

How to Make High Accuracy EVM Measurements for Power Amplifiers Using the Keysight Vector Network Analyzers

Premier innovation for active device and component
characterization

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Nonlinearity Evaluation of a Power Amplifier

In wireless communication systems, the power amplifier (PA) occupies the last stage of the transmission chain to supply required RF power to antennas, as Figure 1 illustrates. The PA plays an important role in determining the quality of the communication service with respect to signal quality and battery life.

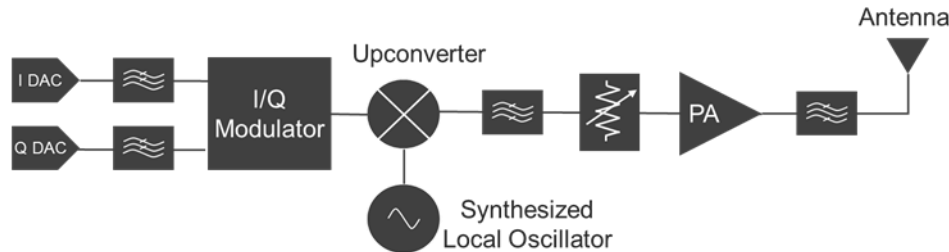


Figure 1. RF transmission chain

The PA is also the most significant contributor to the quality of the RF chain in communication systems. Linearity is critically important in systems with high peak-to-average-power (PAPR) signals, as in modern wireless standards that use orthogonal frequency-division multiplexing (OFDM) for the digital modulation scheme. The nonlinear response of the RF chain directly impacts error in the demodulation, leading to poor bit error rate. Additionally, nonlinearity can create spectral regrowth, causing interference in other bands, so maintaining the linearity of the RF chain is crucial for the quality of wireless communication.

The PA also contributes to the power consumption for wireless communication systems, especially on the user equipment (UE) side, where the power consumption or power efficiency directly impacts the quality of service for wireless communication. A PA with poor efficiency quickly consumes limited battery life, affecting the quality of service.

Since there is a tradeoff between linearity and efficiency, one of the challenges facing the PA industry is designing PAs for linearity and efficiency. The tradeoff is especially pronounced considering demanding and emerging operating conditions, such as millimeter-wave (mmWave) carrier frequency and extremely wide signal bandwidth. Digital pre-distortion (DPD) is a method of linearizing the PA by pre-distorting the input signal with the inverse of the distortion contributed by the PA, ultimately producing a linearized signal at the PA's output.

To quantify PA nonlinearity under a modulated stimulus condition, we commonly use error vector magnitude (EVM) as a figure of merit (FOM) for the in-band distortion. We use adjacent channel power ratio (ACPR) for the out-of-band characteristics. This application note introduces a new method for characterizing PA nonlinearity under a modulated stimulus condition.

Modulation distortion: Nonlinearity of the PA under a modulated signal

Intermodulation (IM) is a common parameter for quantifying PA nonlinearity. In IM measurements, we excite the PA with a two-tone stimulus, then measure intermodulation tones from both high-side and low-side tones. These tones represent nonlinearity under a two-tone stimulus condition. The IM method provides a rough idea of the nonlinear distortion of the PA, but it's important to note that the bias condition of the PA for two tones is different from its actual operation.

Modulation distortion (MOD) is an extension of IM. Instead of two tones, it uses significantly "many" tones to stimulate the device under test (DUT). A modulated waveform signal with a specific bandwidth at a given carrier frequency is a suitable example of a stimulus for modulation distortion. Assuming the PA drives a nonlinear operating condition, we observe spectral regrowth at the output, which we can represent by the equation in Figure 2:

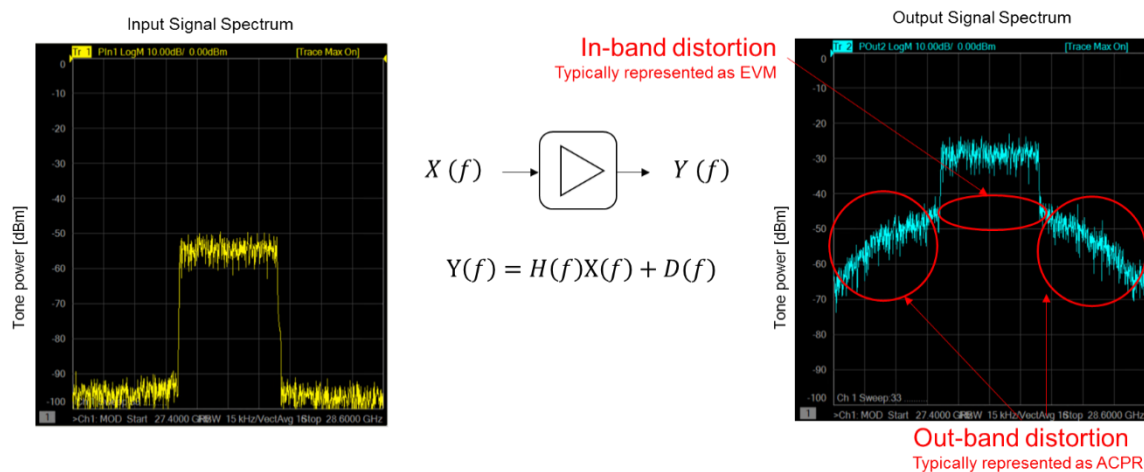


Figure 2. Nonlinear amplification of a modulated input signal

In the equation above, $H(f)$ represents a complex number that signifies a linear transfer function. In contrast, $D(f)$ illustrates a distortion parameter resulting from all the nonlinear interactions of tones present in the PA.

ACPR represents out-of-band distortion. The ACPR measurement is straightforward, comparing channel power between the signal and the adjacent channel.

The challenge arises when quantifying in-band distortion because distortion tones are part of the measured in-band signal. In the next section, we discuss a traditional method for quantifying EVM.

Traditionally, we measure EVM using time domain demodulation on a vector signal generator (VSG) and vector signal analyzer (VSA) setup, as illustrated in Figure 3. In this method, the distortion of the PA contributes to the EVM, along with many other factors, such as noise and I/Q imbalance. Also, we assume that the input signal is perfect and does not contribute to the measured EVM.

Also, the VSA method is a time domain measurement since we capture the output waveform in the time domain by the wideband digitizer.

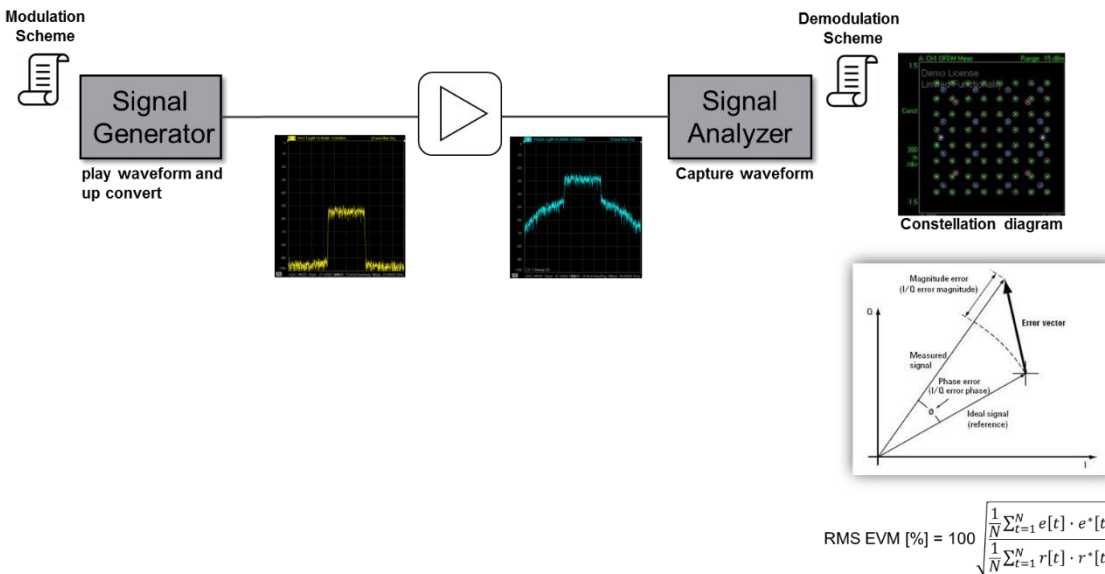


Figure 3. EVM measurement using demodulation or the VSA method

Demodulation Method

- 1) Use a specific modulation scheme (such as 64QAM) to create an IQ waveform.
- 2) Play the IQ waveform on the arbitrary waveform generator (AWG), then upconvert it to the carrier RF frequency using the VSG. The waveform stimulates the PA, which then creates an output that may contain distortion.
- 3) Capture the output using the VSA and quantify it using a wideband digitizer.
- 4) Demodulate the digitized quantity and plot the constellation diagram.
- 5) Evaluate the error vector from the measured constellation for each data point.
- 6) Compute RMS EVM by integrating the error over a certain waveform period.

Challenges in device characterization under a wideband modulated signal

With emerging standards such as 5G, we must measure EVM at mmWave carrier frequencies and extremely wide signal bandwidths. In such a measurement scenario, accurately measuring the device's EVM is increasingly challenging for the reasons appearing in the next section.

Error contributions of stimulus

In the VSA method, we evaluate the error vector between the ideal and measured signal for each constellation. Any imperfection of the signal source directly impacts the measurement result. The distortion of the generated signal must be lower than that of the DUT.

In addition, the VSA method is sensitive to I/Q imbalance and phase noise, which the DUT does not typically create. Random noise can also impact the EVM measurement result, especially when the power level is low. As the signal bandwidth (BW) widens, the signal-to-noise ratio (SNR) of the signal from the signal generator becomes narrower.

Error contributions of the receiver

To analyze a signal with minimal error, we need to digitize the signal without adding nonlinear distortion from the signal analyzer. Also, the receiver's noise floor must be lower than the target signal. However, as the signal BW widens, the SNR of the receiver becomes narrower.

Additionally, to manage the input level of the receiver chain, we carefully control the attenuation and gain settings of the receiver chain, which requires deep knowledge of the analyzer. Iterations for receiver optimization will reduce error contributions but slow down the measurement.

Signal fidelity

In the VSA method, we characterize the test system beforehand, and then we compensate the stimulus using a complex term so that the desired signal stimulates the DUT. We also compensate for the measured result using the complex term.

However, this method can introduce errors due to the widened bandwidth and the frequency response of the mismatch error between the DUT and the test system because we do not remove the DUT mismatch from the measurement.

Modulation Distortion Measurement on VNA

There is a new approach to distortion measurements that overcomes the challenges of the VSA method. Keysight's high-end VNA platforms (PNA-X, ENA-X, or PXI VNA) with a VSG can now characterize the nonlinear distortion of a device with a modulated input signal and then derive EVM and ACPR.

Modulated source such as:

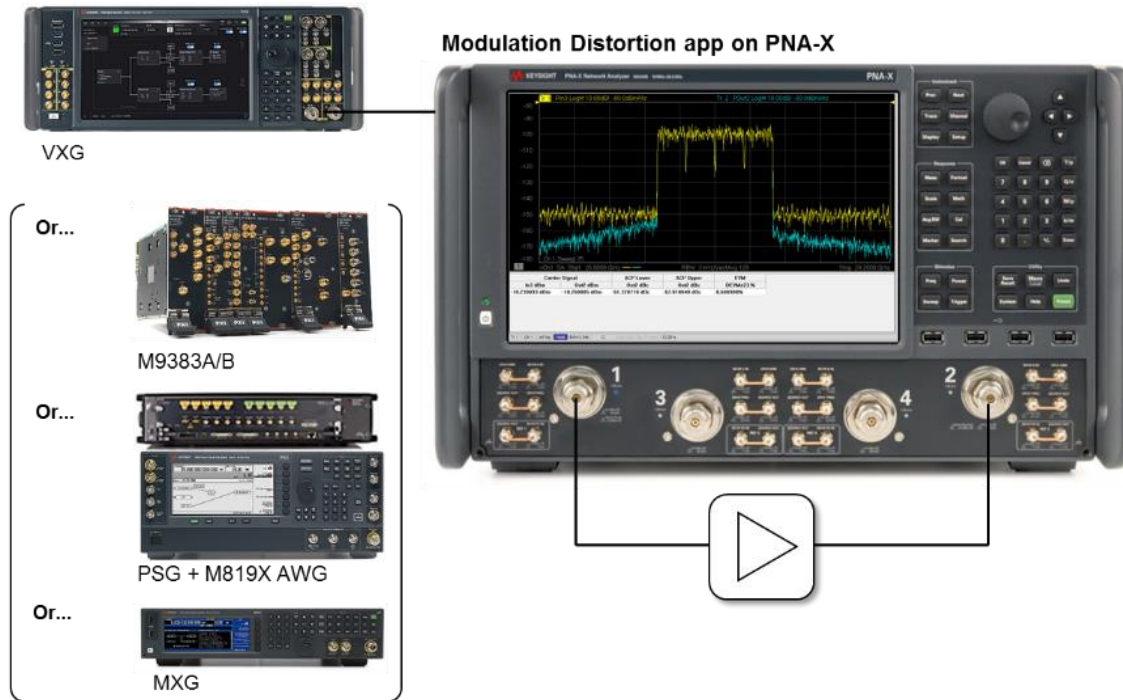


Figure 4. Distortion measurement using the Keysight PNA-X

The new approach is in the VNA's software application and is called modulation distortion, or MOD. The measurement setup comes integrated into the VNA firmware, enabling the user to quickly set up the stimulus and measurements while using state-of-the-art calibration techniques for the best accuracy.

In addition to existing VNA measurements such as S-parameter, gain compression, intermodulation distortion (IMD), and noise figure, the VNA now enables nonlinear distortion measurements under the modulated stimulus condition without requiring changes to the connections of the DUT.

In this section, we will briefly explain the underlying technology for the MOD application. A more detailed discussion analysis is available in the Appendix.

Measuring spectrum using VNA receivers

Figure 5 shows a simplified block diagram of the measurement system based on the VNA. We employ a VSG to generate the stimulus continuously, replaying the modulated signal in a loop. We use the VNA receivers to capture the spectrum at the reference plane — the input and output of the DUT. Note that the output signal spectrum exhibits spectral regrowth, which arises from the nonlinear response of the DUT.

