APPLICATION NOTE

Power Amplifier EVM Measurement with PNA-X Series Network Analyzer

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Nonlinearity Evaluation of Power Amplifier (PA)

PA in the wireless communication system

In wireless communication systems, PAs are located at the last stage of the transmission chain, to supply required RF power to antennas. The PA plays an important role, determining the quality of the communication service, in terms of signal quality and battery life.

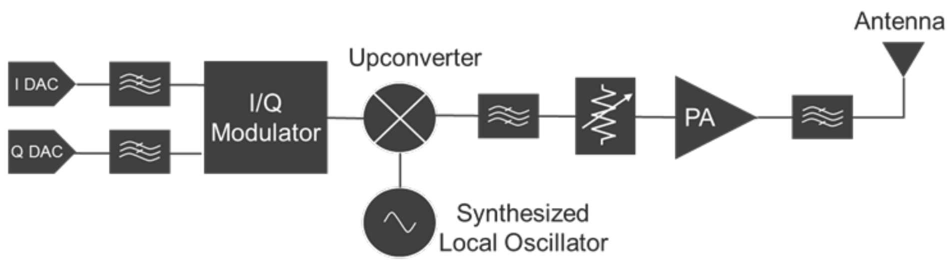


Figure 1. RF transmission chain

The PA is the biggest contributor to the quality of the RF chain in communication systems. Linearity is extremely important in systems with high peak-to-average-power (PAPR) signals, as in modern wireless standards where OFDM is used for the digital modulation scheme, because the nonlinear response of the RF chain directly impacts error in the demodulation, resulting in poor bit-error-rate. Also, spectral regrowth created by nonlinearity can create interference in other bands. Thus, maintaining linearity of the RF chain is critical for the quality of wireless communication.

The PA is also known as a contributor to the power consumption for wireless communication systems. Especially in the user equipment (UE) side, power consumption or power efficiency of the PA directly impacts the quality of service for wireless communication because a PA with poor efficiency consumes limited battery life quickly.

Because there is a trade-off between linearity and efficiency, one of the challenges for the PA industry is to design the PA to be linear as well as efficient, with demanding and emerging operating conditions such as mmW carrier frequency and extremely wide signal bandwidth. Digital predistortion (DPD) is a method of linearizing the PA by predistorting the input signal with the inverse of the distortion contributed by the PA, resulting in a linearized signal at the output of the PA.

To quantify PA nonlinearity under a modulated stimulus condition, EVM is commonly used as a Figure of Merit (FOM) for the in-band characteristics, and Adjacent Channel Power Ratio (ACPR) is used for the out-band characteristics. In this application note, a novel method of characterizing PA nonlinearity under a modulated stimulus condition is introduced.

Modulation distortion: Nonlinearity of the PA under modulated signal

Intermodulation (IM) is a commonly known parameter to quantify PA nonlinearity. In the IM measurement, the PA is excited by a two-tone stimulus, then intermodulation tones of both high side and low side tones are measured, that represents nonlinearity under two-tones stimulus condition. This method provides a rough idea of the nonlinear distortion of the PA, while the bias condition of the PA is for two-tones, which is different from actual operation.

Modulation distortion (MOD) is an extended concept from IM. Instead of two-tones, significantly "many" tones are used to stimulate the DUT. The modulated waveform signal, which has a certain bandwidth at the carrier frequency, is a good example of a stimulus for modulation distortion. Assume the PA has nonlinearity, spectral regrowth will be present at the output, which can be represented in the equation in the following figure.

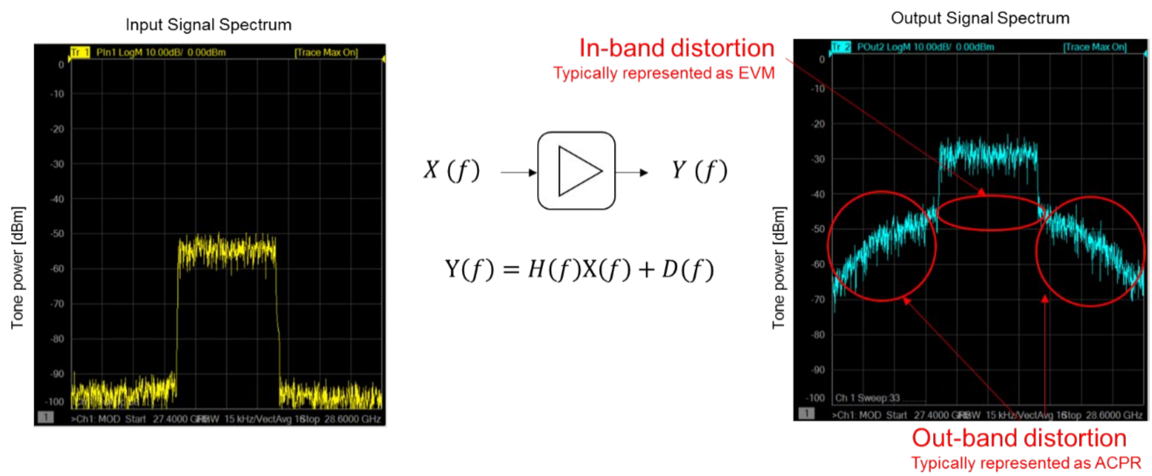


Figure 2. Nonlinear amplification of a modulated input signal

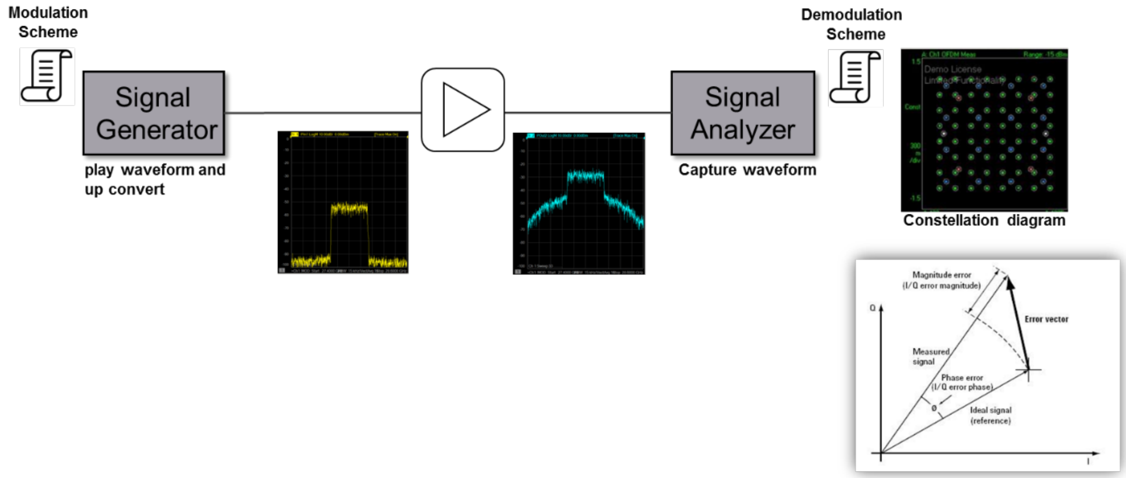
In this equation, H is a complex number that represents a linear transfer function. The D is a parameter called distortion, which is generated from all the nonlinear interactions of tones present in the PA.

ACPR is used to represent out-of-band distortion. The ACPR measurement is straight forward, which compares channel power between the signal and the adjacent channel.

The challenge then becomes quantifying in-band distortion because distortion tones are part of the measured in-band signal. In the next section, a traditional method to quantify EVM is explained.

Measuring EVM using VSA

EVM is one of the figures of merit to quantify the PA distortion that affects signal quality. A Vector Signal Generator (VSG) and Vector Signal Analyzer (VSA) are traditionally used to measure EVM.



$$\text{RMS EVM [\%]} = 100\sqrt{\frac{\frac{1}{N}\sum_{t=1}^{N} e[t]\cdot e^{*}[t]}{\frac{1}{N}\sum_{t=1}^{N} r[t]\cdot r^{*}[t]}}$$

Figure 3. EVM measurement using demodulation (or VSA) method

Demodulation method

- Signal modulation. Specific scheme (such as 64QAM) is used to create IQ waveform
- IQ waveform is played at the AWG, then it is upconverted to carrier RF frequency on the VSG
- Waveform stimulates the PA, then creates an output which may contain distortion
- On VSA, the output is captured by the signal analyzer, then quantified by a wideband digitizer
- Digitized quantity is demodulated then constellation diagram is plotted
- Error vector is evaluated from the measured constellation for each data point
- RMS EVM is computed by taking integration of the error for a certain waveform period

In this method, the distortion of the PA contributes to the EVM, along with many other factors such as noise and IQ imbalance. Also, it is assumed that the input signal is perfect and does not contribute to the measured EVM.

