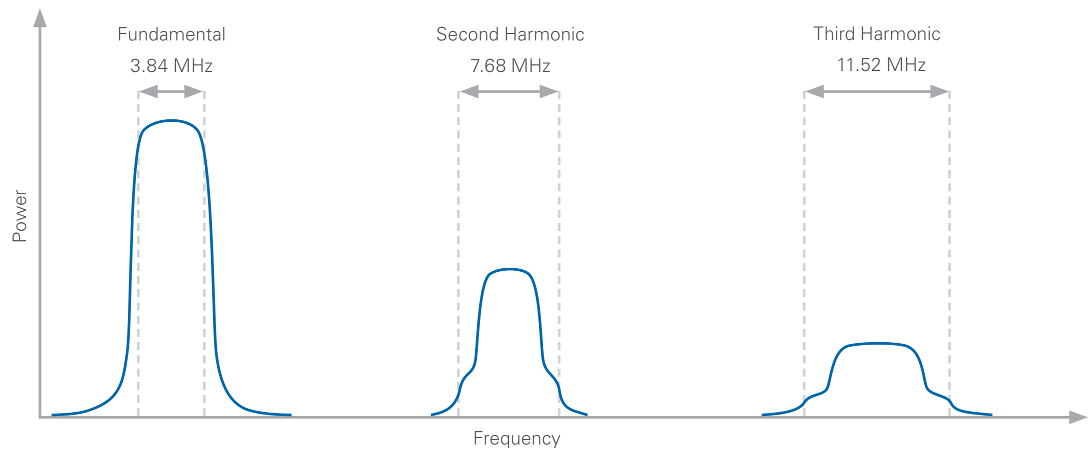


Figure 16. Bandwidth increases with the harmonic order.

Given the wide bandwidth requirements of the harmonics of modern communications signals, engineers can measure harmonics in either the time or frequency domain, depending on the instantaneous bandwidth of a signal analyzer. Time domain harmonics measurements using the zero-span mode of the signal analyzer are preferable but are not always realistic. For example, accurately measuring the third harmonic of a 160 MHz 802.11ac signal requires 480 MHz of instantaneous bandwidth. In this case, it is important to either generate a non-bursted stimulus or carefully configure the signal analyzer's power trigger to ensure that each acquisition is measuring an equivalent snapshot of the bursted signal.

Note that the specifications of cellular standards such as GSM, UMTS, and LTE provide specific guidance on the maximum spurious emissions of a transmission rather than the harmonic power itself. As a result, many engineers characterize wireless PAs according to the spurious emissions limit in addition to the actual harmonic.

Intermodulation Distortion

Another important metric in PA linearity is intermodulation distortion (IMD). Although the IMD metric is an extremely useful tool to describe the linearity of all PAs, this metric is most commonly used on general purpose power amplifiers for which an adjacent channel power measurement is not relevant.

Theory of Intermodulation Distortion

In order to understand IMD, it is worthwhile to review the theory surrounding multi-tone signals through a nonlinear system. Although a single tone stimulus creates harmonic behavior at each multiple of the tone's frequency, the nonlinear products resulting from a multi-tone signal occur at a much broader range of frequencies.

As shown in Figure 17, the second order distortion products at the output of a PA occur at frequencies that are every multiple of the input signal frequency. These distortion products at $f_2 - f_1$, $2f_1$, $f_1 + f_2$, and $2f_2$ include the second harmonics of each of the input tones in addition to two additional distortion products at both the sum and difference of the frequencies of the input tones.