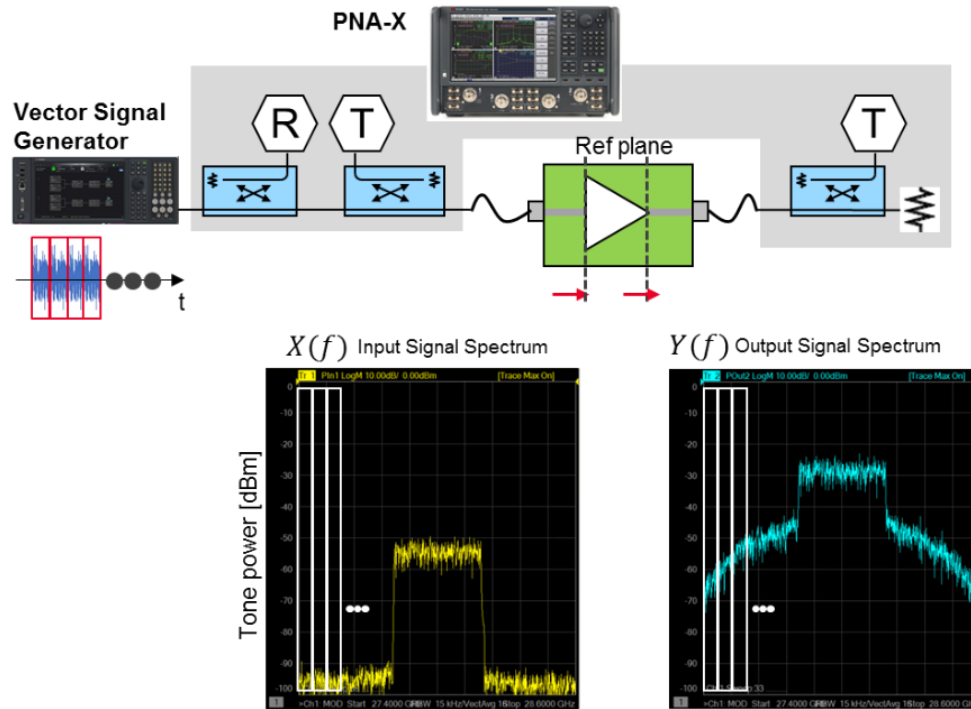


Figure 5. Measurement setup for modulation distortion using Keysight's VNA

The instantaneous bandwidth of the analog-to-digital converter (ADC) in the VNA receivers is about 30 MHz. When MOD measures the spectrum of a signal with a bandwidth greater than 30 MHz, it shifts the local oscillator frequency of the VNA receiver across the signal bandwidth. It measures the spectrum for each instantaneous bandwidth. Subsequently, MOD combines these acquisitions to obtain the entire spectrum response. The PNA-X carefully tunes the receiver's discrete Fourier transform (DFT) parameters to ensure it measures only signal tones without capturing the image.

During each acquisition, MOD uses multiple receivers coherently and applies linear calibration terms to remove the linear error of the measurement system from the raw measurement. This technique enables the VNA to achieve vector-corrected and accurate measurements at the DUT reference plane.

Spectral decomposition

The MOD application reimagines distortion measurements using the VNA to compare the input and output spectrum through spectral correlation. Spectral correlation allows the MOD application to break down the output signal spectrum $Y(f)$ into a linearly correlated part, $H(f)$, and a distortion part, $D(f)$, which appear in Figure 6. We will briefly explain the spectral correlation technique below.

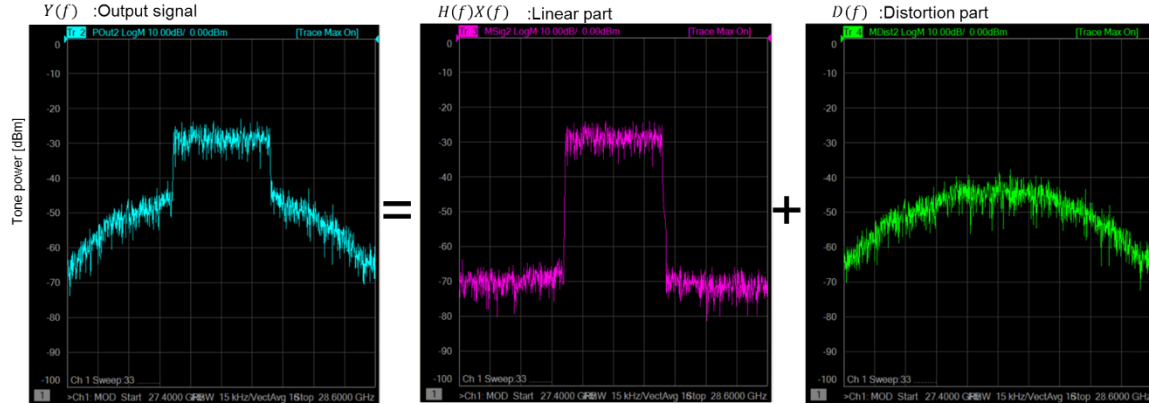


Figure 6. The modulation distortion application performs spectral decomposition

First, we start with equation (1), where $X(f)$ is the input signal, $Y(f)$ is the output signal, $H(f)$ is the linear transfer function, and $D(f)$ is the distortion:

$$(1) Y(f) = H(f)X(f) + D(f)$$

Now, we multiply both sides of the equation by the complex conjugate of X , then calculate the averaged expectation:

$$(2) E[Y(f)X^*(f)] = H(f)E[X(f)X^*(f)] + E[D(f)X^*(f)]$$

In this equation, the expectation operator is $E[.]$. The expectation is practically performed by evaluating the mean of a sufficiently large number, for example, of 100, adjacent tones. Because there is no linear correlation between the input spectrum and the distortion spectrum, (2) could be expressed as:

$$(3) E[Y(f)X^*(f)] = H(f)E[X(f)X^*(f)]$$

Which is the transfer function and the autocorrelation of $X(f)$. From here, we could compute the linear transfer function of the response in the following way:

$$(4) H(f) = E[Y(f)X^*(f)]/E[X(f)X^*(f)]$$

Now that we have determined H , we can proceed to obtain the remaining part, which is the distortion response of the DUT:

$$(5) D(f) = Y(f) - H(f)X(f)$$

Computing figures of merit (FOM)

Computing ACPR is straightforward and follows the same process as the traditional SG and SA approach. It involves evaluating the channel power of the in-band channel (BAND) of interest and the adjacent channel band (AC), then computing the ratio between the BAND and AC.

The following equation represents EVM in the MOD application of the PNA-X:

$$DEVMe (\%) = 100 \sqrt{\frac{\int_{BAND} |H^{-1}(f)D(f)|^2 df}{\int_{BAND} |X(f)|^2 df}} = 100 \sqrt{\frac{\int_{BAND} |X(f) - H^{-1}(f)Y(f)|^2 df}{\int_{BAND} |X(f)|^2 df}}$$

In practical use cases of PAs in modern communication systems, we apply compensation for linear characteristics of the PA, such as frequency dispersion or equalization. In the equation above, the suffix 'e' on *EVMe* refers to equalized EVM. Also, the prefix 'D' is added to the EVM denote that the EVM in the MOD application is computed from the distortion characteristics of the response.

The VNA measures the response of the DUT and then calculates the DFT with the ADC sampling so that the DFT frequency bins align precisely with the coherent signal tones of the modulated signal. By accumulating the power spectral density (PSD) of each tone in the desired band, it becomes possible to compute the EVM of the specified band.

Compacting the modulated waveform

In the MOD application, the measurement does not require a long stimulus waveform, such as a full frame or even a sub-frame, to measure EVM. Since no demodulation is performed, the measurement can use a signal as short as a few orthogonal frequency-division multiplexing (OFDM) symbols. The short waveform period allows an accurate measurement within a relatively short time frame. Creation of a shortened waveform is called "compacting" the waveform. For example, when a user wants to know the response of the DUT to a waveform under a specific modulation scheme or a "parent waveform," the VNA firmware takes a slice of the parent waveform, and the compact waveform inherits the frequency signature, and statistical characteristics, as illustrated in Figure 7. The slice of the waveform is called a compact test signal (CTS).

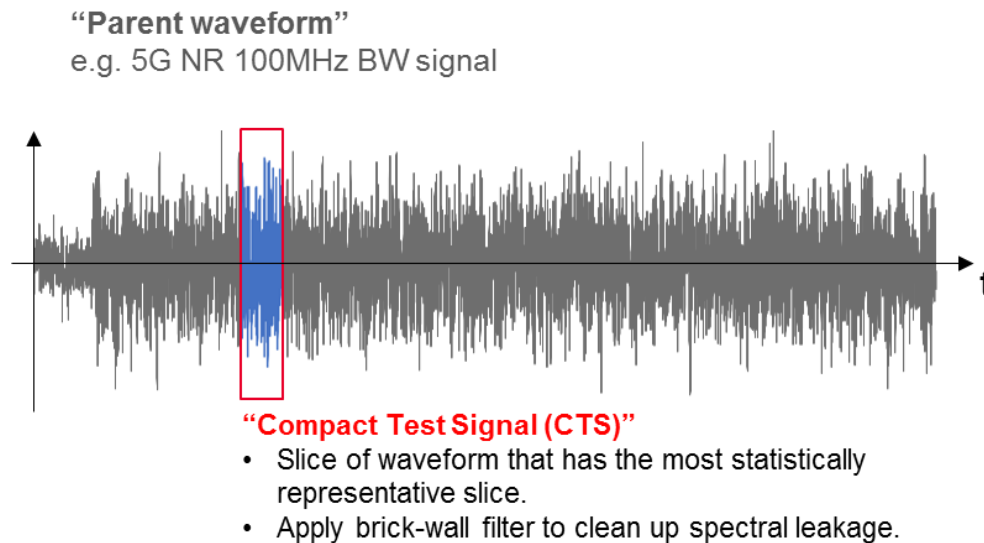


Figure 7. The compact test signal is a statistically representative slice of the parent waveform

The VNA firmware uses a unique algorithm to find the most statistically representative slice from the parent waveform based on user-provided parameters, as shown in Figure 8. Subsequently, the waveform undergoes processing, including applying a brick-wall filter to remove spectral leakage when the CTS plays over time.

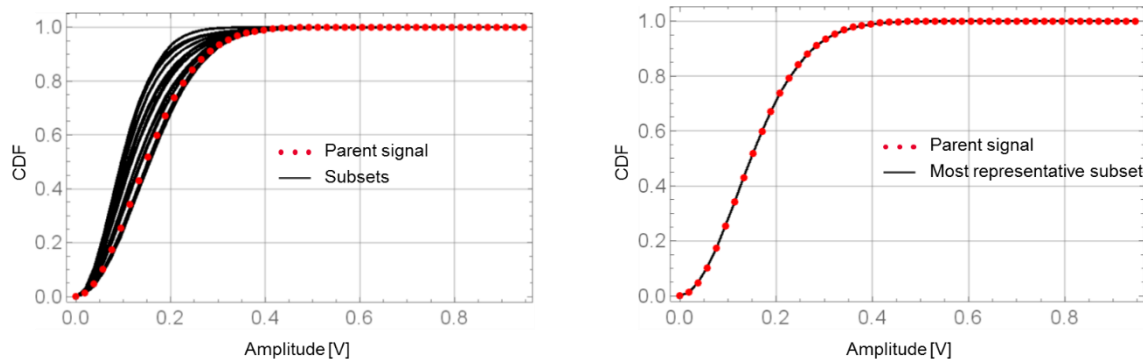


Figure 8. The VNA algorithm finds the most representative compact test signal from the parent signal

The characteristics of the CTS can vary based on user-provided parameters, which can affect the measurement result. For example, consider a 5G NR 100 MHz BW waveform as the parent waveform for creating a CTS. Figure 9 shows two CTS characteristics from different parameters applied to the same parent waveform. The yellow plot indicates the parent waveform and the blue plot indicates the CTS. The top figures show the waveform's spectrum, depicting how the waveform looks in the frequency domain. This figure materializes by performing an FFT on each waveform.

Figure 9 also shows the position of the CTS in the parent waveform. Since the signals appear noise-like, statistics can be used to characterize the modulated signals. The cumulative distribution function (CDF) shows the probability or percentage of time a signal sample falls below a certain value. Determining the likelihood is important because signals at higher levels will clip or distort. The complementary composite distribution function (CCDF) shows the signal statistics in a way that enables a quick characterization of the peak-to-average power. On the left side of the graph, there is a reference representing the signal's average value. The further the CCDF curve extends to the right, the greater the peak-to-average ratio, indicating a greater need for overhead in the amplifier design to avoid distortion.

The bottom two traces in Figure 9 show the CCDF curve for both the parent signal and the CTS, along with a Gaussian distribution in the pink trace. The signal's power is relative to the average power on the horizontal axis, measured in dB. Each line on the chart represents the percentage of time the signal remains at or above a specific power level, providing the probability for that particular power level.

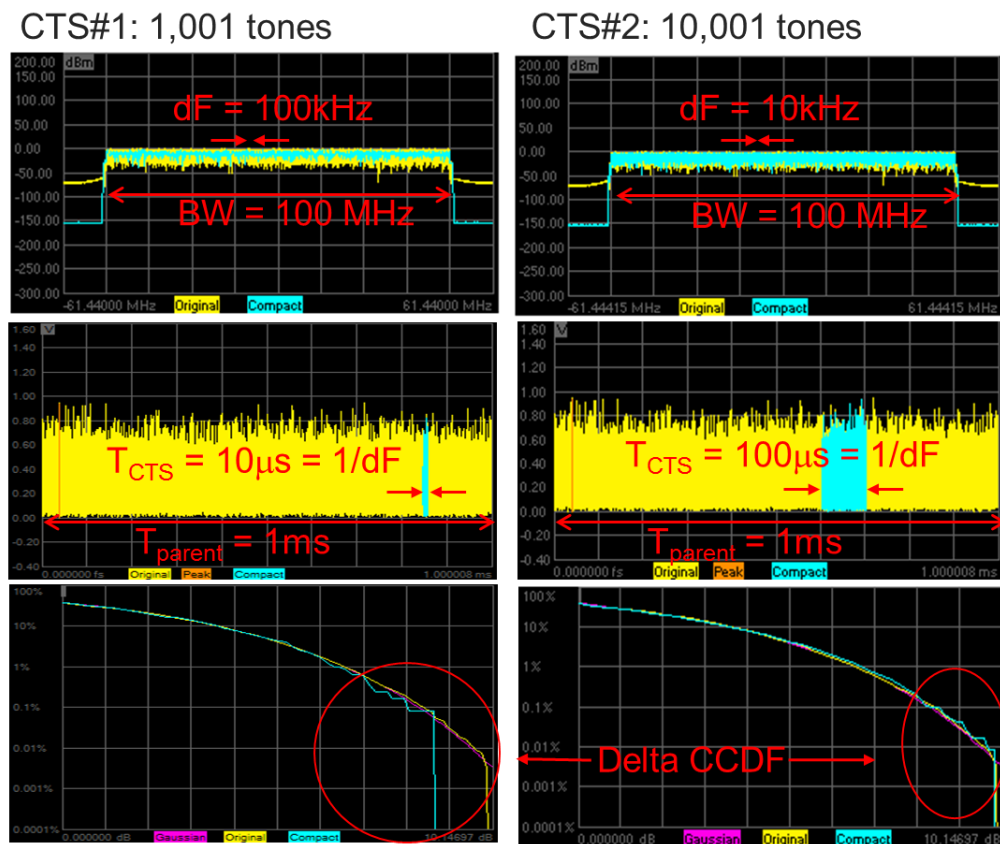


Figure 9. A comparison of CTS characteristics generated with different user-defined parameters

We can discuss the parameters of the CTS in a couple of ways, but for now, let's focus on the number of tones located in the in-band region of the modulated signal. The left CTS includes 1,001 tones, while the appropriate CTS includes 10,001 tones. To clarify, the left CTS has a tone spacing of 100 kHz, while the right CTS has a tone spacing of 10 kHz.

The first thing to notice is that both signals have the same bandwidth as the original signal. The original signal exhibits a higher out-of-band spectrum, while the CTS features a lower out-of-band spectrum, which occurs because of the brick-wall filter.

In the middle figures representing the time domain, the waveform length equals the reciprocal of the tone spacing. The left CTS has a waveform length of 10 us, while the right CTS has a waveform length of 100 us. A narrower tone spacing waveform leads to a longer period for the CTS waveform.