Note that the VSA method is considered a time domain measurement, as the output waveform is captured in the time domain by the wideband digitizer.

Challenges in device characterization under a wideband modulated signal

With emerging standards such as 5G, EVM needs to be measured at mmW carrier frequency and extremely wide signal bandwidth. In such a measurement scenario, it is more and more challenging to accurately measure EVM of the device for the following reasons:

Error contributions of stimulus

In the VSA method, the Error Vector is evaluated between the ideal and measured signal for each constellation. Thus, any imperfectness of the signal source directly impacts the measurement result. The distortion of the generated signal needs to be lower than the DUT.

Also, the VSA method is sensitive to IQ imbalance and phase noise, which is normally not created by the DUT.

Random noise can also impact the EVM measurement result, especially if the power level is low. As signal bandwidth (BW) widens, Signal-to-Noise Ratio of the signal from the signal generator gets narrower.

Error contributions of receiver

To analyze a signal with minimum error, the signal needs to be digitized without added nonlinear distortion from the signal analyzer. Also, the noise floor of the receiver needs to be lower than the target signal but as the signal BW widens, the signal-to-noise-ratio (SNR) of the receiver gets narrower.

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

Signal fidelity

In the VSA method, the test system is characterized beforehand, then the stimulus is compensated using a complex term so that the DUT is stimulated by the desired signal. Also, the measured result can be compensated using the complex term.

However, this method can include error due to widened bandwidth and the frequency response of the mismatch error between the DUT and test system because mismatch of the DUT is not removed from the measurement.

Modulation Distortion measurement on PNA-X

There is a new approach to distortion measurements that overcomes the challenges of the VSA method. The Keysight PNA-X with a VSG can now characterize nonlinear distortion of a device with a modulated input signal, then derive FOMs such as EVM and ACPR.

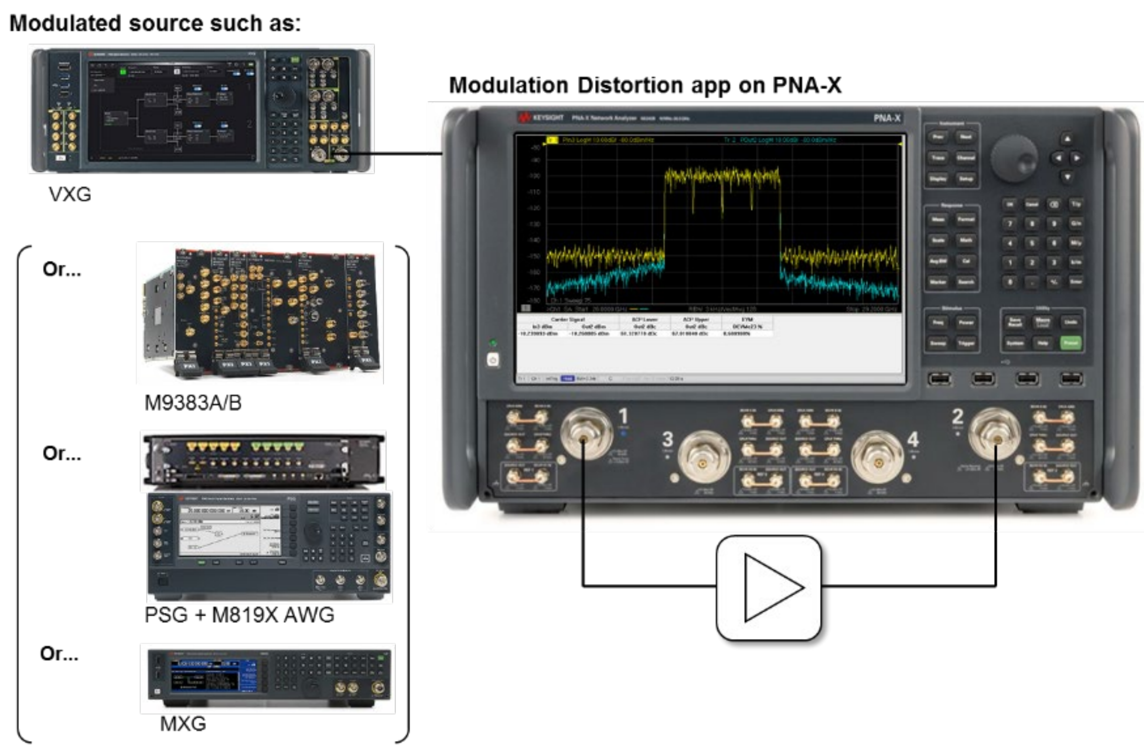


Figure 4. Distortion measurement using the Keysight PNA-X

This new approach is in the software application of the PNA-X, called “Modulation Distortion”, or “MOD”. The measurement setup is integrated into the PNA-X firmware, which allows the user to easily set up the stimulus and measurements and leveraging state-of-the-art calibration techniques for the best accuracy.

In addition to existing PNA-X measurements such as S-parameter, Gain Compression, IMD and Noise Figure, the PNA-X now allows nonlinear distortion measurement under the modulated stimulus condition, without changing connections to the DUT.

In this section, the underlying technology for the MOD will be briefly explained. A detailed discussion also will be explained in the Appendix.

Compacting the modulated waveform

In the MOD application, the measurement does not require a long waveform, such as a full frame or even a sub-frame. To perform an accurate enough measurement within a short measurement time, the MOD application uses a shortened waveform period. This is called “compacting” the waveform. For example, when a user would like to know the response of the DUT to a waveform under a specific modulation scheme, called a “Parent waveform”, the PNA-X firmware takes a slice of the parent waveform, and the compact waveform will inherit the frequency signature and statistical characteristics. The slice of the waveform is called the “Compact Test Signal” or CTS.

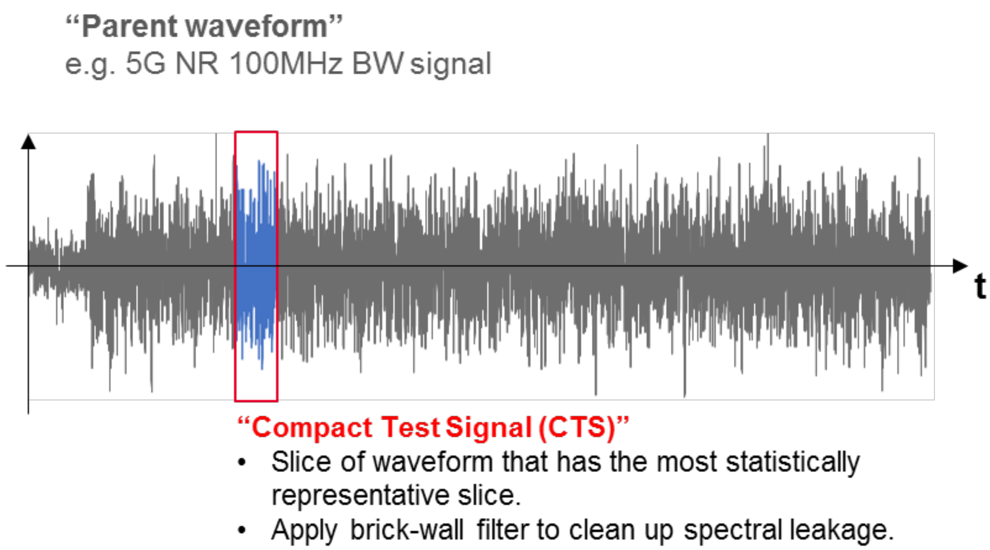


Figure 5. Compact test signal that is statistically representative slice of parent waveform

The PNA-X firmware uses a unique algorithm to find the most statistically representative slice from the parent waveform given parameters provided from the user, then processes the waveform by applying a brick-wall filter to remove spectral leakage when the CTS is played over time.