THEORY OF INTERMODULATION DISTORTION

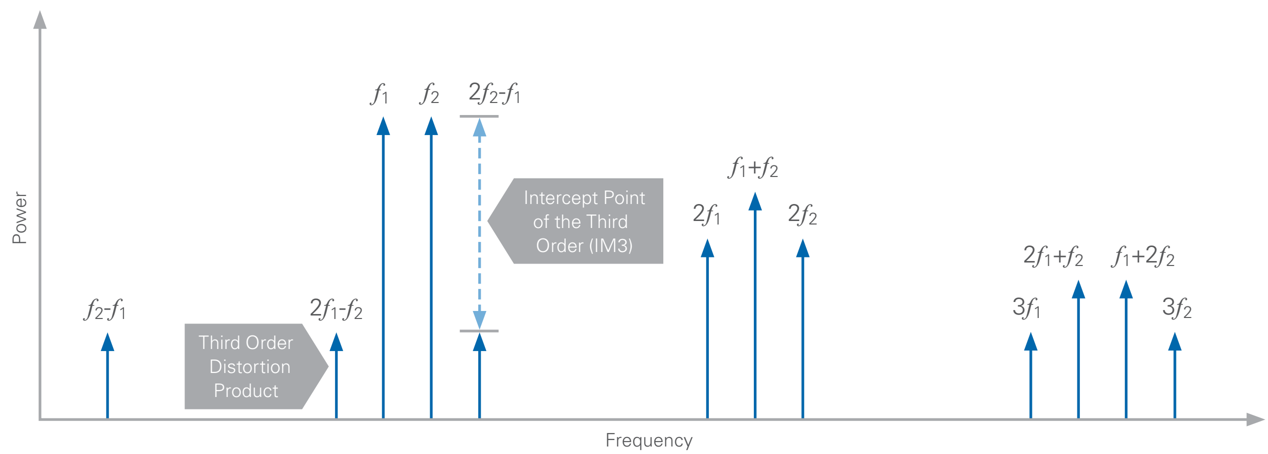


Figure 17. Theory of Intermodulation Distortion

Third order distortion products describe the interaction between the first order fundamental tones and each second order distortion product. In fact, working through the mathematics, you can visualize how two specific third order distortion products occur at a frequency that is relatively close to the fundamental tones. In a practical application, where the PA is transmitting a modulated signal, the effect of third order distortion occurs as in-band distortion that is adjacent to the band of interest.

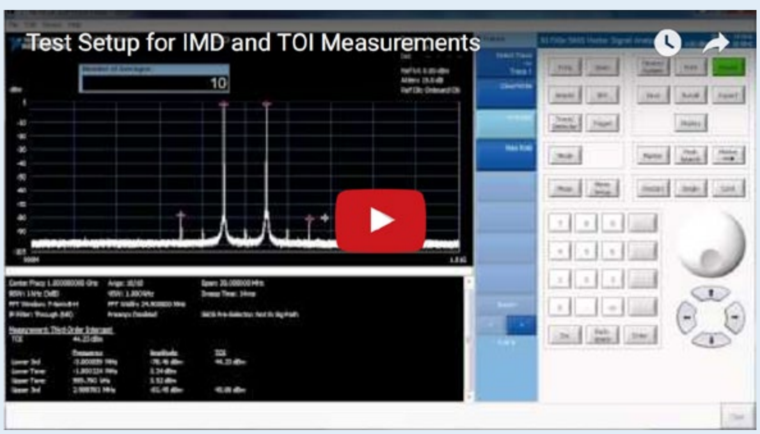
The IMD measurement describes the ratio of the power difference between the fundamental tones and the adjacent third order distortion products, and is expressed in dB. One important characteristic of the IMD measurement is that the power ratio between first order and third order distortion products is entirely dependent on the absolute power level of the tones.

In many devices, the ratio of first order tones and third order distortion products is often quite high in the linear region of operation. However, as the input power of the fundamental tones increases, the third order distortion products increase as well. In fact, IMD products should increase by 3 dB for every 1 dB increase in the power of the fundamental tones.

In theory, as third order distortion products increase in power at a faster rate than the fundamental tones, the two types of signal are eventually equal in power level as illustrated in Figure 18. The intercept point is the point at which the fundamental tones and third distortion products are equal in theoretical power. This point is also known as the third order intercept (TOI) or intercept point of the third order (IP3).

IMD and TOI Measurements Using PXI Signal Analyzers

Intermodulation distortion (IMD) and third order intercept (TOI) are built-in measurements of the NI-RFSA Soft Front Panel (SFP). When performing these measurements, you can center the signal analyzer on the two fundamental tones, ensuring that the third order distortion products are visible above the noise floor. Select detect tones in the NI-RFSA SFP to produce the measurement results. The NI-RFSA SFP automatically accounts for power differences in the fundamental tones along with power differences in the third order distortion products to present the correct measurement result. For more information on PXI RF signal analyzers, please visit ni.com/rf/test.