Finally, the bottom plots show the statistical characteristics of each waveform. We observe that the CCDF of the parent waveform aligns with a Gaussian distribution. This alignment with the Gaussian distribution reflects the noise-like characteristics inherent in OFDM. When we compare the CCDF of the CTS to the parent waveform, we find that the right CTS nicely aligns with the parent waveform. This alignment indicates that the power of the CTS waveform closely matches the average power of the parent waveform. On the other hand, the left CTS exhibits alignment only until around 0.1% probability, after which a significant discrepancy emerges, resulting in the power of the CTS waveform having lower power than the average power of the parent waveform.

In theory, the CCDF demonstrates good alignment until the reciprocal of the number of tones. To illustrate, to maintain good alignment until 0.1%, a minimum of 1,000 tones for the CTS is required.

Correlation of EVM measured in time domain to EVM measured in frequency domain

We have two distinct methods for calculating EVM: The time domain demodulation method using a VSG and VSA and the frequency domain spectral correlation method using a VSG and VNA. It's valuable to analyze the correlation between these two EVM measurement approaches.

Mathematically, both methods share an equivalent definition of EVM, as dictated by Parseval's theorem. They both focus on assessing the nonlinearity (distortion) of the DUT.

$$\sqrt{\frac{\int_{BAND} |H^{-1}(f)D(f)|^2 df}{\int_{BAND} |X(f)|^2 df}} = \sqrt{\frac{\int_{TIME} |(h^{-1} * d)(t)|^2 dt}{\int_{TIME} |x(t)|^2 dt}}$$

In both measurement methods, statistically equivalent waveforms stimulate the DUT; consequently, the DUT exhibits the same nonlinear response.

To compare the EVM results obtained in the time domain (VSG and VSA demodulation method) with those in the frequency domain (VSG and VNA MOD method), we will measure a DUT using a practical parent waveform and a CTS generated from that parent waveform.

Setup

Figure 10 shows the measurement setup for the correlation study, where a broadband power amplifier serves as the DUT. First, we used the traditional SA / SG setup to measure the DUT's EVM. The process involved creating a waveform with a specific modulation scheme and playing it from the VSG. The amplified signal from the DUT was then captured by the wideband digitizer of the UXA and subsequently processed by the VSA to compute the EVM in the time domain.

In the MOD setup, we generated the CTS using the same parent waveform as above. Subsequently, the VSG played the CTS continuously. We measured the input and output spectrum in the frequency domain and conducted spectral correlation using the PNA to compute the EVM.

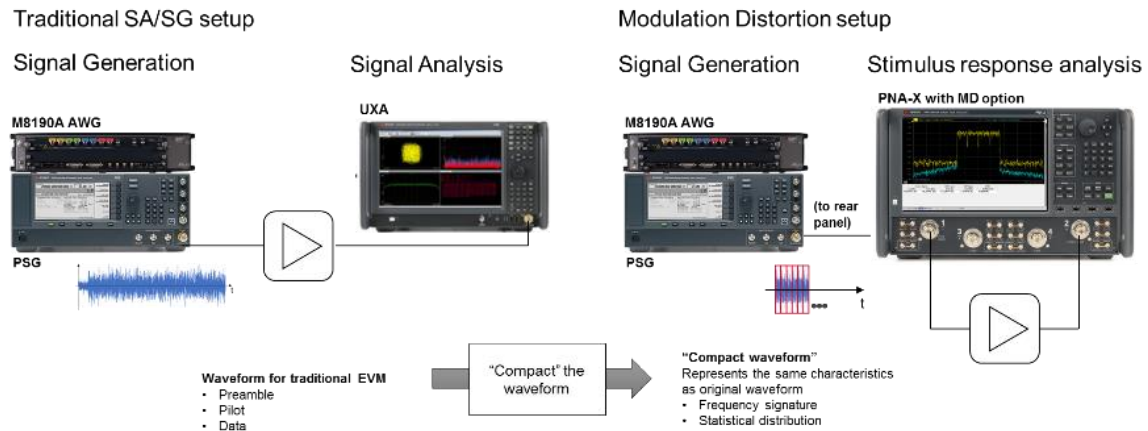


Figure 10. Measurement setup to conduct a correlation study that compares the VSA and PNA-X methods for determining EVM

Result

This study employed several waveforms: Verizon 5G 1 channel 100 MHz BW, NR 400 MHz 64 QAM, NR 800 MHz 64, and 256 QAM signals. We measured the DUT at carrier frequencies of 27, 28, and 29 GHz. The results appear in Figure 11, where we plotted EVM as a function of the output power level, known as the bathtub curve.

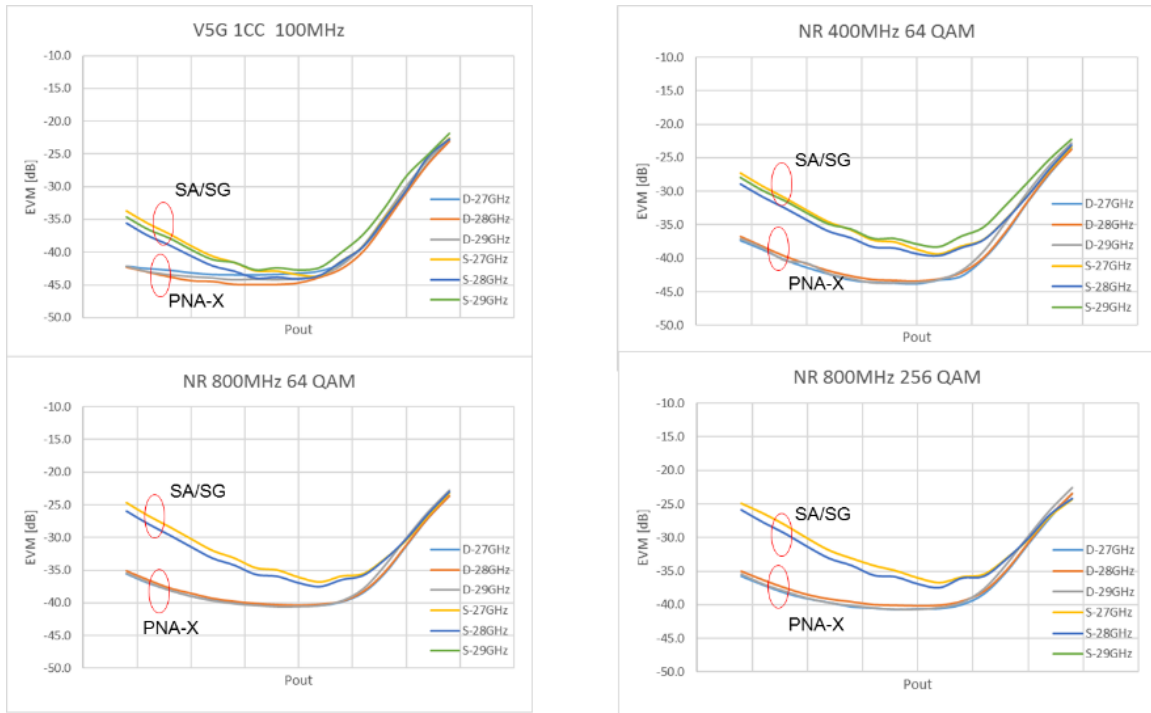


Figure 11. EVM results of a correlation study comparing VSA and VNA methods

In each plot, the right-hand side represents the high-power region, where the DUT's nonlinear response (distortion) predominantly affects EVM. In this region, both methods exhibit good alignment.

On the left-hand side of each plot lies the low-power region, where EVM is influenced mainly by the measurement system's signal-to-noise ratio (SNR). Due to the lower system noise in the MOD setup, the EVM measurement result is lower when compared to the SG/SA method. The reduced residual EVM of the test system provides users with better insight into DUT behavior, including DUT nonlinearity as it relates to output power.

When characterizing a PA for wideband applications, understanding when the PA begins to exhibit distortion with a specific modulation signal is crucial. By using the MOD application in the VNA, the user can accurately measure the performance of the PA, achieving lower residual EVM

VNA unique implementation

In this section, we will explain the VNA's unique implementation of the EVM measurement, showing how to perform measurements for a power amplifier.

PNA-X hardware configuration

The block diagram in Figure 12 shows the measurement system comprising a PNA-X and a VSG. To synchronize the PNA-X and the VSG, we employ a 10 MHz reference signal. We connect the output of the VSG to the rear panel of the PNA-X (J10) so that the signal goes through the internal path of the PNA-X. This internal path allows us to access the modulated signal from test port 1 of the PNA-X, which

connects to the DUT. Then, we connect test port 2 of the PNA-X to the DUT's output. To measure the input and output spectrum of the modulated signal, we employ receivers R1, A, and B coherently.

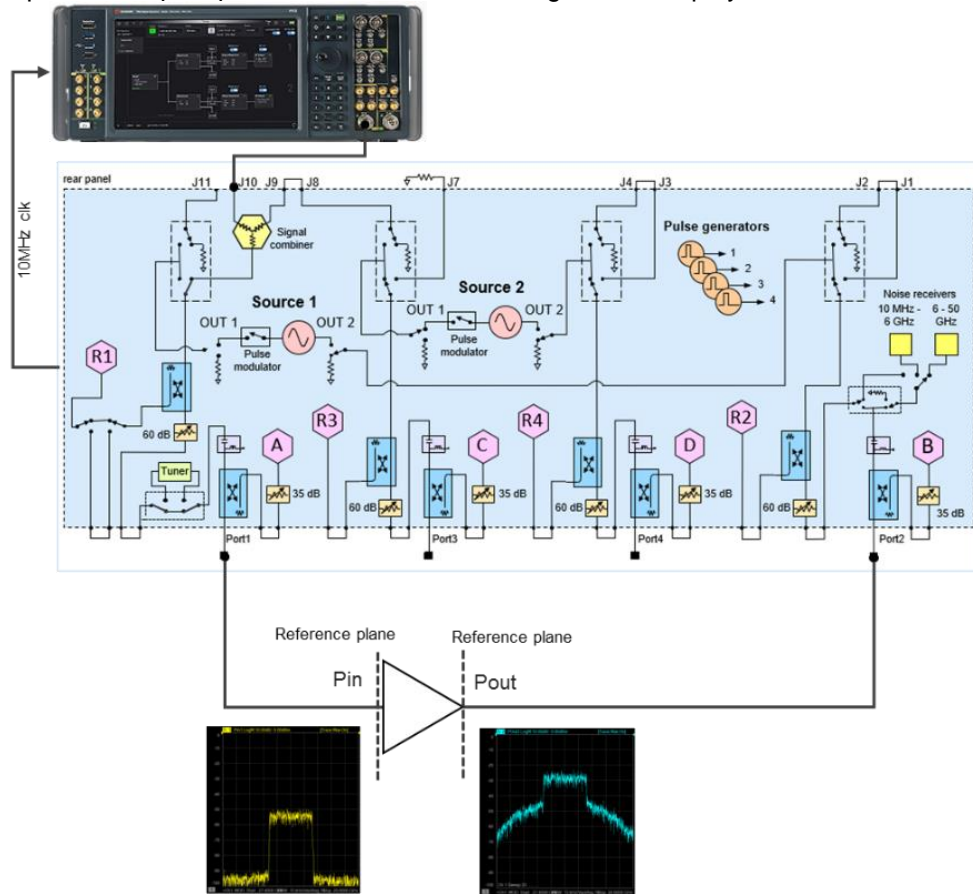


Figure 12. Measurement setup for MOD measurement using an external VSG signal generator

High-power configuration

In addition to the basic configuration in Figure 12, you can achieve flexible configurations with the PNA-X hardware by using external couplers and direct receiver access ports. Figure 13 illustrates a configuration for the PNA-X hardware and high-power amplifier testing.

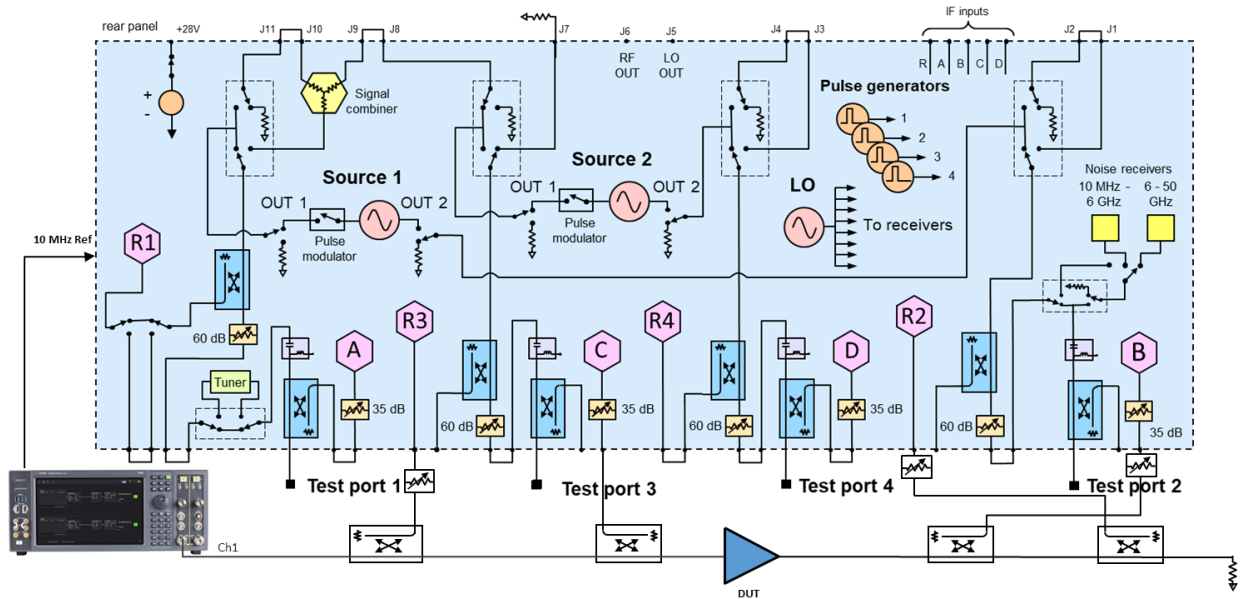


Figure 13. High-power amplifier test setup

For more information

To address the unique challenges involved with testing high-power amplifiers check out [Recommendations for Testing High-Power Amplifiers Using the PNA Microwave Network Analyzers](#) (publication number 5966-3319EN).

ENA-X hardware configuration

Figure 14 shows a block diagram of the measurement system, comprising an ENA-X and a VSG like Keysight's MXG. To synchronize the ENA-X and the MXG, we use a 10 MHz reference signal. We connect the output of the MXG to the front panel of the ENA-X (BBin), which enables us to up-convert the vector-modulated signals up to the maximum frequency of the VNA (maximum 44 GHz). The VNA has built-in upconverter hardware, enabling the up-conversion of signals below 6 GHz. This feature eliminates the need for users testing a mmWave DUT to acquire a high-frequency signal generator for generating modulated signals covering the entire measurement frequency range. You can access the modulated signal from test port 1 of the ENA-X, which connects to the DUT. The test port 2 of the ENA-X connects to the DUT's output. Measure the input and output spectrum of the modulated signal using receivers R1, A, and B coherently.