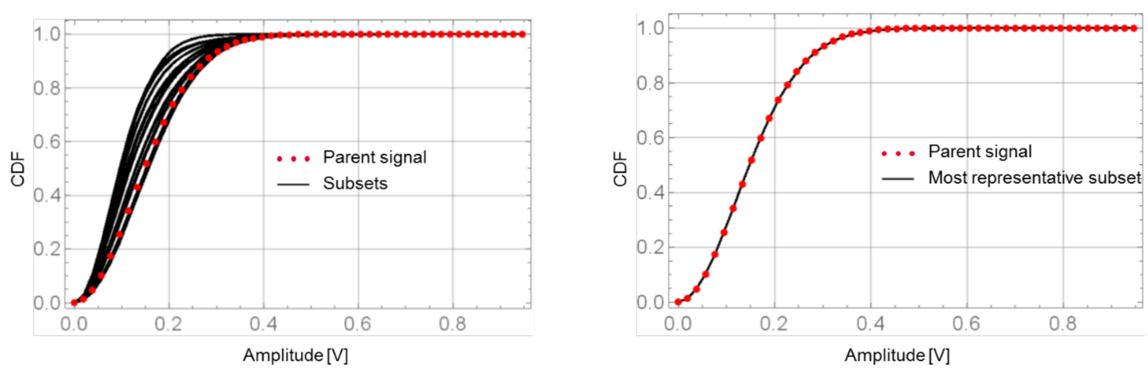


Figure 6. PNA-X algorithm to find most representative compact test signal from parent signal

The CTS has different characteristics depending on user provided parameters, which can affect the measurement result. As an example, let's use a 5G NR 100 MHz BW waveform as the parent waveform to create a CTS. The figure shows two different CTS characteristics given different parameters for the same parent waveform. The yellow plot indicates the parent waveform, and the blue plot indicates the CTS. The top figures show the spectrum of the waveform, namely how the waveform looks in the frequency domain. This figure can be obtained by an FFT (Fast Fourier Transform) of each waveform.

The figures in the middle show the position of the CTS in the parent waveform. Since the signals appear noise-like, modulated signals can be characterized using statistics. The cumulative distribution function (CDF) shows the probability or percentage of time that a sample of a signal is less than some value. This is important because signals at higher levels will clip or distort. The CCDF (complementary composite distribution function) shows the signal statistics in a way that allows the peak-to-average power to be quickly characterized. The left side of the graph is the reference and represents the average value of the signal. The further to the right that the CCDF curve extends, the greater the peak-to-average ratio and the greater need for overhead in the design of the amplifier to avoid distortion.

The bottom figures show the CCDF curve of the parent signal, CTS, as well as the Gaussian distribution in the pink trace. The horizontal axis is the signal's power relative to the average power, measured in dB. The percentage of time that the signal spends at or above each line gives the probability for that particular power level.

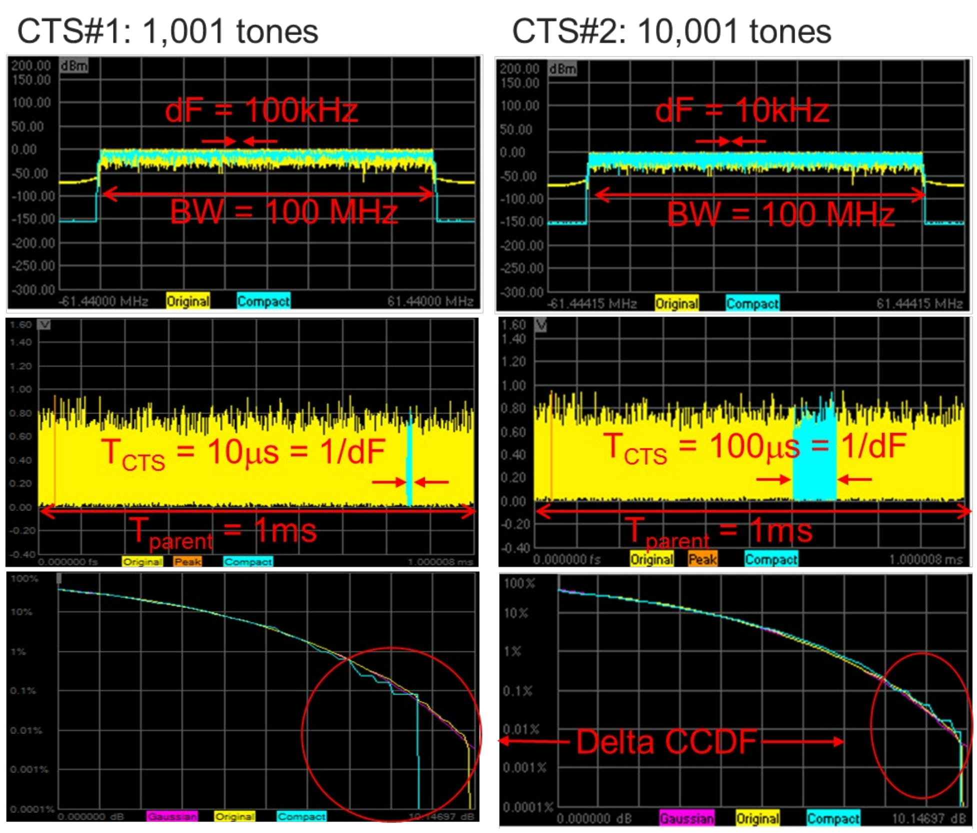


Figure 7. Comparing CTS characteristics generated with different user defined parameters

There are a couple of ways to discuss parameters of the CTS, but for now, the focus will be on the number of tones located in the in-band region of the modulated signal. The left CTS consists of 1001 tones, whereas the right CTS consists of 10,001 tones. In other words, the left CTS has 100 kHz tone spacing, while the right CTS has 10 kHz tone spacing.

The first thing to notice is that the frequency signature of both signals have the same bandwidth as the original signal. The original signal has higher out-of-band spectrum, whereas the CTS has low out-of-band spectrum, this is due to the brick-wall filter.

For the middle figures, which are in the time domain, the waveform length is the reciprocal of tone spacing. The left CTS has a 10 us waveform length, whereas the right CTS has a 100 us waveform length. A finer tone spacing waveform results in a longer period of the CTS waveform.

Finally, for the statistical characteristics of each waveform, the CCDF of the parent waveform is aligned with a Gaussian distribution. This is the nature of OFDM, which has noise-like characteristics. If the CCDF of the CTS is compared to the parent waveform, the right CTS is nicely aligned with the parent waveform, meaning that the power of the CTS waveform is very close to the average power of the parent waveform. On the other hand, the left CTS has alignment until around 0.1% probability, then a big discrepancy is observed, and the power of the CTS waveform is lower than the average power of the parent waveform.

In theory, the CCDF has good alignment until reciprocal of number of tones. For example, in order to have good alignment until 0.1%, a minimum of 1000 tones for the CTS are required.

Measuring spectrum using PNA-X receivers

Now the PNA-X measurement of the spectrum of the CTS will be explored. Figure 8 shows a simplified block diagram of the measurement system.