RELATIONSHIP OF OUTPUT POWER AND IMD

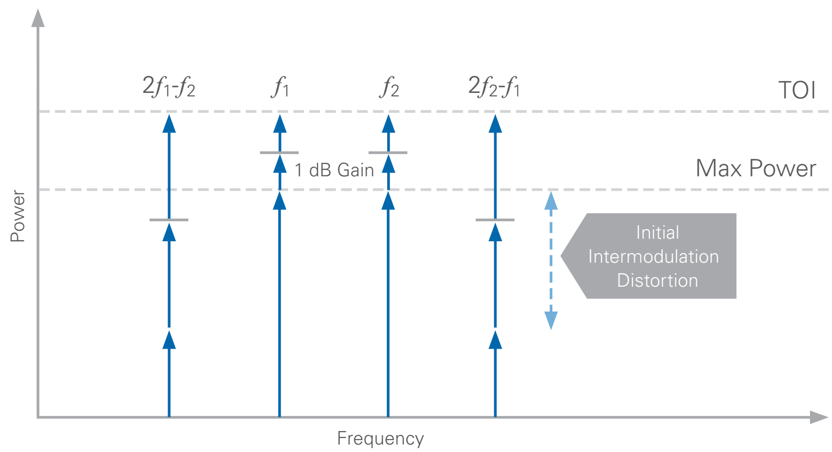


Figure 18. Every 1 dB increase in fundamental signal power produces a 3 dB increase in third order distortion products.

In practice, IP3/TOI is calculated rather than measured. Given the 3:1 ratio between the power increase of the first and third order products, you can calculate IP3 using Equation 24.

$$IP3 = \frac{IMD}{2} + P_{Fundamental}$$

Equation 24. Converting IMD to IP3

TOI is an extremely useful metric of PA performance because the IMD ratio is dependent on power level. The TOI measurement combines an element of IMD performance with absolute power level and presents the performance as a single number.

IMD Measurement Configuration

As the theory of IMD measurements suggests, performing this measurement requires a two-tone stimulus. In most applications, the preferred approach to configuring a two-tone stimulus is to use RF signal generators connected to an RF power combiner, as shown in Figure 19.

IMD MEASUREMENT CONFIGURATION

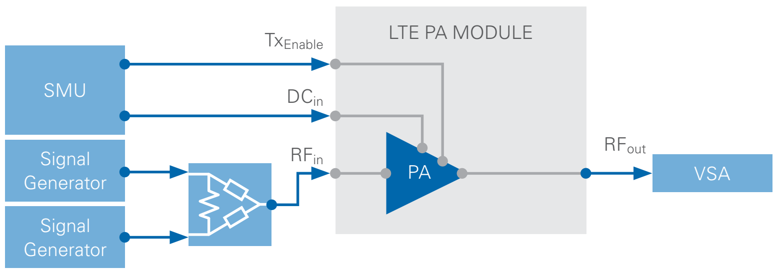


Figure 19. IMD measurements require two signal generators connected to a power combiner.

Because IMD is a common measurement, many RF signal analyzers have built-in measurement functions to measure IMD/TOI. In fact, the NI-RFSA SFP automatically detects both fundamental tones and third order distortion products and calculates the IMD ratio.

When testing high-performance PAs, it is important to generate the cleanest two-tone signal possible. In some cases, a combiner alone does not provide sufficient isolation between the two signal generators to offer a two-tone signal that is sufficiently clean. In these instances, energy from one source can leak into the other source, creating small intermodulation products that are introduced to the DUT by the test instruments.

One way to improve the isolation is to choose a combiner with a high port-to-port isolation. Generally, purely resistive combiners feature only between 6 dB and 12 dB of isolation, depending on the resistor topology. A good rule of thumb is that roughly 40 dB of isolation is required to measure IP3 numbers above +25 dBm. In the event that the combiner's isolation is insufficient, you can improve the port-to-port isolation of the combiner using either attenuators, isolators, or even an amplifier.

Assuming the source power is sufficiently high, a method to improve isolation is to introduce a pad, or attenuator, between each source and the power combiner, as shown in Figure 20. The attenuator provides additional isolation for signals traveling in the reverse direction. Additional measures to increase the isolation using either directional couplers or isolators can provide up to 50 dB of isolation if used at both ports. However, couplers are often limited to a single octave and are thus not suitable for broadband applications.

IMPROVING SOURCE ISOLATION WITH ATTENUATORS

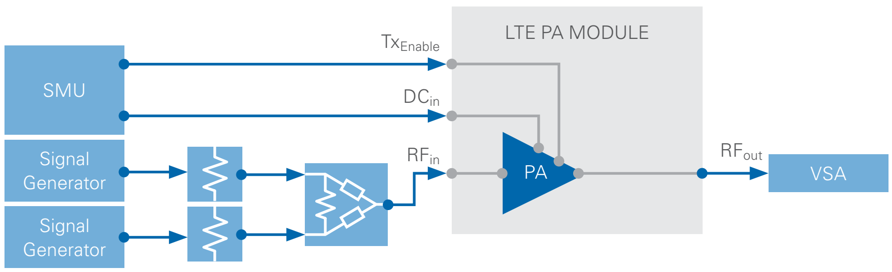


Figure 20. Attenuators can improve the quality of the stimulus signal by isolating each of the signal generators.