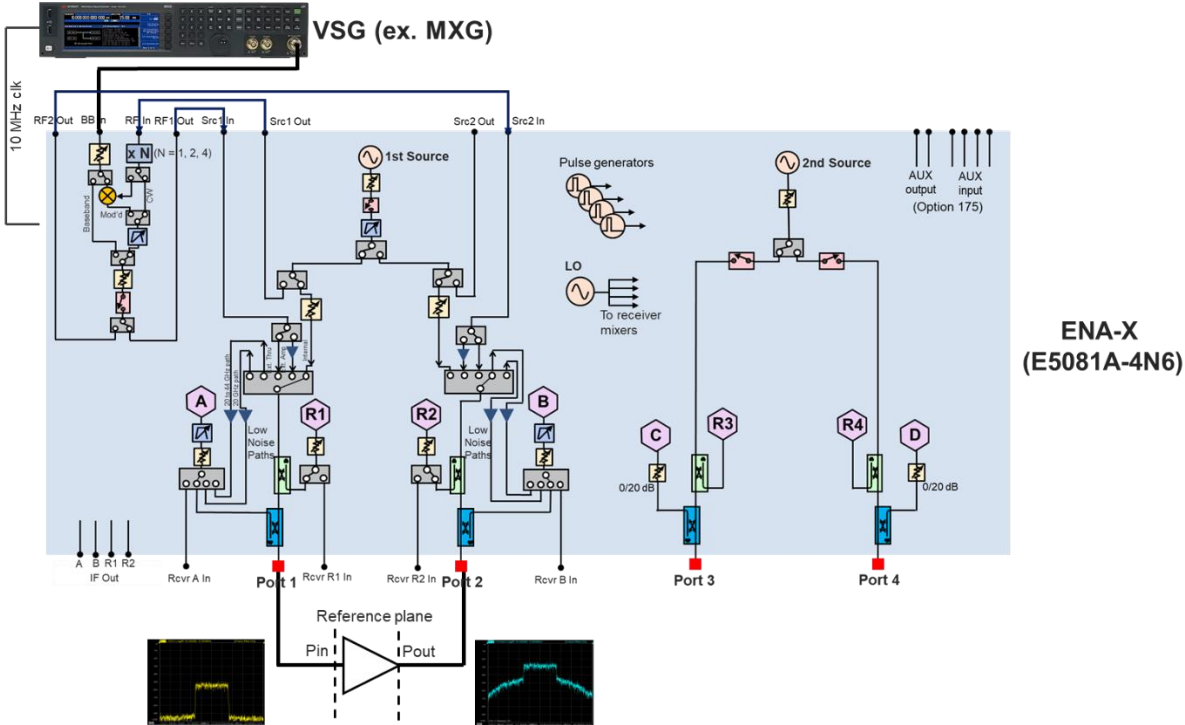


Figure 14. Measurement setup for MOD measurement using the ENA-X and MXG

PXI VNA hardware configuration

Figure 15 shows the measurement system consisting of an M9837A PXIe VNA, M9410A VXT PXI vector transceiver, and M9300A PXIe frequency reference. To synchronize the VNA and VXT, we use a frequency reference from the M9300A. We connect the output of the VTX to the front panel of the VNA (BBin), enabling the conversion of vector-modulated signals to the maximum frequency of the VNA. The VNA has built-in upconverter hardware, allowing signals below 6 GHz to undergo up-conversion. You can access the modulated signal from test port 1 of the VNA, which connects to the DUT. The test port 2 of the VNA connects to the output of the DUT. Receivers R1, A, and B measure the input and output spectrum of the modulated signal coherently.

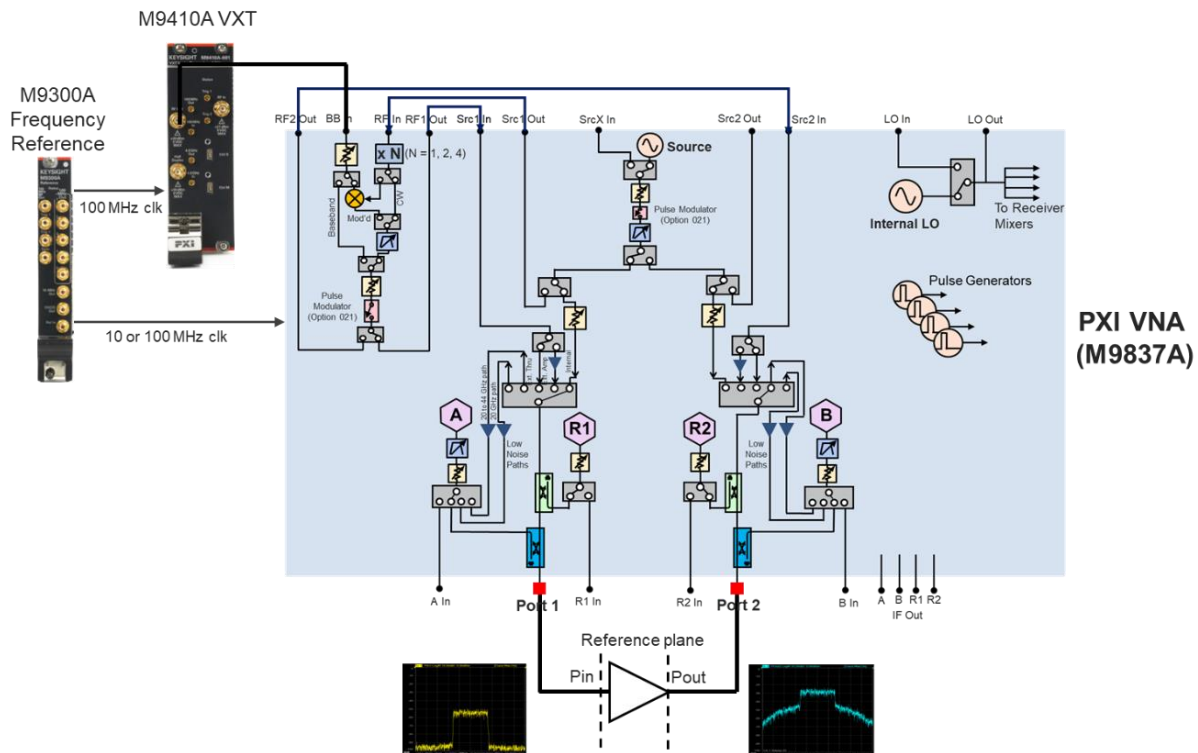


Figure 15. Measurement setup for MOD measurement using the PXI, VNA, and the VXT

Modulation distortion application

You can perform all the required configurations to conduct a distortion measurement using the MOD software application which is integrated in the VNA's firmware. In this section, we will explain the detailed setup procedure.

MOD measurement class creation

Like other VNA application software, initiate the measurement process by creating a MOD channel, as shown in Figure 16. The channel contains all the stimulus-response and calibration information required for the measurement. Like other measurement classes, you can create multiple channels for the MOD class as needed.

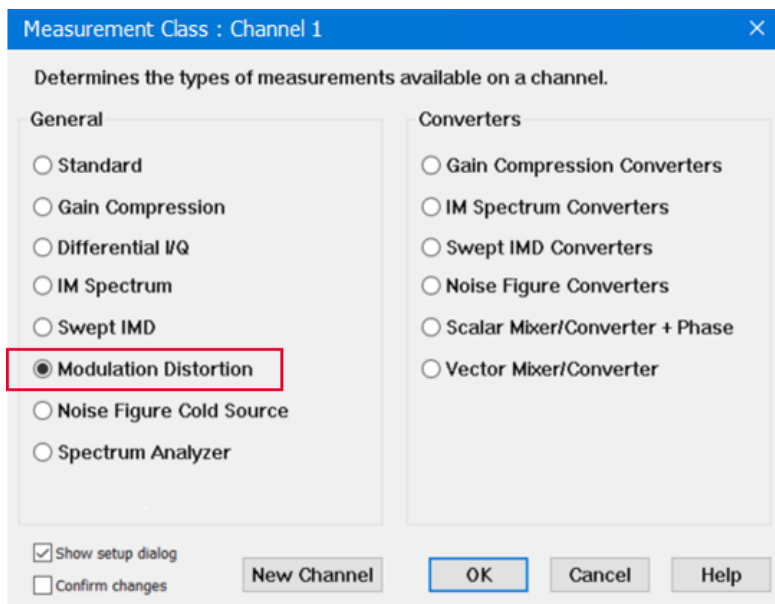


Figure 16. Creating a modulation distortion measurement class in a PNA-X modulation distortion application

Modulation distortion setup window

You can easily configure all the required settings for distortion measurements using a setup window. Figure 17 shows the setup for a measurement under the conditions outlined below.

Example measurement conditions

- Carrier frequency: 28GHz
- Span for spectrum analysis: 500 MHz
- CTS: 1001 tone from 5G NR 100 MHz BW waveform
- Measurement: EVM and ACPR for high side and low side

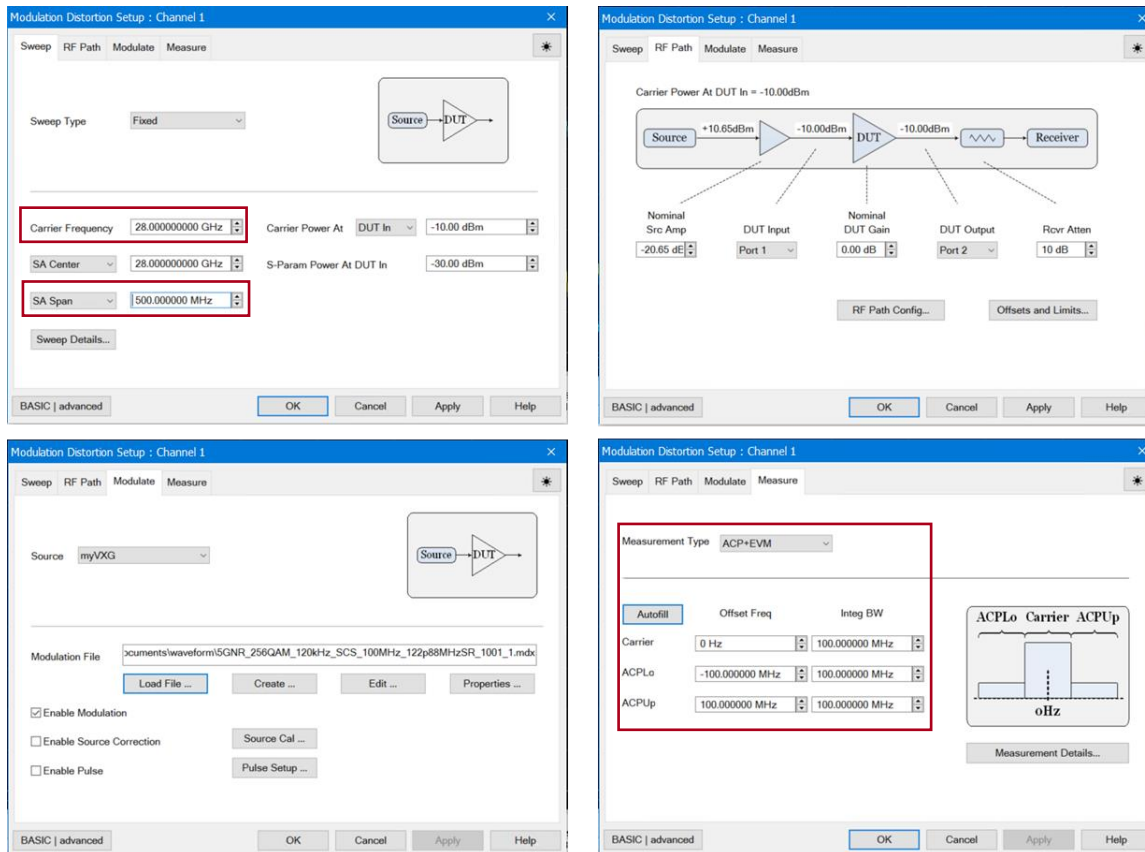


Figure 17. Modulation distortion measurement setup

You can easily specify the VSG for this measurement within the external device configuration window of the VNA firmware. In the example in Figure 17, we connected the VXG via LAN. The PNA-X communicates with the VXG via high-speed LAN instrument protocol (HiSLIP).

CTS creation

You can generate the CTS within the "Modulate" tab of the modulation distortion setup window. Figure 18 shows an example of CTS creation that uses a waveform file (*.wfm file format) generated by Keysight Signal Studio software as a parent signal. In addition, you have the option to use the *.csv file format.

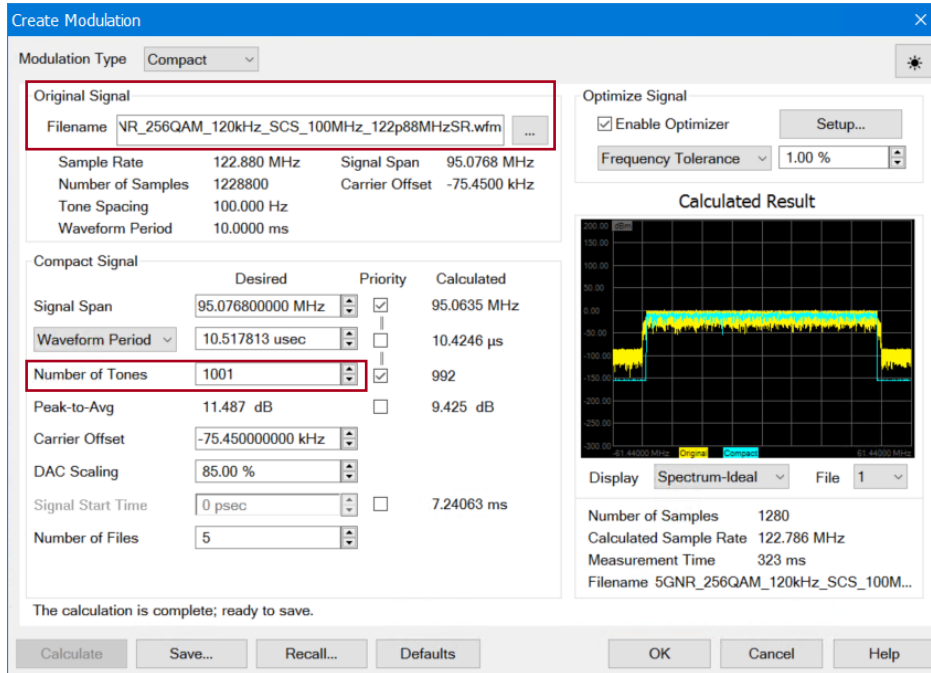


Figure 18. Generate a compact test signal from a parent waveform

In Figure 18, the parent signal has a waveform period of 10 ms and operates at a sample rate of 122.88 MHz. Based on the parameters specified in the compact signal area, the firmware generates a CTS.

In this example, we create a CTS targeting 1,001 tones. Given the provided parameters, the PNA-X uses its algorithm to find the most statistically representative slice from the parent waveform. It then performs resampling and brick-wall filtering to optimize the CTS for the PNA-X to capture the spectrum.

As a result, the modulation distortion channel created a CTS with 992 tones within the in-band spectrum, which has a waveform period of 10.4246 us.

Calibration of the system

The calibration of the measurement system plays a crucial role in achieving an accurate EVM measurement. This section will explain the calibration procedure using the setup in Figure 19.