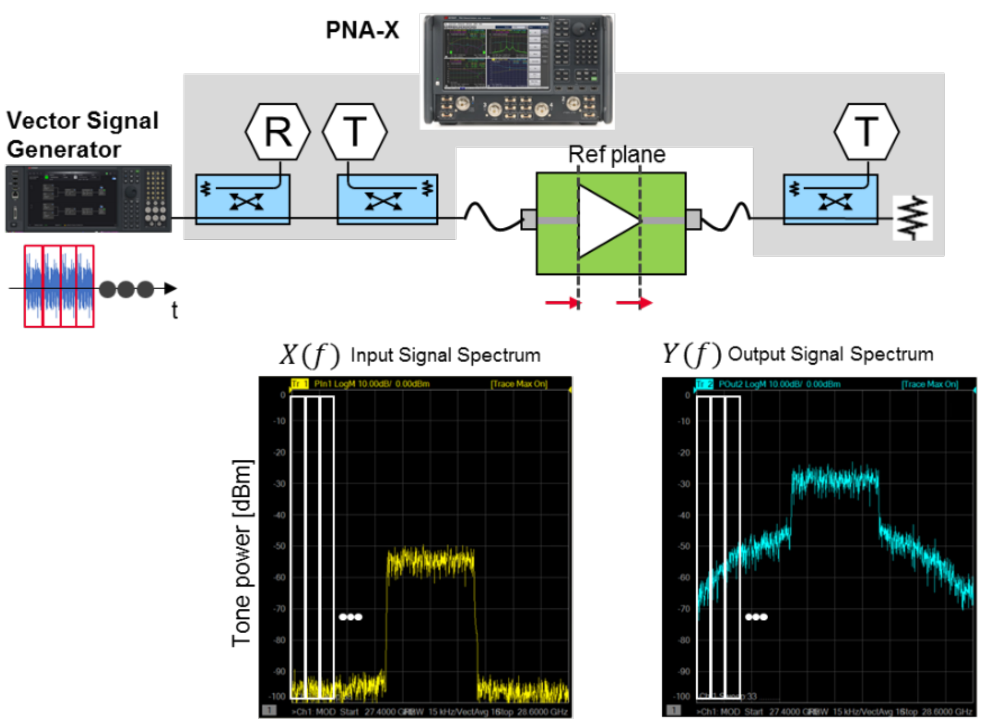


Figure 8. Measurement setup for modulation distortion using Keysight's PNA-X

A vector signal generator (VSG) is used to generate the stimulus. The VSG keeps replaying the CTS repeatedly without any gaps. Three PNA-X receivers are used to capture the spectrum at the reference plane (input and output of the DUT). Note that the instantaneous bandwidth of the PNA-X ADC is about 30MHz. When MOD measures the spectrum of a signal that has > 30MHz, it moves the local frequency of the PNA-X and measures the spectrum for each instantaneous bandwidth, then combines chunks to obtain the entire spectrum response. The PNA-X carefully tunes the DFT parameters of the receiver, so that only signal tones are measured without the image.

When MOD measures the spectrum for each of chunk, it uses multiple receivers coherently and applies linear calibration terms, so that the linear error of the measurement system can be removed from the raw measurement. With this technique, the PNA-X realizes vector corrected accurate measurements at the DUT reference plane.

This figure illustrates how PNA-X makes measurements of the input and output spectrum. The CTS is repeated from a signal generator, while the PNA-X receiver measures the input and output spectrum in the frequency domain. Note that the output signal spectrum has spectral regrowth, which is created by the nonlinear response of the DUT.

Spectrum decomposition

The MOD application has reimagined distortion measurements by using the PNA-X to compare the input and output spectrum using a technique called spectral correlation. Spectral correlation allows the MOD application to decompose the output signal spectrum $Y(f)$ into a linearly correlated part, $H(f)$, and distortion part, $D(f)$. The spectral correlation technique is briefly explained below.

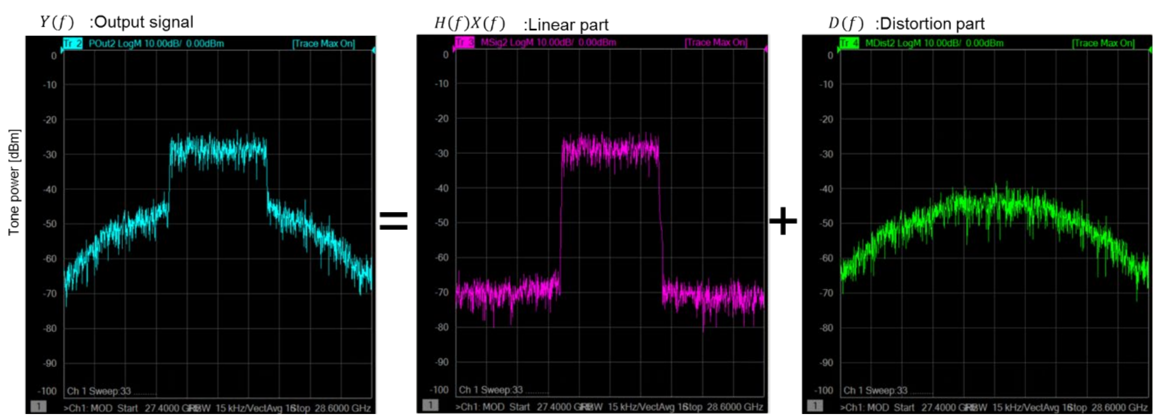


Figure 9. Spectral decomposition performed by modulation distortion application

First, starting from 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 the complex conjugate of X on both side of the equation, 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, $E[.]$ denotes expectation operator. The expectation is practically performed by evaluating the mean of a significant enough number e.g. 100, of adjacent tones. Because there is no linear correlation between the input spectrum and 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:

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

By knowing H, we could now obtain the remaining part, that is the distortion response of the DUT:

(5) $D(f) = Y(f) - H(f)\,\mathrm{X}(f)$

Computing figures of merit (FOM)

Now, we're ready to compute FOMs of the nonlinear response, such as ACPR and EVM.

Computing ACPR is straightforward and is the same as the traditional SG and SA approach, where channel power of in-band channel (BAND) of interest and adjacent channel band (AC) are evaluated, then the ratio between the BAND and AC is computed.

On the MOD application of the PNA-X, EVM is defined by the following equation.

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

In the practical use case of PAs in modern communication systems, compensation of linear characteristics of the PA (such as frequency dispersion) is applied, known as equalization. The suffix e on $EVMe$ in the above equation means this is equalized EVM. Also, the prefix D is added to the EVM, to denote that the EVM on the MOD app is computed from distortion characteristics of the response.

The PNA measures the response of the DUT then calculates the Discrete Fourier Transform (DFT), with the ADC sampling so that the DFT frequency bins exactly land on coherent signal tones of the CTS. Accumulating PSD of each tone in the desired band allow for computation of EVM of the specified band.

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

Since there are two different ways of computing EVM, namely the time domain demodulation method using a VSG and VSA and the frequency domain spectral correlation method using a VSG and PNA-X, it is insightful to review how the two EVM measurements are correlated.

Mathematically, the definition of EVM in both methods is equivalent, per Parseval's theorem, and focuses on 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, the DUT is stimulated by statistically equivalent waveforms and the DUT exhibits the same nonlinear response.

To compare the result of EVM measured in the time domain (VSG and VSA demodulation method) and in the frequency domain (VSG and PNA-X MOD method) a DUT will be measured using a practical parent waveform and a CTS generated from the parent waveform.

Setup

The following figure shows the measurement setup for the correlation study. A broadband power amplifier is used as the DUT. First, the EVM of the DUT was measured with the traditional SA/SG set up by creating a waveform with a specific modulation scheme, then playing it from the VSG. The amplified signal from the DUT was captured by the wideband digitizer of the UXA, then processed by the VSA to compute the EVM in the time domain.

In the MOD setup, the CTS was generated by the same parent waveform above, then the CTS was played continuously by the VSG. The input and output spectrum was measured in the frequency domain, then spectral correlation performed by the PNA to compute the EVM.

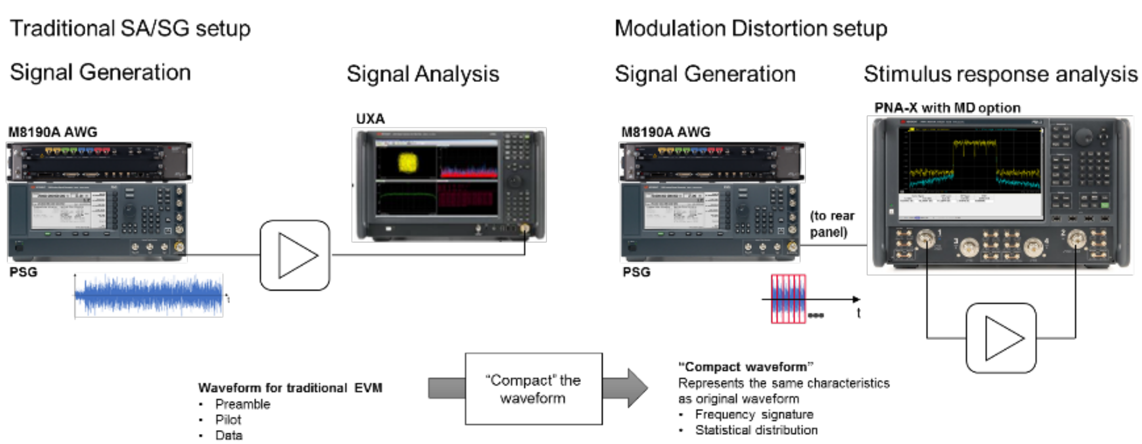


Figure 10. Measurement setup for correlation study comparing VSA and PNA-X method of finding EVM

Result

Several waveforms were used in this study – Verizon 5G 1 channel 100MHz BW, NR 400MHz 64 QAM, NR 800MHz 64 and 256 QAM signals. The DUT was measured at the carrier frequencies of 27, 28, and 29 GHz. EVM is plotted as a function of output power level, known as the bathtub curve.