Amplifiers with ample reverse isolation are an excellent option when you require a high-power stimulus. In addition to providing isolation between ports, an amplifier can offer gain to the stimulus to enable the generation of a high-power two-tone stimulus.

Efficiency

Efficiency describes the ability of the PA to convert DC energy into RF energy. The two most commonly used metrics of PA efficiency are drain efficiency and PAE. Each measurement involves a signal generator, signal analyzer, and power supply or source measure unit (SMU), as shown in Figure 21.

TEST CONFIGURATION POWER EFFICIENCY

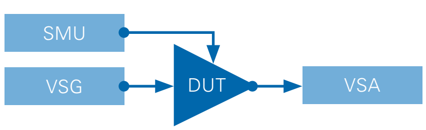


Figure 21. Configuration for Power Efficiency and Power Added Efficiency

The SMU is a critical instrument when measuring PA efficiency because of its ability to measure the DC current consumption of the PA. Typically, engineers measure PA efficiency over a range of supply voltages and use the SMU to measure current consumption at each voltage supply.

Drain Efficiency

Drain efficiency is a metric of PA efficiency and describes the percentage of the DC power that is converted to RF energy. The term *drain efficiency* comes from PA implementations that use a field-effect transistor (FET), in which the DC power is supplied to the drain of the device.

You can calculate drain efficiency by dividing the output power of the PA by the supplied DC power. Although most RF signal analyzers display measured power in watts, you can convert units of power from dBm to watts using Equation 25.

$$P_{Watts} = 10\left(\frac{P_{dBm} - 30}{10}\right)$$

Equation 25. Converting Power in dB to Power in Watts

With measured output power expressed in watts, you can calculate drain energy as the output power divided by the DC power, as shown in Equation 26.

$$Drain\ Efficiency\ (n) = \frac{P_{RF\ Output}}{P_{DC\ Supply}}$$

Equation 26. Drain efficiency is a function of RF output power and DC supplied power.

Drain efficiency is a useful metric of PA performance, but it is less useful in PAs with lower gain where the input power is often significant. As a result, a second key metric of PA efficiency that factors into the supplied input power is PAE.

Power Added Efficiency

Calculating PAE is similar to calculating drain efficiency except that it characterizes the power added by the PA instead of the power at its output. Determining PAE requires knowledge of the input power to a PA and can be calculated using Equation 27.

$$\text{Power Added Efficiency } \eta = P_{RF\ Output} - P_{RF\ InputPDrain\ Efficiency}\ (\eta) = \frac{P_{RF\ Output} - P_{RF\ Input}}{P_{DC\ Supply}}$$

Equation 27. Calculating PAE

In practical use, PAE depends on a range of operating conditions, including the bias voltage and output level of the amplified signal. As shown in Figure 22, the maximum PA efficiency occurs at or around the saturation point. In addition, Figure 22 illustrates that the PAE versus output profile is also dependent on the PA bias voltage. In general, a higher bias voltage tends to lead to higher efficiency and higher maximum PAE.

POWER ADDED EFFICIENCY FOR AN ENVELOPE TRACKING PA

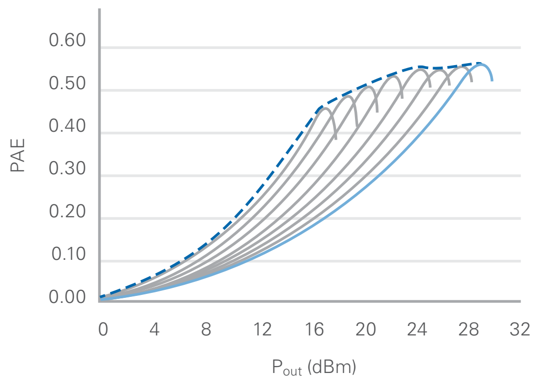


Figure 22. Power Added Efficiency (PAE) as a Function of Output Power for Various VCC Levels

Because PAE is dependent on the output power, the use of signals with higher peak to average power ratios (PAPRs) can significantly influence the PAE of the device. Modern wireless standards such as 802.11ac and LTE are based on OFDM technology, which inherently involves amplifying signals with high PAPR. Because efficiency is often highest close to the compression region of a PA and because driving a PA into compression introduces distortion, the input power of a PA must be slightly backed off when amplifying high PAPR signals. Variations in the power level result in the PA spending a smaller percentage of time operating near its compression point, and the unfortunate trade-off is a lower PAE. As a result, PA designers are increasingly looking towards techniques, such as envelope tracking, to improve overall PAE.