The MOD application involves two distinct types of calibration. First, we establish a receiver calibration plane that removes linear errors from the raw measurement results. Then, we perform modulated source correction with the DUT connected to optimize the signal across the passband so the desired modulated waveform aligns with the DUT reference plane.

Receiver calibration

The receiver calibration procedure aligns with other S-parameter calibrations performed on the VNA using the "Cal All" function. You can employ conventional accessories such as a mechanical calibration kit, ECal, and power sensor. The VNA uses its internal continuous wave (CW) source during this procedure. You can use the conventional fixturing feature to adjust the calibration plane as required.

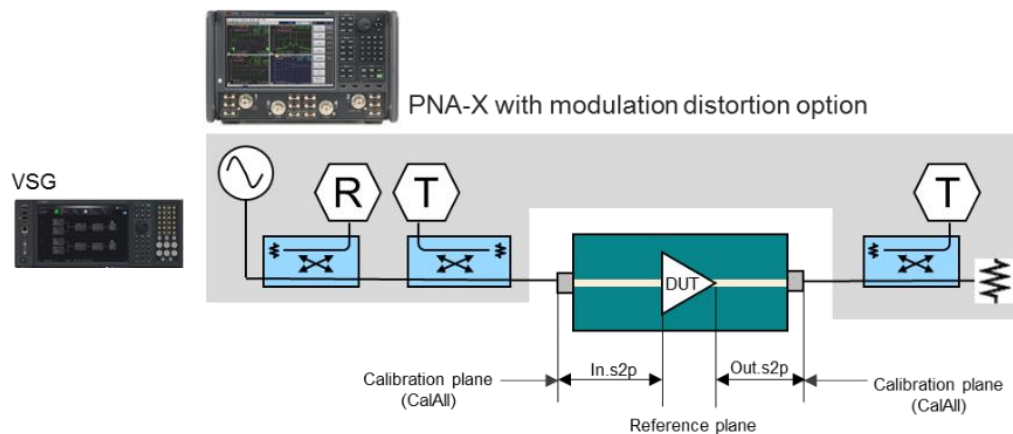


Figure 19. Calibration plane for MOD measurement

Through the calibration process, the VNA establishes a calibration plane that removes the linear errors within the measurement system when capturing the complex values of the input and output signal spectrum at the reference plane. Note that the corrected measurement removes input mismatch error the DUT and test system create, then provides accurate amplitude and phase values of the spectrum. Also, receiver calibration enables an accurate measurement of the amplitude and phase relationship between the input and output of the spectrum.

Modulated source correction

After completing the receiver correction, the measurement system can precisely capture the input and output signal spectrum at the desired reference plane shown in Figure 20. Using this system, you can perform modulated source correction so that the DUT receives stimulation from the desired modulated signal. When modulated source correction is complete, you can properly account for the DUT's mismatch in the calibration.

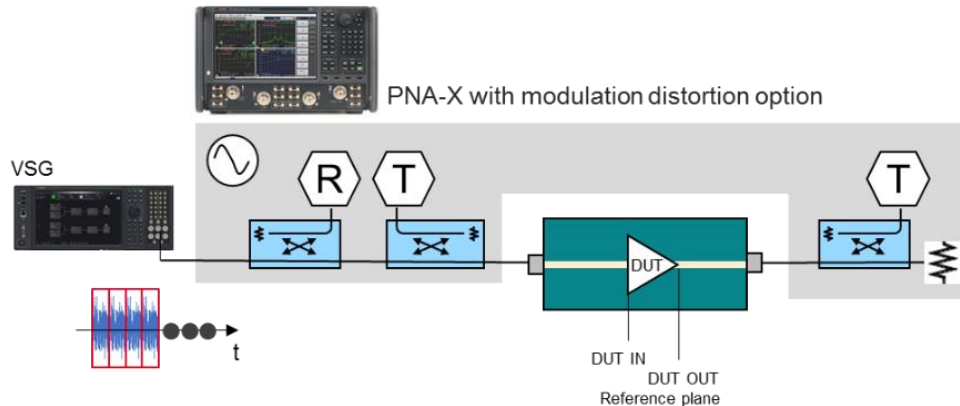


Figure 20. Calibration plane options for source correction

Choose the calibration plane from either the input (DUT IN) or the DUT output output (DUT OUT). Typically, to characterize the distortion performance of the DUT, select the input (DUT IN). Also, you can linearize the DUT response for more advanced applications by choosing the DUT output (DUT OUT).

There are different types of calibration to perform:

- 1) **Power:** Ensures the accurate calibration of the total output power of the carrier; it is integrated over the signal span.
- 2) **Equalization:** Involves a linear pre-distortion calibration that equalizes the modulated signal's magnitude and phase. The Cal Span defaults to the occupied BW of the Carrier Signal.
- 3) **LO Feedthru:** Aims to minimize the LO Feedthru tone.
- 4) **EVM:** Focuses on minimizing the vector error of the modulation signal over the Cal Span. The Cal Span defaults to the occupied BW of the Carrier Signal and includes an equalization calibration when performing a distortion calibration.
- 5) **NPR Notch:** Nullifies the NPR notch. The Cal Span gets set to the frequency range of the notch. If there are multiple notches with different spans, then Cal Span displays Various.
- 6) **ACP Upper / ACP Lower:** Nullifies the ACP upper / lower sidebands of the signal. The Cal Span is set to the frequency width of the calibrated ACP sideband. The default Cal Span is equal to the carrier span.

After completing calibration, we play the modified IQ waveform in such a way that we achieve the targeted waveform at the defined reference plane.

Performing the MOD measurement

In what follows, we explain the MOD measurement result of the setup in the previous section.

Distortion measurement

MOD measures the DUT's nonlinear response to the CTS. Figure 21 displays both the input signal spectrum (yellow trace) and the output signal spectrum (blue trace), revealing the presence of spectral regrowth in the output signal spectrum.

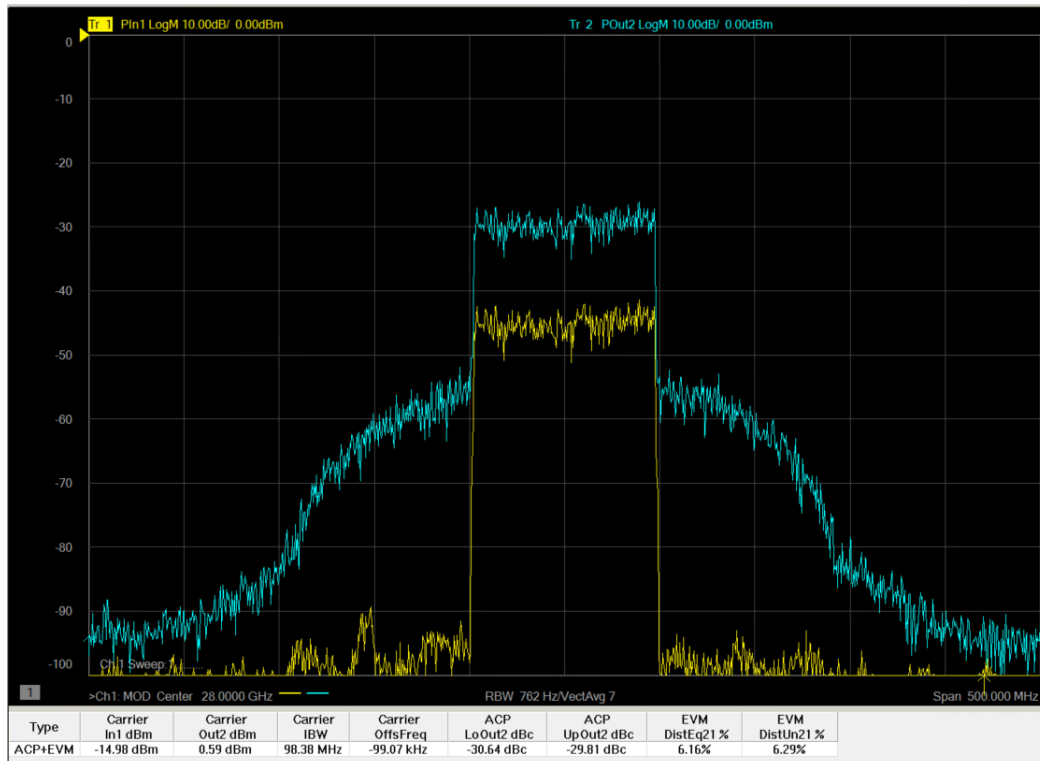


Figure 21. Measured input and output spectrum in a MOD distortion application

When you choose EVM as the measurement type in the MOD setup window, the software automatically calculates the spectral correlation between the input and output signals. The background processes include spectral correlation and signal decomposition, followed by the computation of FOMs.

The measurement in Figure 21 uses a CTS based on a 5G NR 100 MHz BW signal at 28GHz carrier frequency to stimulate a DUT with approximately 16dB of gain. A spectrum analysis was conducted over a span of 500 MHz. ACP and EVM measurements appear in the table below the input and output spectrum.

Optimizing MOD measurement

Optimizing the setup is crucial to ensure accurate measurements. In this section, we will discuss the optimization of the MOD measurement. Figure 23 shows a simplified setup for a MOD measurement, depicting the specific areas where you can create optimizations for the MOD measurement.

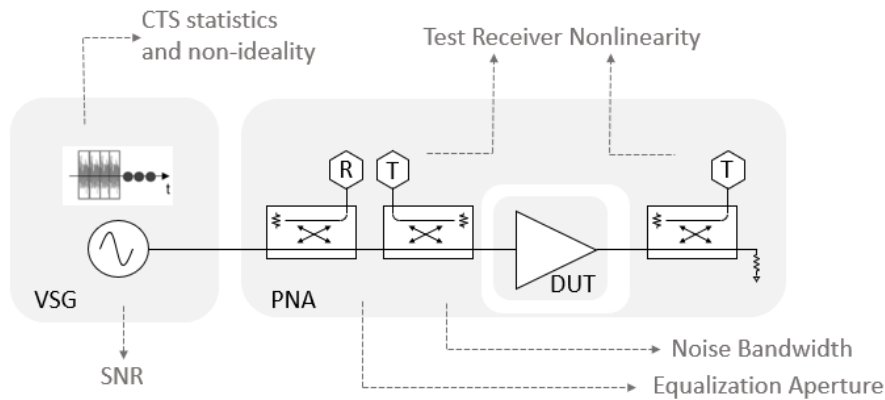


Figure 23. MOD measurement optimization

Signal-to-noise ratio of the spectrum

Optimizing the SNR of the measurement can enhance the EVM measurement. When random noise is the dominant factor contributing to the EVM, likely on the left side of the bathtub curve, improving the measurement's SNR can improve the accuracy of the measurement.

Noise bandwidth

The noise bandwidth (NBW) depends on both the resolution BW (RBW) and the number of coherent averages in the measurement. The NBW is, by default, set to approximately 1 kHz. NBW determines the SNR of the measurement system and measurement time. When the NBW decreases, the MOD increases the underlying coherent averaging, resulting in wider SNR and longer measurement time. The firmware automatically configures RBW based on the CTS waveform length, and it's a discrete number. The NBW is always a discrete number set to the closest value to the value the user enters. In scenarios where random noise dominates the EVM, EVM percentage exhibits an inverse proportionality to the square root of NBW.

SNR of VSG

Several factors can influence the SNR of the VSG, with the number of tones of the CTS being the most significant factor. Increasing the number of tones leads to a decrease in the SNR of the VSG. For example, consider comparing a modulated signal with 100 tones to one with 1,000 tones with the same signal bandwidth and channel power. Since the channel power is distributed across only 100 tones instead of 1,000 tones, the power level of each tone is 10 times bigger on the 100-tone CTS when the AWG creates the baseband IQ. When the SNR of the source is predominantly determined by the AWG, the choice of the number of tones for the CTS directly affects the SNR of the VSG.

Nonlinearity of test receiver

When there is a nonlinear response in the test receiver, it becomes challenging for the VNA to differentiate between the nonlinearity from the DUT or the test system's receiver. It's critical to optimize the MOD measurement to ensure the VNA test receiver is not overdriven when measuring the signal. Optimizing the MOD measurement becomes especially important when dealing with subtle DUT nonlinearity, such as an EVM level of 1%. In such a case, keeping the power level at the test port less than -5 dBm and using the receiver attenuator for power level adjustments is advisable.

Non-ideal CTS

We can correct linear errors attributed to the test system using the modulated source correction feature. This correction ensures that the desired CTS is precisely aligned with the DUT reference plane, involving adjustments to channel power and linear flatness response. Also, we can effectively suppress out-of-band spectral regrowth by correcting ACPR.

It is essential to know that the EVM measurement in the MOD application does not require a perfect stimulus signal. What matters most is that the stimulus preserves the statistical characteristics of the parent signal.

CTS statistical characteristics

It is essential to understand that the nonlinear characteristics of the DUT under a modulated signal condition depend significantly on the stimulus signal. While correcting the stimulus signal spectrum to align with the desired value at the DUT reference plane is essential, it's equally important to create a CTS that effectively stimulates the DUT with statistically representative characteristics relevant to the practical usage of the DUT.

OFDM is a commonly used modulation scheme in modern digitally modulated communication applications. When the waveform comes from OFDM, it exhibits a statistical characteristic represented by a Gaussian distribution.

Peak-to-average

While the peak-to-average value is a common metric for describing the statistical characteristics of a waveform, it's important to note that solely relying on peak-to-average can be risky when discussing CTS statistical characteristics. In a Gaussian-distributed signal, the peak-to-average increases as the waveform length and the number of samples (number of tones) increase. Generally, the CTS's waveform length is much shorter than the parent signal. Therefore, the parent signal's peak-to-average value is much higher than the peak-to-average of the CTS unless you check the peak-to-average priority in the CTS creation process.

In most use cases, we do not recommend enabling peak-to-average priority because of the short waveform length of the CTS. Generating the CTS with a peak-to-average equal to the parent signal can lead to overdriving the DUT. Overdriving occurs because the peak value of the CTS presents to the DUT much more frequently within the same period than the parent waveform.

Best practices involve aligning the CCDF of the CTS with the CCDF of the parent waveform. However, as discussed earlier, the CCDF cannot be the same since the CTS is a slice of the parent waveform. As a rule of thumb, aim to match the CCDF of the CTS and parent waveforms at a probability of 0.1%. You can achieve this by using a minimum of 1,000 tones for the CTS.

Measurement throughput

Reducing measurement time to increase throughput is a desirable objective. Several parameters can determine measurement time, including SA span, NBW, CTS number of tones, and reusing linear characteristics.

There is typically a tradeoff between measurement throughput and measurement accuracy. Striking the right balance of speed and accuracy is crucial, and you need to modify the parameters in the MOD setup based on the target measurement value.

Equalization Aperture

Channel Response

Let's consider data transmission over a channel, which could be physical (like a cable) or a wireless medium. An ideal channel exhibits constant gain and linear phase characteristics, meaning that the channel behavior would be the same across all frequencies of interest, also known as a "flat" channel response.

For any practical channel, the amplitude of the frequency response is not constant (flat), and the phase response is not linear with frequency.

Impacts of channel response on modulated signals

When a modulated signal passes through a non-ideal channel, the modulated signal undergoes frequency-dependent changes in both magnitude and phase due to the channel's non-flat response. These conditions lead to the occurrence of linear distortion.