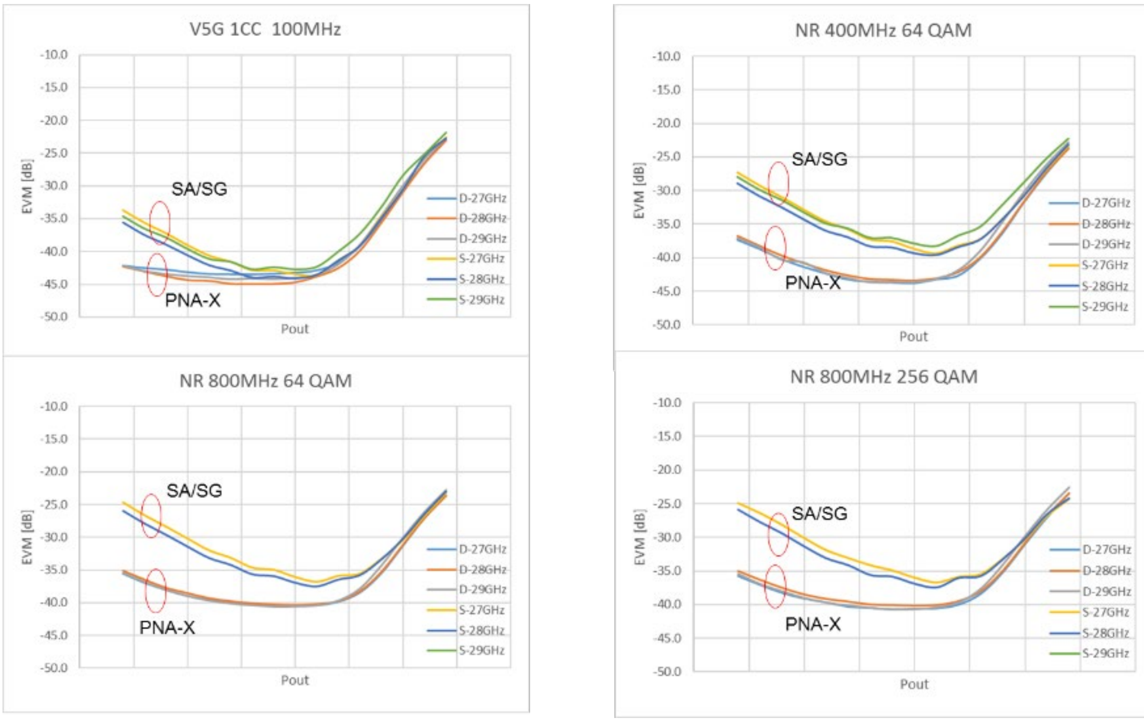


Figure 11. EVM results of correlation study comparing VSA and PNA-X methods

The right-hand side of each plot is the high-power region. EVM is predominantly affected by the nonlinear response (distortion) of the DUT. In this region, there is good alignment between both methods.

The left-hand side of each plot is the low power region. EVM is predominantly affected by the signal-to-noise ratio (SNR) of the measurement system in this region. As a result of lower system noise in the MOD setup, the EVM measurement result is lower compared to the SG/SA method. The lower residual EVM of the test system gives the user better insight into DUT behavior, including DUT nonlinearity as a function of output power.

When characterizing a PA for wide-band applications, it is important to know when the PA starts to show distortion given a specific modulation signal. Using the MOD application in the PNA-X, the user can accurately measure the performance of the PA with lower residual EVM.

PNA-X unique implementation

In this section, the PNA-X's unique implementation of the EVM measurement will be explained, showing how measurements can be done for a power amplifier.

Hardware configuration

The following figure shows the block diagram of the measurement system, that consists of a PNA-X and a VSG. A 10MHz reference is used to synchronize the PNA-X and the VSG. The output of the VSG is connected to the rear panel of the PNA-X (J10) so that the signal goes through the internal path of the PNA-X. Through the internal path, the modulated signal will be available from test port1 of the PNA-X, which is connected to the DUT. The test port2 of the PNA-X is connected to the output of the DUT. The input and output spectrum of the modulated signal is measured using receivers R1, A, and B coherently.

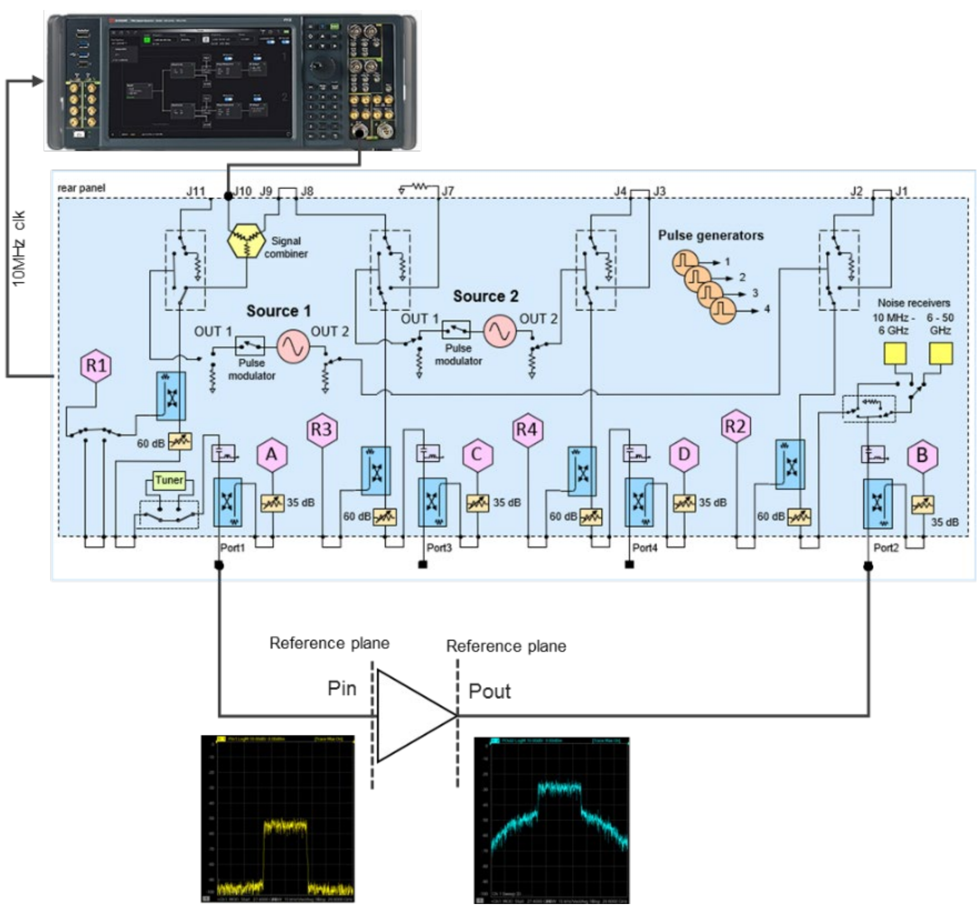


Figure 12. Measurement setup for MOD measurement using external VSG signal generator

In addition to the basic configuration shown above, the PNA-X hardware allows flexible configuration using an external coupler and direct receiver access port. The PNA-X hardware can be configured for high-power amplifier testing.

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).

Modulation distortion application

All the required configurations to make a measurement can be done with the MOD software application integrated into the firmware of the PNA-X. In this section, the detailed setup procedure will be explained.

MOD measurement class creation

Similar to other PNA-X application software, the measurement starts by creating a MOD channel. The channel contains all of the stimulus-response information, as well as calibration information required for the measurement. Similar to the other measurement classes, multiple channels for the MOD class can be created as needed.

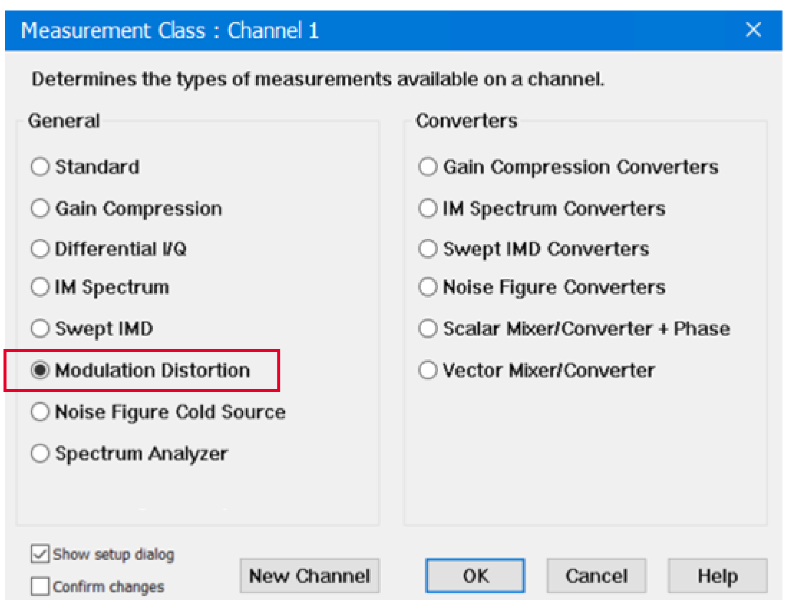


Figure 13. Creating a modulation distortion measurement class in PNA-X modulation distortion application

Modulation distortion setup window

All of the required configurations for distortion measurements can be easily done using a setup window. In this example, a measurement with the below conditions was performed in the block diagram shown earlier.

Example measurement conditions

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

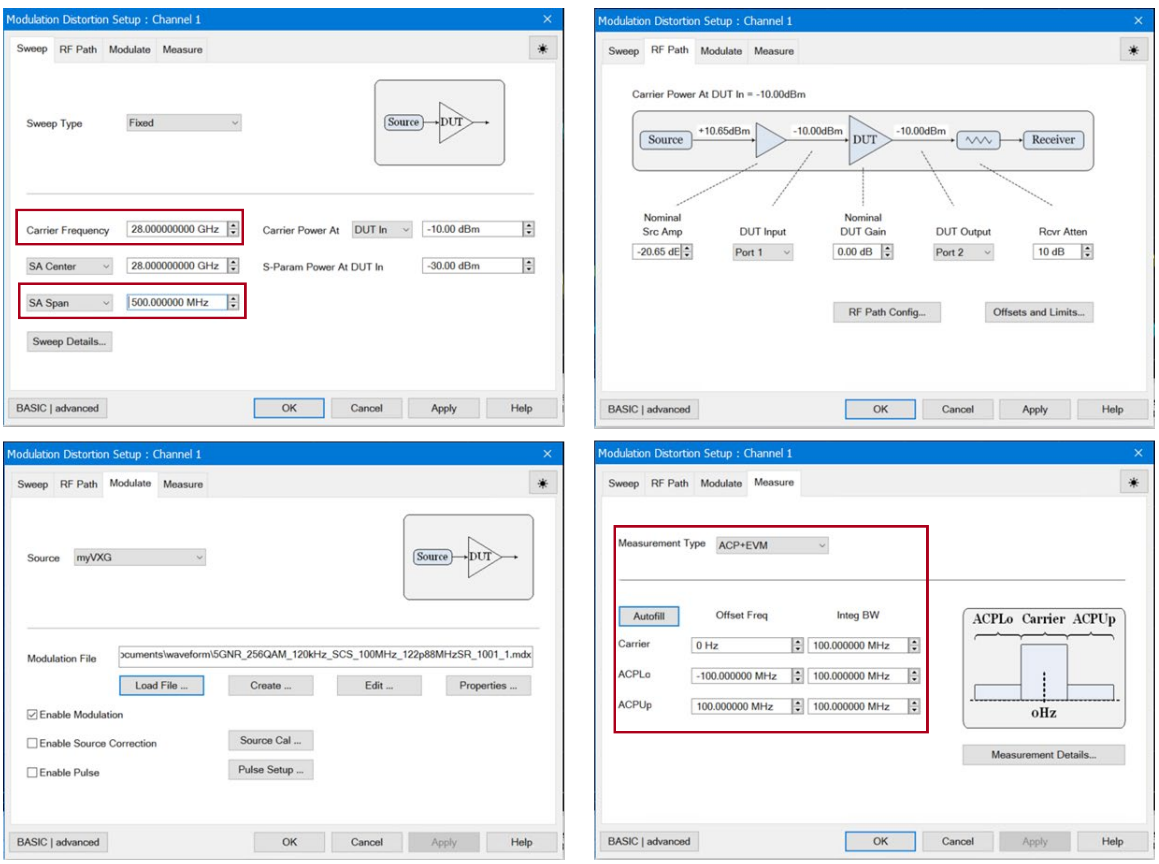


Figure 14. Modulation distortion measurement setup

The VSG used in this measurement can easily be specified in the external device configuration window of the PNA-X firmware. In this example, the VXG is connected via LAN, and the PNA-X is communicating with the VXG via HiSLIP.

CTS creation

The CTS can be generated in the “Modulate” tab of the Modulation Distortion Setup window. In this example, a waveform file (*.wfm file format) generated by Keysight Signal Studio software is used as a parent signal. In addition to the *.wfm file, the *.csv file format can be used.

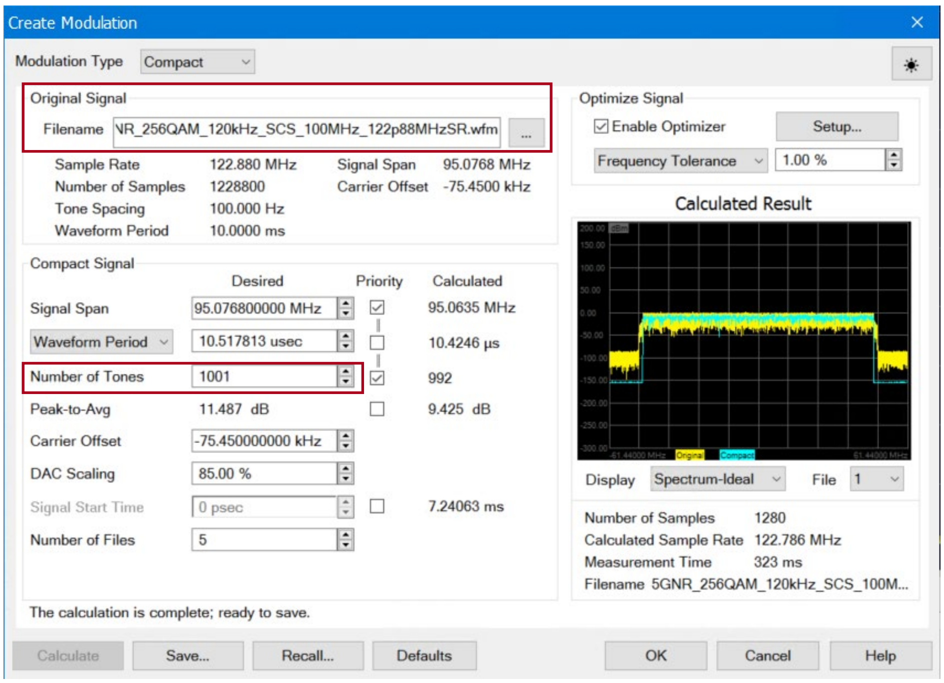


Figure 15. Generating a compact test signal from a parent waveform

In this example, the parent signal has a 10ms of waveform period, with a 122.88 MHz sample rate. Based on the parameters specified in the “Compact Signal” area, the firmware generates a CTS.

In this example, a CTS is created targeting 1,001 number of tones. The PNA-X uses its own algorithm to find the most statistically representative slice from the parent waveform given provided parameters, then applies resampling and brick-wall filtering so that the CTS is optimized for the PNA-X to capture the spectrum.

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

Calibration of The System

Calibration of the measurement system is the critical part to making an accurate EVM measurement. In this section, the procedure of the calibration will be explained using the setup shown in the previous sections.

There are two different types of calibration involved in the MOD application. First, a receiver calibration plane is established that removes the linear error from the raw measurement result. Then, modulated source correction is performed with the DUT connected to optimize the signal across the passband so the desired modulated waveform is at the DUT reference plane.