The role of equalization

We use equalization to compensate for the linear distortion caused by the degrading effects of a transmission path. In communication systems, receiver equalization is a common practice for removing linear distortion by addressing the non-ideal effects of the channel. The equalization process entails creating an inverse response to the channel to make the channel appear ideal (flat). A simplified illustration appears in Figure 24.

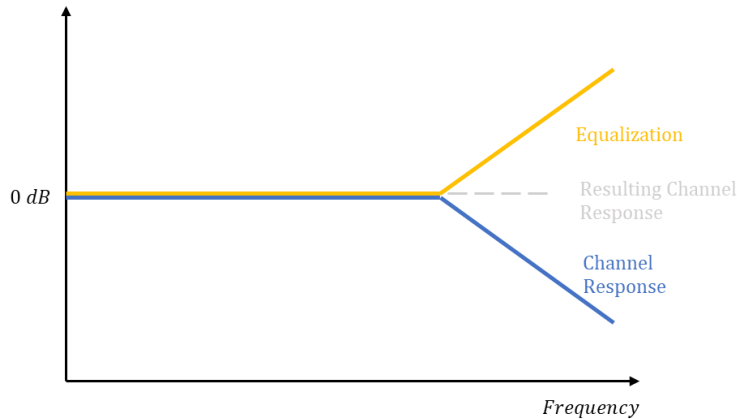


Figure 24. Channel equalization

For channel equalization, the challenge becomes finding the channel's frequency response.

The role of frequency response in the modulation distortion application

Similarly, the response of an amplifier, much like that of a practical channel, exhibits non-flat characteristics versus frequency. In the modulation distortion application, we apply concepts related to frequency response identification to isolate the linear from the nonlinear response of the amplifier.

The relationship between the output spectrum $Y(f)$, input spectrum $X(f)$, frequency response function $H(f)_{actual}$ and distortion $D(f)$ is given by

$$Y(f) = H(f)_{actual} X(f) + D(f)$$

We use $H(f)_{actual}$ to indicate the true frequency response function of the DUT, whereas $H(f)_{measured}$ indicates the frequency response function obtained through measurements in the modulation distortion application. The goal of the modulation distortion application is to minimize the difference between $H(f)_{measured}$ and $H(f)_{actual}$.

The modulation distortion application uses the modulated input signal as the stimulus signal to find the frequency response function. This approach differs from the conventional practice of sweeping a constant amplitude CW stimulus signal over a large bandwidth to measure this response. We measure the input spectrum $X(f)$ and output spectrum $Y(f)$ and $H(f)_{measured}$ is determined by linear regression. $H(f)_{measured}$ is reported as MGain21.

Use the following formula to find the distortion:

$$D(f) = Y(f) - H(f)_{measured} X(f)$$

The role of equalization aperture

In the MOD application, the equalization aperture is a setting that determines the ability of the linear regression algorithm to fit ripples in the DUT frequency response function, $H(f)_{actual}$. The equalization aperture represents the frequency resolution for which $H(f)_{measured}$ is fit to $H(f)_{actual}$. When there is a ripple at the output, opting for a narrow equalization aperture enhances the level of detail captured within the ripple and results in a more accurate fit, as illustrated in Figure 25.

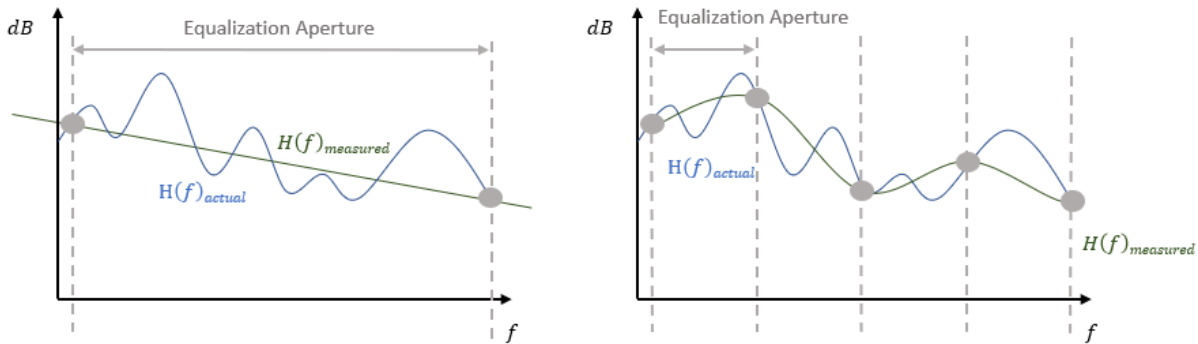


Figure 25. The width of the equalization aperture determines how well the linear regression algorithm fits the DUT frequency response function

In the illustrations in Figure 25, the dotted lines indicate the width of the equalization aperture.

How equalization aperture affects the distortion measurement

A narrow equalization aperture leads to insufficient sampling, resulting in a significant level of uncertainty in $H(f)_{measured}$. Conversely, while providing less detail of the output ripple, a wider aperture makes the measurement less susceptible to noise. It involves a tradeoff between fitting ripples with high uncertainty, resulting in the $H(f)_{measured}$ trace appearing "noisy" or not fitting ripples but having low uncertainty in the measurement.

Narrow Aperture

Provides more detail of DUT output ripple

Measurement uncertainty is higher

Wide Aperture

Provides less detail of DUT output ripple

Measurement uncertainty is lower

Recommendations

The recommendation is to set the equalization aperture at 5% of the BW, with a minimum of 150 tones. It's typical to use 3,000 tones. You can calculate the appropriate equalization aperture in the frequency domain using the following formula:

$$\text{Equalization aperture} = \frac{\text{signal bandwidth}}{\# \text{ of ripples}}$$

In the time domain, we calculate the appropriate equalization aperture from:

$$\text{Equalization aperture} = \frac{\text{symbol rate}}{\# \text{ of filter taps}}$$

The aperture settings denote a "tap equivalent," and they prove useful when comparing EVM results between the modulation distortion application and a VSA. For receiver equalization, the taps represent the values of the discrete filter applied to the received signal for correcting a non-flat channel response. The number of taps corresponds to the number of ripples within the signal bandwidth and, as such, corresponds with the OFDM.

When we compare the EVM results between the modulation distortion application and a VSA, we expect the OFDM results of both methods to match closely. This expectation arises because the OFDM equalization in a VSA relies on measuring pilot tones and performing the equalization in the frequency domain, resulting in a near-perfect equalization, consistent with the calculations performed by the modulation distortion application.

Single Carrier

When users compare EVM results between the modulation distortion application and a VSA for single carrier signals, they should configure the equalization aperture in the modulation distortion application to the equivalent equalization filter in the signal analyzer.

Users can perform an S21 measurement of the DUT in the standard channel to determine the appropriate equalization aperture. Figures 26 and 27 illustrate how the S21 measurement can guide the selection of the equalization aperture.

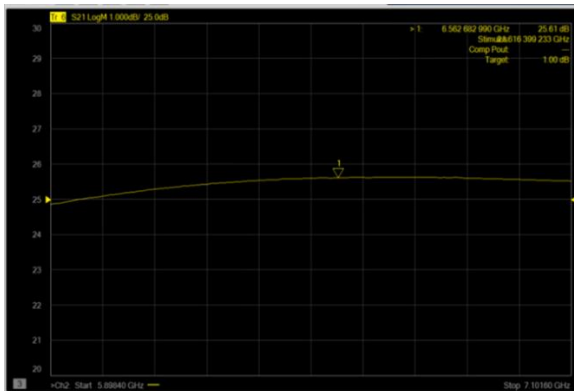


Figure 26. For an amplifier with a relatively flat frequency response function, the equalization aperture can be set to a wider value.

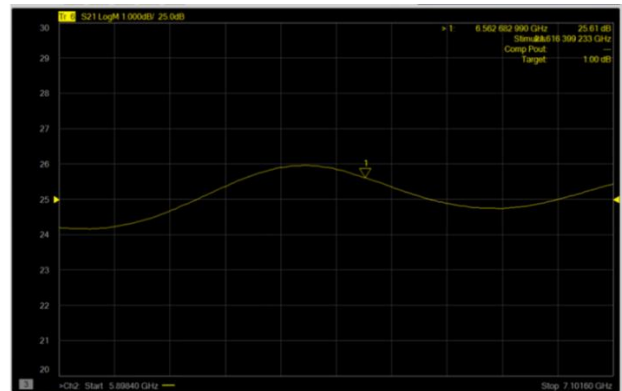


Figure 27. A narrower equalization aperture can be used to fit the ripples.

Limits: Equalization aperture = too wide

If the equalization aperture is very wide and there is ripple at the output, you will find that the frequency response function $MGain_{21}$ no longer fits $H(f)_{actual}$. The tone-by-tone modulation gain, or $PGain_{21}$ (P_{out2}/P_{in1}), appears in Figure 28 below. The equalization aperture setting for this measurement would underestimate the linear contributions to $Y(f)$, resulting in a non-ideal equalization. Consequently, part of the linear dispersion is included in the reported distortion EVM value, leading to a slightly higher distortion EVM value.

Figure 28 shows that $H(f)_{measured}$ is missing the ripple in $H(f)_{actual}$.

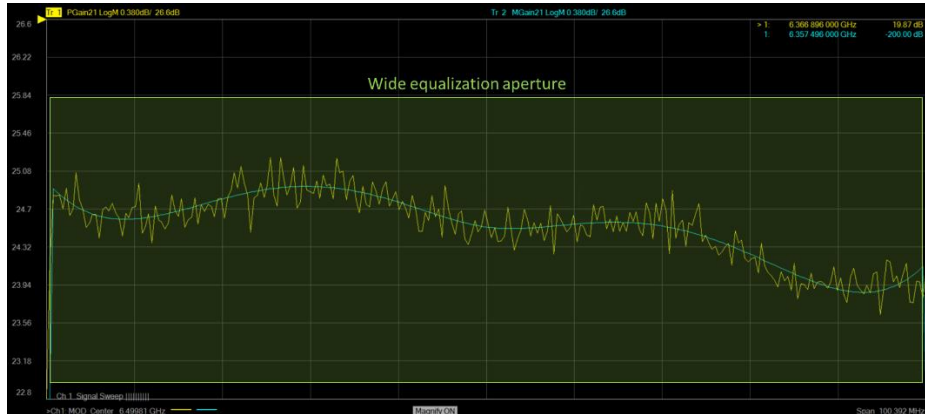


Figure 28. The equalization aperture width is too wide and is missing ripples in the DUT output

Limits: Equalization aperture = too narrow

In Figure 29, the measurement screen capture shows the equalization aperture was set too narrowly. As a result, the $MGain_{21}$ trace appears the same as the $PGain_{21}$ trace, creating a "noisy" appearance. The noise arises from the reduced number of tones used to calculate the gain. The measurement becomes highly uncertain when the regression interval (equalization aperture) lacks adequate tones for a meaningful gain measurement.

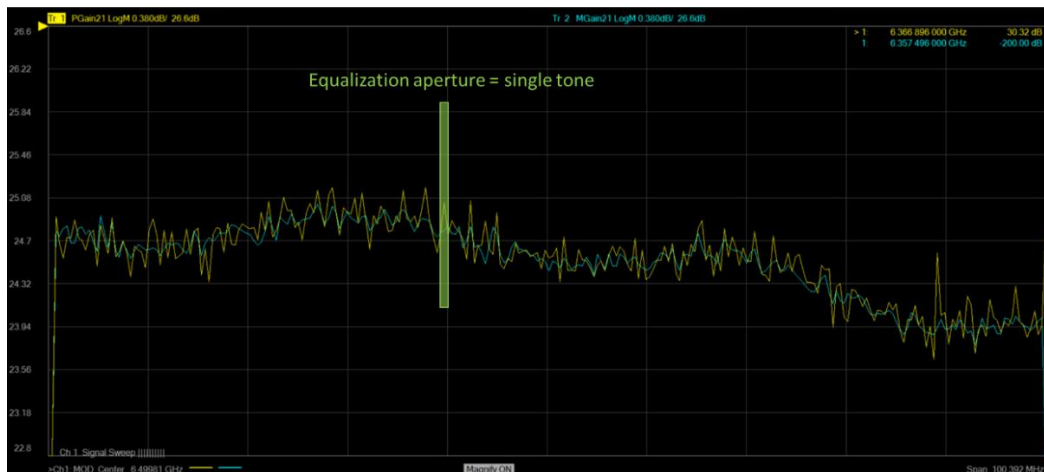


Figure 29. The equalization aperture width is too narrow, resulting in high measurement uncertainty

Equalization Aperture = Reasonable

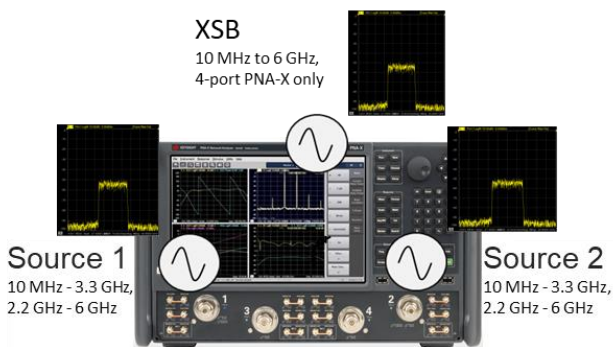
When you set the equalization aperture appropriately, as shown in Figure 30, MGain21 follows the ripples in $H(f)_{actual}$.



Figure 30. An appropriately set equalization aperture shows the MGain trace following ripples in the DUT output

PNA-X Internal AWG

The PNA-X can internally generate modulated signals, including multitone and complex digital modulation signals, to simplify the setup for active device characterization and perform EVM, NPR, and ACPR measurements.



The Internal 6 GHz arbitrary waveform generation (AWG) application uses the PNA / PNA-X's direct digital synthesizer (DDS) sources and works with MOD or the spectrum analysis application. Waveforms are limited to 6.2866 us (131k samples at 19.2 GSa/s) and must be on a frequency grid of $(M/N) * 600$ MHz with a minimum tone spacing range of 150 kHz. This feature requires an additional SW license, S93072B. It's worth noting that certain limitations in signal generation bandwidth apply to ports 1 and 3 of the PNA-X due to the use of analog filters to provide the best-in-class source harmonic performance. These limitations do not affect ports 2 and 4 or the XSB source.

DPD

Modulated signals with a non-constant envelope and a high peak-to-average power ratio require linear PAs with a wide dynamic range. You can choose between two methods to avoid distortion caused by high-power peaks in modulated signals. The first method involves operating the PA with high input back off, ensuring it doesn't compress when handling high-power signals. However, this method results in the PA operating at wasted capacity for a significant time. The second method focuses on efficiently using the PA and addressing the distortion separately. Balancing linearity and efficiency is an inherent tradeoff. Operating the PA with high input back off while reducing distortion also decreases efficiency, decreasing signal transmission range and shorter battery life. Consequently, it is not the preferred method for distortion reduction in a PA.

DPD is a method that linearizes the PA by applying predistortion to the input signal with the inverse of the distortion contributed by the PA. This process results in a less distorted signal at the output of the PA. The MOD application now includes a new DPD analysis capability that enables users to generate both direct DPD and model DPD. PA designers can use the DPD analysis feature in the MOD to determine the minimum achievable EVM and DPD cost by comparing the characteristics of various DPD models. The DPD feature is only available on the PNA and PNA-X platforms.