Receiver calibration

The receiver calibration procedure is the same as other S-parameter calibrations in the PNA-X that utilize the “CalAll” function of the PNA-X. Conventional accessories (mechanical calibration kit, Ecal, and power sensor) can be used for CalAll. During this procedure, the PNA-X uses its internal CW source. The conventional fixturing feature of the PNA-X is available to move the calibration plane as needed.

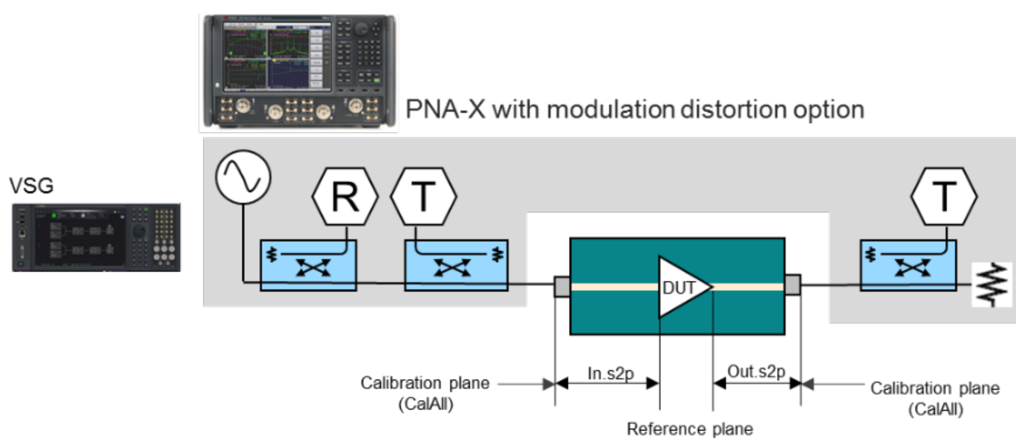


Figure 16. Calibration plane for MOD measurement

By performing this calibration, the PNA-X establishes a calibration plane that removes the linear error of the measurement system when the complex value of the input and output signal spectrum at the reference plane is measured. Note that the corrected measurement removes input mismatch error created by the DUT and test system, then provides accurate amplitude and phase values of the spectrum. Also, with receiver calibration, the amplitude and phase relationship between the input and output of spectrum is accurately measured.

Modulated source correction

Once the receiver correction is made, the measurement system can accurately capture the input and the output signal spectrum at the desired reference plane shown in the figure below. Using this system, modulated source correction is performed so that the DUT can be stimulated by the desired modulated signal. When modulated source correction is done, the mismatch of the DUT is properly accounted for in the calibration.

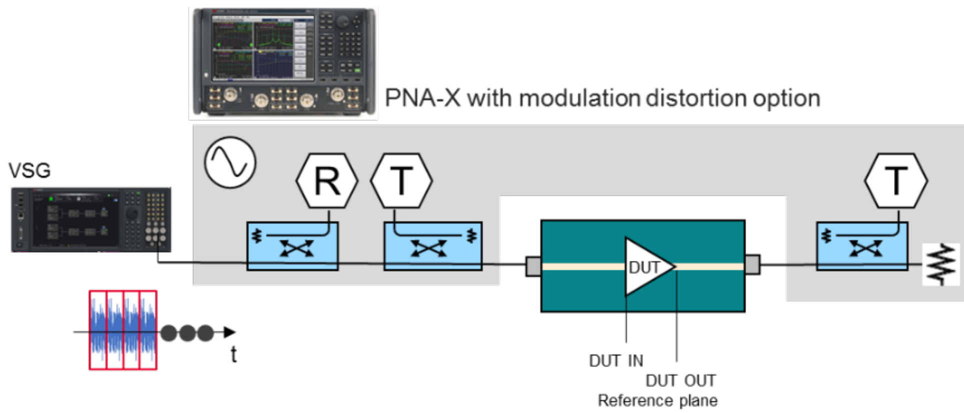


Figure 17. Calibration plane options for source correction

The calibration plane can be selected either from the input (DUT IN) or the output (DUT OUT) of the DUT. Typically, input (DUT IN) is selected to characterize distortion performance of the DUT. Also, as an advanced application, DUT response can be linearized by selecting output (DUT OUT) of the DUT.

There are different types of calibration to perform:

- **Power** calibrates the total output power of the carrier integrated over the signal span.
- **Equalization** performs a linear pre-distortion calibration which equalizes the magnitude and phase of the modulated signal. The Cal Span defaults to the occupied BW of the Carrier Signal.
- **LO Feedthru** minimizes the LO feedthru tone.
- **EVM** minimizes the vector error of the modulation signal over the Cal Span. The Cal Span defaults to the occupied BW of the Carrier Signal. An Equalization calibration is included when a distortion calibration is performed.
- **NPR Notch** nulls the NPR notch. The Cal Span defaults to the frequency range of the notch. If there are multiple notches having different spans, then Cal Span displays Various.
- **ACP Upper/ACP Lower** nulls the ACP upper/lower sideband of the signal. The Cal Span is set to the frequency width of the ACP sideband being calibrated. The default Cal Span is equal to the carrier span

Once calibration has been completed, the modified IQ waveform is played such that the targeted waveform is obtained at the defined reference plane.

Performing the MOD Measurement

In this section, the MOD measurement result of the setup shown in the previous section will be explained.

Distortion measurement

MOD measures the nonlinear response of the DUT to the CTS. In the figure, the input signal spectrum (yellow trace) and the output signal spectrum (blue trace) are shown. Spectral regrowth is observed in the output signal spectrum.

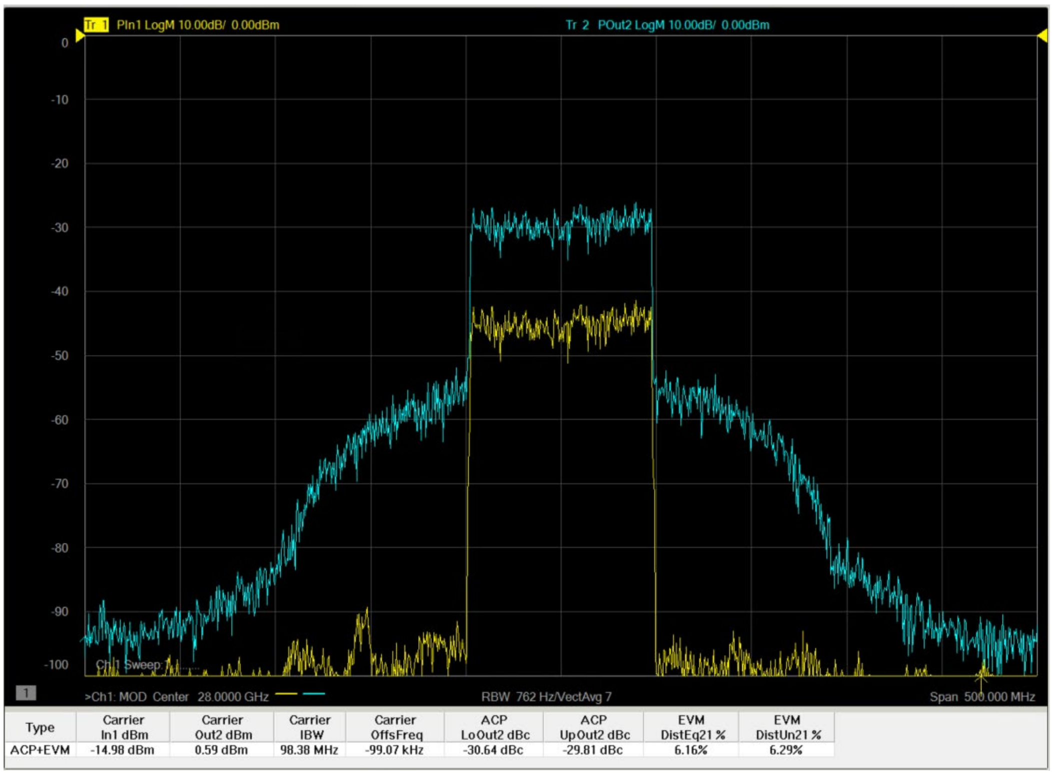


Figure 18. Measured input and output spectrum in MOD distortion application

When EVM is selected as the measurement type in the MOD setup window, spectral correlation of the input and output signal is computed automatically. Spectral correlation and signal decomposition is processed in the background, then the FOMs are computed.