Load-pull

Typically, designers create PAs for a fixed output load, such as 50 Ω. Variations in loading conditions can significantly impact PA efficiency, linearity, output power capability, and reliability. The process of load-pull involves varying (pulling) the load impedance presented to the DUT away from the nominal 50 Ω so changes in performance, like distortion parameters, can be measured under non-50Ω load impedance conditions.

The modulation distortion application includes integrated load control for distortion measurements or arbitrary load control (ALC) for MOD. The powerful modulation distortion feature can compute FOMs such as EVM and ACPR while using a modulated input signal under varying load conditions; see Figure 31.

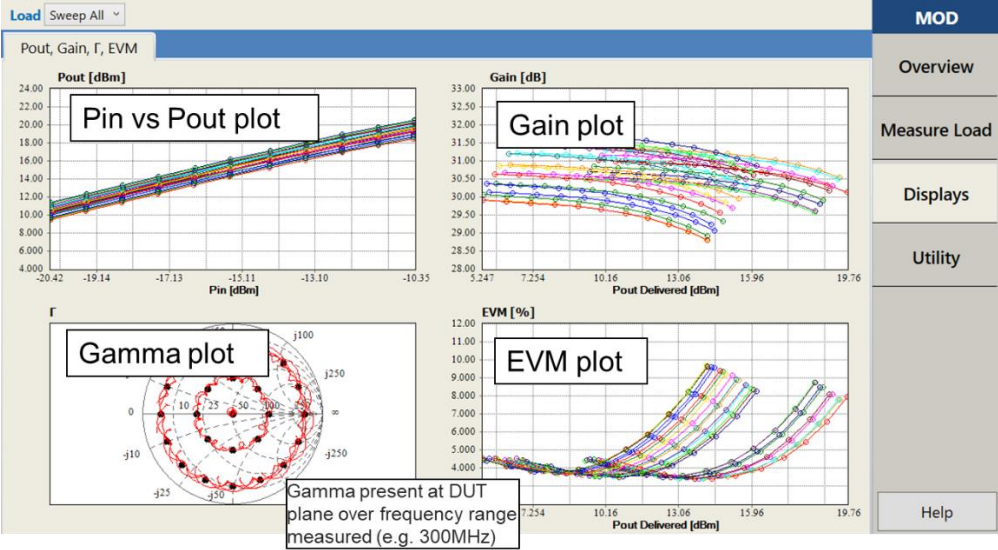


Figure 31. Arbitrary load control for MOD distortion measures power, gain, and EVM for varying gamma

Additional Challenges in Device Characterization Under Wideband Modulated Input Signals

The load impedance presenting to the PA varies in mobile communication devices because mobile devices are rarely stationary. Depending on user habits and usage locations, load conditions may vary greatly. In 5G massive multiple-input, multiple output (MIMO) base stations (BSs) with beamforming capabilities, the input impedance of different antenna elements changes due to their mutual coupling. Since the impedance of each antenna element deviates from $50\ \Omega$, the power amplifier in the front end of the beamformer IC may experience performance degradation. As phased arrays become ubiquitous, the use and significance of load sensitivity tests for power amplifiers are rising.

Performing Load-Pull Measurements

Figure 32 demonstrates the use of ALC for MOD to perform a load pull measurement on a PA under modulated input signal conditions. The illustration displays the setup for a focus microwave tuner (left) and a Maury microwave tuner (right).

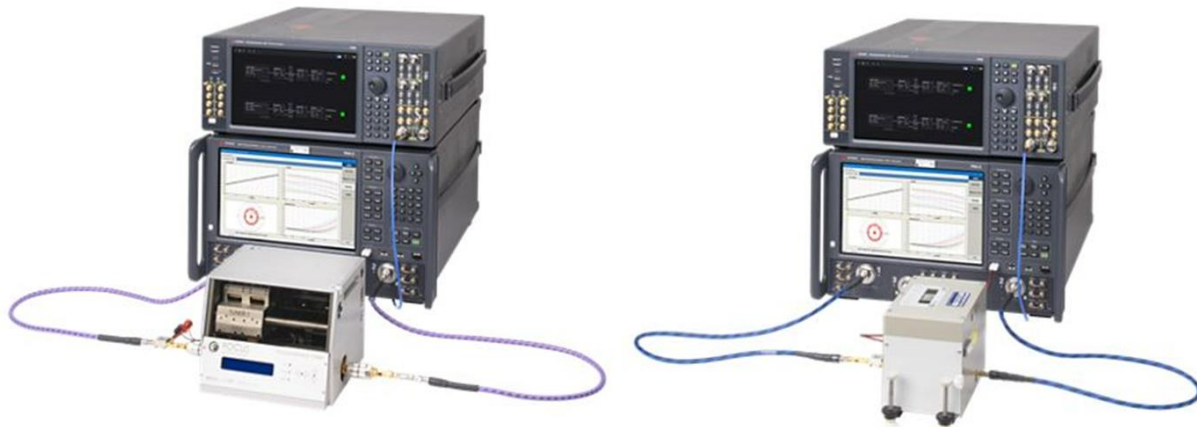


Figure 32. Arbitrary load control for a MOD distortion measurement setup

The VXG generates the modulated input signal while the tuner connects to the DUT as a load. The PNA controls the X-Y positions of the tuner, varying the load impedance presented to the DUT at a specific frequency point. After configuring the impedance at a specific frequency point, the PNA measures the gamma for the bandwidth of interest. Subsequently, you can make EVM, NPR, and ACPR measurements as a function of that load condition. ALC for MOD works with Focus microwaves and Maury microwave tuners.

Passive Tuners

Passive electromechanical tuners typically have two parallel plates, a precision $50\ \Omega$ slab line center conductor, and a metallic probe that can move in the X and Y directions. A motor precisely controls the position of the probe in an electromechanical tuner. When the probe moves in the Y direction, it disrupts the signal's electric field entering the tuner, resulting in a change in the magnitude of reflection (γ).

Conversely, moving the probe in the X direction (toward or away from the DUT) changes the phase of the reflection. This way, the tuner can present nearly any impedance to the DUT. The range of impedance values that the tuner can present to the DUT is limited by the tuner type and any loss in the test set. Passive tuners reflect the signal to the DUT; the reflected wave is always less than the incident wave.

Calibrating the Tuner

Tuner calibration involves mapping the X and Y positions of the probe in the tuner to s-parameters. During load-pull measurements, the PNA-X uses the mapping to present user-specified impedances to the DUT. When calibrating for a carrier frequency of interest, the user goes through the process using a standard channel on the PNA. Fundamental tuners, such as those used in the ALC for MOD application, set one specific frequency gamma but do not exert control over harmonics.

Setting the load grid

The load grid represents a set of impedance values the user defines and will present to the DUT during the MOD measurement. The ALC for MOD software simplifies the generation of the load grid by either creating a grid from the magnitude and phase values or allowing the user to input custom values, as shown in Figure 33.

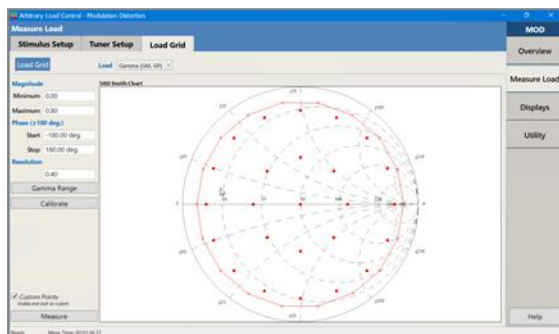


Figure 33. User-defined gamma load values

Calibrating the load grid

To ensure accurate load-pull measurements, it is essential to precisely measure the gamma at the DUT reference plane. Since the MOD channel is calibrated with the tuner initially set to the 50 Ω condition, the PNA calculates correction terms for each gamma value in the load grid and applies the correction terms to the MOD channel. During the load grid calibration, s2p data at each gamma point are measured and saved, then are used during the measurement. By doing so, accurate gamma values present at the DUT reference plane. The load grid calibration maintains the validity of the PNA calibration for the 50 Ω gamma value and each user-defined load impedance.

The ALC for MOD software indicates verified gamma values within a specified tolerance by highlighting them on the load grid, as shown in Figure 34.

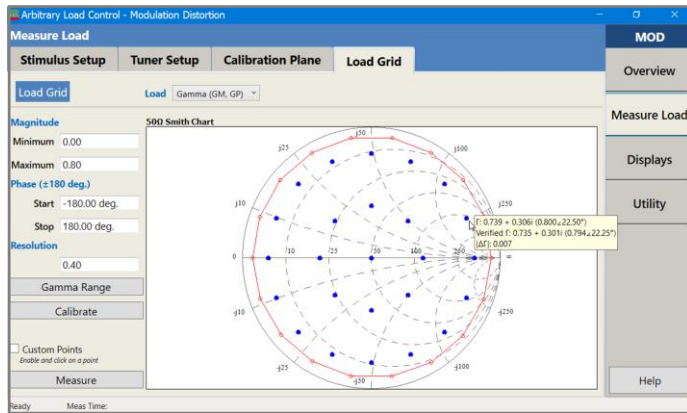


Figure 34. Load grid calibration for precision measurements

Optimizing the load-pull measurement

Achieving precise presentation of the specified load impedance values at the DUT reference plane requires meticulous care. During the calibration of the MOD channel, tuner, and load grid, the user needs to be aware of the positions of the reference planes for both the DUT and the tuner. After tuner calibration, the user must also account for any additions to the test set. The ALC software guides, including any front block components positioned between the DUT reference plane and the tuner calibration plane, ensure the correct gamma is presented to the DUT. It's important to note that only the calibration plane on the load side is of interest in load-pull measurements.

The ALC software shows the possible locations of the DUT, tuner, and PNA calibration planes. Users are guided through the necessary steps to align the planes as applicable to their specific test set configuration.

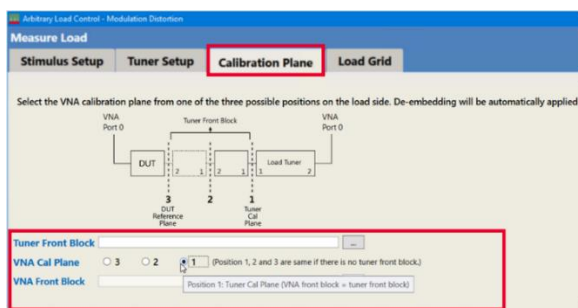


Figure 35. A calibration plane tab in the ALC guides the user to correctly de-embed the setup

Another consideration for load-pull measurements involves minimizing any loss within the test set to achieve the maximum gamma range. Figure 36, in the Smith chart, illustrates how small amounts of loss can limit the maximum gamma for the DUT.

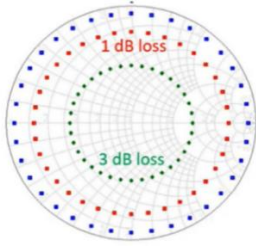


Figure 36. Test set loss limiting maximum gamma range

VSA-link

The PathWave Vector Signal Analysis (VSA) software provides a comprehensive toolkit for demodulation and vector signal analysis. With VSA-link, you can transmit the IQ data from the PNA-X to the VSA, where the IQ data demodulates, and the EVM is calculated in the time domain.

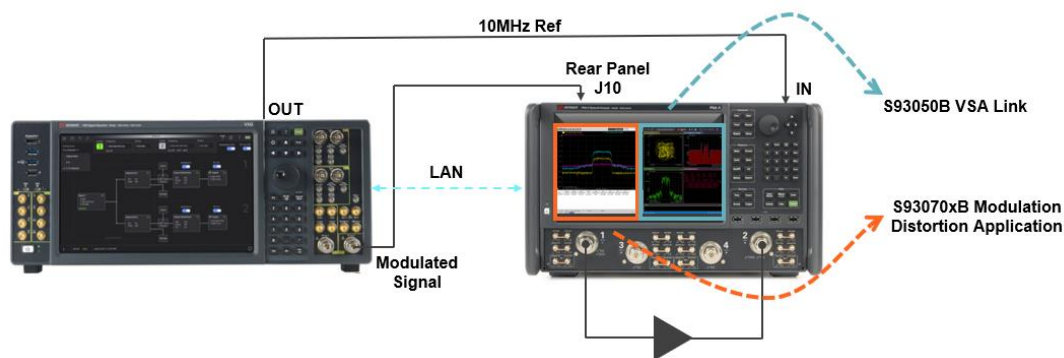


Figure 37. Modulation distortion measurement with VSA link for signal demodulation

Once the PNA-X acquires and processes the data through inverse fast Fourier transform (IFFT), the resulting IQ data is forwarded to the VSA for conducting AM-AM and AM-PM measurements, supporting ultra-wideband analysis alongside complete EVM demodulation.

PA designers who require EVM under a specific modulation scheme such as 5G NR, Wi-Fi, and DBSX find VSA-link particularly useful. Beyond examining memory effects and gain compression, VSA-link enables you to correlate EVM measurements in the time domain with those in the frequency domain.

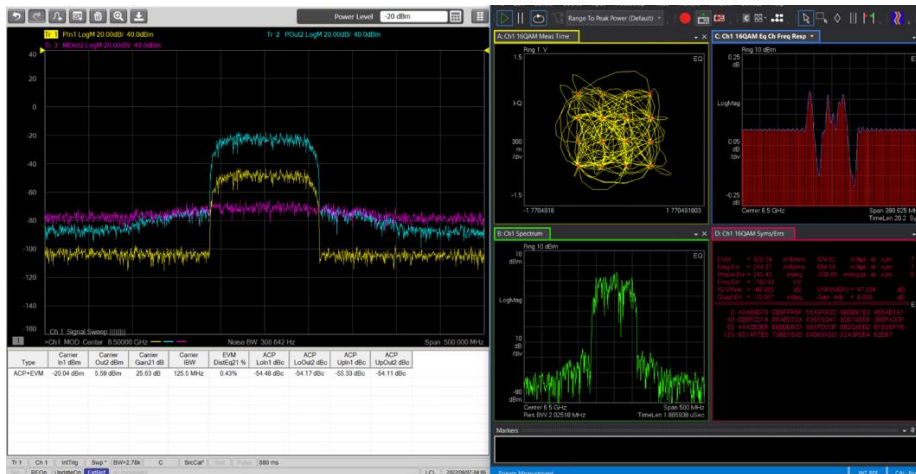


Figure 38. MOD distortion measurement EVM results show MOD distortion measurement IQ data streamed to VSA for demodulation (left) and EVM calculation (right)

Appendix A: Technique comparison for measuring EVM for amplifiers

Measuring the EVM of components has historically followed the same method as measuring the EVM at the system level or characterizing transmitter performance. When using Keysight 89600 Vector Signal Analyzer (VSA) software, the input signal undergoes measurement in the time domain. The captured waveform is then demodulated using known modulation format parameters and compared to a calculated reference waveform. The EVM is the delta between the reference and demodulated waveform at the decision point or ideal constellation.

In system level and transmitter testing, EVM measurements hold significant importance. The reason is that the aggregate EVM contains contributions from various sources, including nonlinear distortion, added noise, carrier phase noise, potential assimilation onto subcarriers, and IQ imbalances such as quadrature, gain differences, and DC offsets. All these errors contribute to the delta between the measured signal and the calculated reference. Previous technical papers and application notes have discussed how to interpret the results, aiding designers in understanding the individual distortion contributions.

Now consider the EVM contribution of a component such as a PA when it is used in a system or transmitter. The EVM of a component is primarily influenced by contributions from distortion or noise, depending on the amplifier's operating point. IQ balance errors do not originate in the amplifier. Another system error is phase noise, with the main contribution from the system oscillator and not the amplifier. The amplifiers generate residual phase noise, but you can disregard the contribution from residual phase noise compared to the noise or distortion contribution.

You can summarize the error contribution of EVM for a component by using a typical "bathtub" curve for an active component (see Figure 39). The "bathtub" curve represents the measured EVM versus power and shows the different error contributions since they are power-dependent. For low power levels, the signal-to-noise or noise figure dominates the EVM; at high power, nonlinear distortion dominates EVM. Note that most system designs will operate somewhere on the "distortion side" to optimize power-added efficiency (PAE) while maintaining overall system-level performance targets.

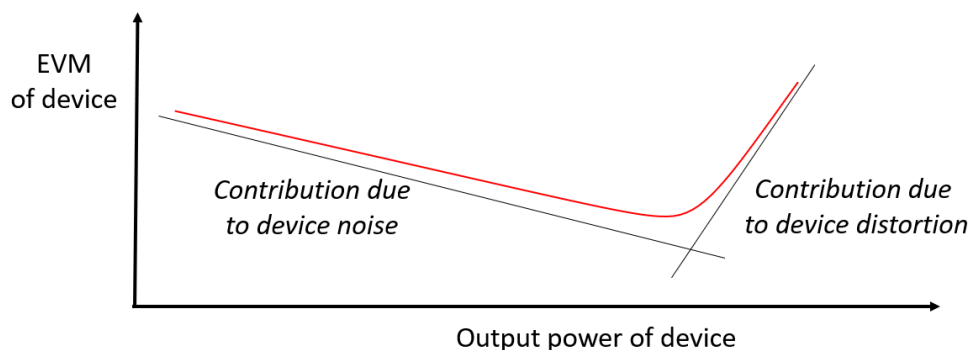


Figure 39. EVM bathtub curve

An additional error not discussed for component measurements is the contribution of the input signal nonidealities. When measuring components using the traditional VSA method, there is an inherent

assumption that the input signal is perfect. However, the input signal has nonidealities or errors from the ideal signal that will set the minimum EVM measurable by the receiver.

For modulation distortion on the VNA, the distortion EVM measurement focuses on the critical contributions of the device. It is immune to input signal EVM since it is measured simultaneously with the output signal. The measured output signal separates into a linear portion, yielding gain and the nonlinear distortion contribution EVM. The measurements are conducted with high precision and high signal-to-noise measurement of the VNA receiver. The residual EVM noise floor is improved versus the traditional VSA approach for devices by eliminating the input signal's contribution and leveraging the VNA receiver's good sensitivity.

The benefits of each approach to measuring EVM appear in Figure 40.

	89600 VSA and X-Apps	S9x070xB Modulation Distortion
Supported instruments	Keysight signal analyzers, oscilloscopes, PXI VSAs	Keysight PNA-X, ENA-X, PXI VNA
Benefits	<ul style="list-style-type: none"> • Standard compliant algorithms • Flexible views of EVM (time, frequency, subcarrier) • Constellation diagrams • Measures all system and transmitter EVM contributors 	<ul style="list-style-type: none"> • Wide BW measurements are limited only by signal generator bandwidth • Lower residual EVM enabled by good sensitivity and eliminating contributions of the input signal • Measures contribution of the device only

Figure 40. Comparison of Keysight's EVM measurement methods

Appendix B: Spectral correlation theory

The Modulation Distortion is a frequency domain analysis method. Nevertheless, Appendix B will show that the total amount of distortion measured in the frequency domain is equal to that in the time domain as measured using the existing EVM method.

To explain the theory, one starts by looking at the nature of the real and imaginary parts of the spectrum $M(f)$ corresponding to a theoretical infinitely long modulated input waveform. If the modulation corresponds to a stationary stochastic process, the real and imaginary parts of $M(f)$ are uncorrelated, and normally distributed stochastic variables with zero mean. The variance of the real and imaginary parts combined is a function of frequency and is more commonly known as the "power spectral density" (PSD). The modulation distortion method described in the app note originates from statistical correlations of such stochastic spectra. The spectral correlation of a signal $U(f)$ and a signal $V(f)$ is denoted by $S_{UV}(f)$ and is defined:

$$S_{UV}(f) = E[U^*(f)V(f)],$$

Whereby $E[.]$ stands for the statistical "expectation" or "mean" operator, and the superscript "*" denotes the complex conjugate operator. $S_{UV}(f)$ is also called the cross-spectral density of $U(f)$ and $V(f)$. The PSD of any signal $M(f)$ is equal to $S_{MM}(f)$, the cross-correlation of the stochastic spectrum with itself (this is an autocorrelation). Under linear small signal operating conditions, the relationship between the spectrum of the input signal $X(f)$ and the spectrum of the output signal $Y(f)$ is given by

$$Y(f) = g_{SM}(f)X(f),$$

Whereby $g_{SM}(f)$ represents the complex transfer function corresponding to the small signal gain of the amplifier. When the input power increases, the amplifier will start to behave in a nonlinear way. This nonlinear behavior manifests itself in 2 ways: the ratio between the input spectrum and output spectrum changes (this is gain compression), and noise-like distortion signals occur both in-band (EVM) and out-of-band (ACPR). This is mathematically expressed as follows:

$$Y(f) = c_G(f)g_{SM}(f)X(f) + D(f),$$

With $c_G(f)$ representing the gain compression function and with $D(f)$ representing the nonlinear distortion that does not linearly correlate with $X(f)$. The linear uncorrelatedness appears mathematically as:

$$S_{XD}(f) = 0.$$

Note that $c_G(f)$ is a complex-valued continuous frequency response function, equal to one under small signal linear operating conditions, and that $D(f)$ is a stochastic variable similar in nature to $X(f)$ and $Y(f)$. $D(f)$ can be interpreted as the superposition of all cross-frequency intermodulation products that end up near the carrier frequency.

We assume in the following that, for a given amplifier, $c_G(f)$ and $S_{XD}(f)$ are uniquely and solely determined by two things: The power spectral density of the input signal $S_{XX}(f)$, on the one hand, and the statistical distribution of the amplitude of the corresponding time domain complex envelope $x(t)$ on the other hand. We express the statistical amplitude distribution by the complementary cumulative distribution function (CCDF) of the instantaneous power levels. The CCDF is a function of power and represents the percentage of time the instantaneous output power is higher than any given power level.

The modulation distortion measurement problem is defined as follows. For a given input signal, $X(f)$, corresponding to a given CCDF and PSD, decompose the output signal, $Y(f)$, into a function $c_G(f)g_{SM}(f)X(f)$, which is linearly correlated with $X(f)$, and the remaining part $D(f)$, which contains the nonlinear distortion. We explain the measurement procedure below:

Start by measuring $g_{SM}(f)$ using a VNA. Next, excite the DUT with a signal $X(f)$ that has a CCDF and PSD corresponding to the modulation format and power level you want to know $c_G(f)$ and $D(f)$. Next, use the VNA to measure the complex vector ratio $Y(f)/X(f)$, denoted by $R(f)$, as well as $|X(f)|$ and $|Y(f)|$. Note that these measurements require both a power calibration and a vector calibration. The spectral correlation quantities $S_{XY}(f)$, $S_{XX}(f)$ and $S_{YY}(f)$, which will be used as intermediate variables, are calculated as follows:

$$S_{XY}(f) = E[|X(f)||Y(f)|e^{j\varphi(R(f))}] = E[X^*(f)Y(f)],$$

$$S_{XX}(f) = E[|X(f)|^2] = E[X^*(f)X(f)], \text{ and}$$

$$S_{YY}(f) = E[|Y(f)|^2] = E[Y^*(f)Y(f)].$$

With a dense enough tone spacing, the expectation operator $E[\cdot]$ is practically performed by calculating the mean for a significant number, e.g., 100, of adjacent tones. Next one calculates $c_G(f)$ using

$$c_G(f) = \frac{S_{XY}(f)}{g_{SM}(f)S_{XX}(f)},$$

and substitution of (8) in to (3) and solving for the nonlinear distortion component $D(f)$ results in

$$D(f) = Y(f) - \frac{S_{XY}(f)}{S_{XX}(f)}X(f).$$

The PSD of the nonlinear distortion, $S_{DD}(f)$, is given by

$$S_{DD}(f) = (1 - \gamma_{XY}^2(f)) S_{YY}(f), \text{ with}$$

$$\gamma_{XY}^2(f) = \frac{|S_{XY}(f)|^2}{S_{XX}(f)S_{YY}(f)}.$$

In the quantity $\gamma_{XY}^2(f)$ is referred to as the "linear coherence function." Equations (8), (10), and (11) are derived as follows. Frequency dependency is omitted in the following equations. Multiplying both sides of (3) by X^* and calculating the expectation results:

$$E[X^*Y] = E[X^*(c_G g_{SM} X + D)], \text{ expressed as}$$

$$S_{XY} = c_G g_{SM} S_{XX} + S_{XD}.$$

Substitution of (4) into (13) and solving for c_G results in (8).

Multiplying both sides of (3) by Y^* and calculating the expectation results in

$$E[Y^*Y] = E[(c_G g_{SM} X + D)^* (c_G g_{SM} X + D)],$$

which, after substitution of **Error! Reference source not found.**, can be written as

$$S_{YY} = |c_G|^2 |g_{SM}|^2 S_{XX} + S_{DD}.$$

Substitution of (8) into (15) results in

$$S_{YY} = \frac{|S_{XY}|^2 |g_{SM}|^2 S_{XX}}{|g_{SM}|^2 |S_{XX}|^2} + S_{DD},$$

which, when solved for S_{DD} results in

$$S_{DD} = \left(1 - \frac{|S_{XY}|^2}{S_{XX}S_{YY}}\right) S_{YY}, \text{ thereby proving (10) and (11).}$$

Appendix C: Compact Test Signals

As mentioned in Appendix B, the measurements of $c_G(f)$ and $S_{DD}(f)$ require the use of an input signal $X(f)$ that matches the CCDF and the PSD of a given modulation format — 5G NR. Because the VNA can only detect discrete tones, the input signal must have a discrete spectrum, or in other words, the input signal needs to be periodic. The following section explains the creation of a periodic input signal.

We start by generating a time-domain "parent IQ waveform" that is assumed to represent the modulation format one wants to investigate. The waveform could, for example, be an IQ waveform corresponding to a 10ms long radio frame of a 5G NR signal with a modulation bandwidth of 800 MHz. One option would be to repeat the radio frame, resulting in a signal with a discrete spectrum and a tone spacing equal to the reciprocal of the period, in this case, 100Hz. The total number of tones within the signal bandwidth would equal 8,000,000. As one typically measures across a span equal to three times the input signal's bandwidth to characterize up to third-order spectral regrowth, the total number of tones to be measured by the VNA would be about 24,000,000. Although not impossible, measuring that many tones using a VNA is inefficient and takes a long time. Our option is to use a significantly increased tone spacing, e.g., 100kHz. Such a signal only requires 8,000 tones to cover the signal bandwidth, and the VNA only needs to measure 24,000 tones. The reduced number of tones, 8,000, is high enough to guarantee a faithful representation of the PSD and CCDF of the parent waveform. The process of reducing the number of tones has been automated and is explained in the following passage.

To reduce the number of tones while maintaining a faithful representation of the parent waveform, one starts by calculating the CCDF of the parent waveform. Next, one selects the most statistically representative subset of the parent waveform, whereby the subset has a duration equal to the reciprocal of the desired 100kHz tone spacing; in our case, this corresponds to 10 μ s. The most statistically representative subset is the one that has a CCDF of its own that most closely matches the CCDF of the parent waveform. In the next step, the spectral leakage, caused by the fact that the beginning and ending samples of the subset are not aligned, is cleaned up by applying a brick-wall filter corresponding to the bandwidth of the parent waveform. The resulting signal, called a compact test signal (CTS), matches the CCDF and PSD of the parent waveform and is used as the input signal.

Appendix D: The relationship to existing EVM and ACPR

This Appendix explains the relationship between the existing methods for measuring EVM and ACPR and the new method.

With EVM, one measures the output signal $y(t)$ using a VSA. For modulation formats based on OFDM and standard for 4G, 5G, and Wi-Fi, EVM is the normalized mean squared error (NMSE) between the bandpass filtered $y(t)$ and the ideal input signal $x(t)$. Where $y(t)$ is compensated for complex gain, group delay, and frequency dispersion, and $x(t)$ is derived through demodulation of $y(t)$. Note that EVM is typically expressed in % or in dB. The following equation represents EVM:

$$EVM = \sqrt{\frac{\sum_{i=1}^N |x(t_i) - (y * e)(t_i)|^2}{\sum_{i=1}^N |x(t_i)|^2}},$$

Whereby N represents the total number of time samples acquired by the VSA, "*" stands for convolution and $e(t)$ stands for the equalization filter, which typically includes compensation for group delay, phase rotation, and frequency dispersion. The time instances t_i are determined by the demodulation process. The numerator and denominator in the righthand side of (18) are sums of amplitudes squared. They are, as such, equal to their frequency domain equivalents in accordance with Parseval's theorem and expressed below:

$$EVM = \sqrt{\frac{\sum_{i=1}^N |X(f_i) - E(f_i)Y(f_i)|^2}{\sum_{i=1}^N |X(f_i)|^2}},$$

Whereby $X(f_i), Y(f_i)$ and $E(f_i)$ represent the discrete Fourier transforms of $x(t_i), y(t_i)$ and $e(t_i)$.

Consider now the definition of a new quantity called equalized distortion-error-vector-magnitude ($DEVM_e$), which is based on our measurement of $c_G(f)$, $g_{SM}(f)$, $X(f)$ and $Y(f)$:

$$DEVM_e = \sqrt{\frac{\sum_{i=1}^N |X(f_i) - c_G^{-1}(f_i)g_{SM}^{-1}(f_i)Y(f_i)|^2}{\sum_{i=1}^N |X(f_i)|^2}},$$

Whereby the frequencies f_i correspond to the in-band frequencies of the compact test signal. Considering (19) and (20), one concludes that the classic measure of EVM is identical to $DEVM_e$ under the condition that the equalization filter used for the classic EVM is the inverse of the gain compression times the small signal gain.

The modulation format associated with the EVM may not perform an equalization, in which case $E(f_i)$ represents a group delay τ and a complex gain G . In that case, (19) and (20) become:

$$EVM = DEVM_u = \sqrt{\frac{\sum_{i=1}^N |X(f_i) - G e^{-j2\pi f_i \tau} Y(f_i)|^2}{\sum_{i=1}^N |X(f_i)|^2}},$$

Whereby the subscript "u" in $DEVM_u$ stands for "un-equalized". $DEVM_u$ will always be higher than $DEVM_e$ as it contains the linear distortion contributions due to frequency dispersion and the nonlinear contributions, whereas $DEVM_e$ only includes the nonlinear contributions. It is important to note that our method enables us to accurately determine EVM without demodulation, which significantly simplifies and speeds up the data processing.

ACPR is the ratio between the amount of spectral regrowth in an adjacent channel and the power in the channel itself. ACPR can be calculated directly based on the measurement of $S_{YY}(f)$ as it is defined in the frequency domain.

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