In this example, a CTS based on a 5G NR 100MHz BW signal, at 28GHz carrier frequency has been used to stimulate a DUT, which has approximately 16dB of gain. Spectrum analysis was computed for a span of 500MHz. ACP and EVM are reported in the table below the input and output spectrum.

Making a power sweep

During PA evaluation under a modulated input signal, measurements are done at multiple power levels, then the result is plotted. Using the MOD application, required data can be easily obtained by automating the measurement.

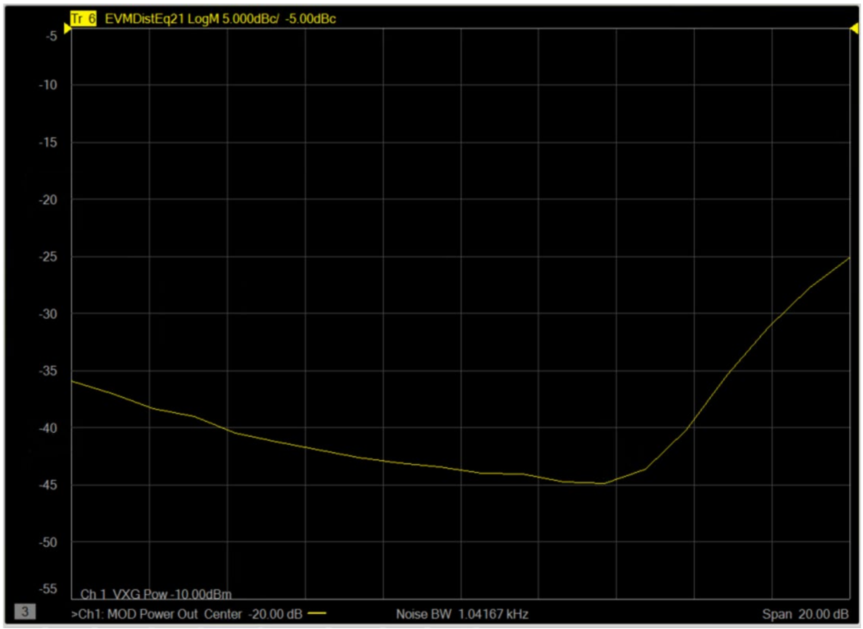


Figure 19. Measured equalized EVM plotted as a function of measured output power

When the DUT response is non-linear, the EVM is predominantly affected by the nonlinearity or distortion of the DUT. In this region, EVM result increases as the power level increases. This can be seen on the right side of the plot.

When the DUT response is linear, noise becomes the key contributor to EVM. This can be seen on the left side of the plot.

The measured noise is a combination of noise from the measurement system and noise from the DUT. Due to the nature of the noise, it is not possible to separate the noise floor from the noise contributions of the DUT, or from the test system. Optimization by the MOD application will also decrease noise from DUT, which will artificially reduce the EVM, giving an EVM result lower than it actually is. To avoid this, the MOD application has a feature to compute the EVM based on the Noise Figure of the DUT, which can be given by the user. The effect of the random noise can be decreased by optimizing the measurement system, which will be shown in the next section.

Optimizing MOD Measurement

It is critical to optimize the setup to make an accurate measurement. In this section, optimization of the MOD measurement will be explained. In the following figure a simplified setup for a MOD measurement is shown, depicting where optimizations for the MOD measurement are made.

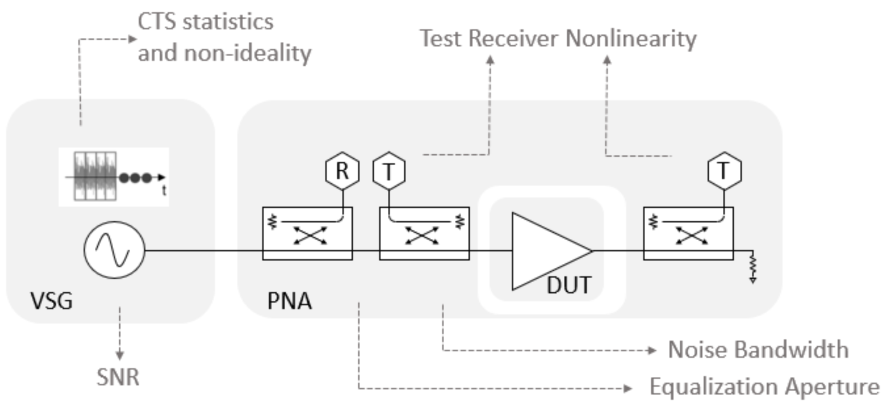


Figure 20. MOD measurement optimization

Signal-to-Noise Ratio of the spectrum measurement

In the left-side of the power vs EVM plot, EVM measurement can be improved by optimizing the SNR of the measurement. When random noise is the dominant factor of the EVM, likely on the left side of the bathtub curve, improving the SNR of the measurement can improve the accuracy of the measurement.

Noise BW (NBW)

The NBW is determined by the Resolution BW (RBW) and the number of coherent averages of the measurement. As a default, NBW is set roughly 1kHz. NBW determines the Signal-to-Noise Ratio (SNR) of the measurement system and measurement time. As NBW is decreased, MOD increases the underlying coherent averaging, resulting in wider SNR and longer measurement time. RBW is automatically set by the firmware based on the CTS waveform length and is a discrete number. The NBW is always a discrete number which is the closest value to the value entered by the user. When random noise is the predominant factor of the EVM, EVM [%] is inversely proportional to the square root of NBW.

SNR of VSG

The SNR of the VSG can differ depending on multiple factors. Practically, the biggest factor is the number of tones of the CTS. As the number of tones is increased, SNR of the VSG decreases. As an example, a modulated signal with 100 tones can be compared with a modulated signal with 1,000 tones, given the same signal bandwidth and channel power. Since the channel power is divided between only 100 tones instead of 1,000 tones, the power level of each tone is 10 times bigger on the 100 tones CTS when the AWG creates the baseband IQ. When the SNR of the source is predominantly determined by the AWG, the SNR of the VSG is directly affected by the number of tones chosen for the CTS.

Nonlinearity of test receiver

If there is a nonlinear response in the test receiver, the PNA cannot distinguish between the nonlinearity that comes from the DUT or from the receiver of the test system. Thus, to optimize the MOD measurement, the PNA-X test receiver needs to not be overdriven when measuring the signal. This is especially important when subtle DUT nonlinearity, such as an EVM level of 1%, is measured. In such a case, it is recommended to keep the power level to be less than -5 dBm at the test port and to use the receiver attenuator to adjust the power level.

Non-ideal CTS

Using the modulated source correction feature, linear error due to the test system can be corrected so that the desired CTS is at the DUT reference plane by correcting channel power and linear flatness response. Also, out-of-band spectral regrowth can be suppressed by correcting ACPR.

It is important to know that the EVM measurement in the MOD application does not require a perfect stimulus signal, as long as the stimulus preserves the statistical characteristics of the parent signal.

CTS statistical characteristics

It is critical to understand that the nonlinear characteristics of the DUT under a modulated signal condition is highly dependent on the stimulus signal. Correcting the stimulus signal spectrum to be the desired value at the DUT reference plane is important, but it is also important to create a CTS that will stimulate the DUT with the most statistically representative characteristics to the practical usage of the DUT.

In modern digitally modulated communication applications, OFDM is a commonly used modulation scheme. When the waveform is based on the OFDM, the statistical characteristic of the signal is a Gaussian distribution.

Peak-to-average

The peak-to-average value is commonly used to describe the statistical characteristics of a waveform, but it is risky to only look at the peak-to-average when discussing CTS statistical characteristics. In a Gaussian distributed signal, the peak-to-average gets bigger as the waveform length increases and/or number of samples (number of tones) increases. Generally, the waveform length of the CTS is much shorter compared to the parent signal. Thus, the peak-to-average value of the parent signal is much higher compared to the peak-to-average of the CTS, unless peak-to-average priority is checked in the CTS creation process.

For most use cases, it is recommended not to check the priority of peak-to-average because of the short waveform length of the CTS. If the CTS is generated to have a peak-to-average equal to that of the parent signal, the resulting CTS waveform can overdrive the DUT, because the peak value of the CTS will be presented to the DUT much more frequently within the same period of time, compared to the parent waveform.

CCDF

It is best practice to align the CCDF of the CTS with the CCDF of the parent waveform. However, as previously discussed, the CCDF can NOT be exactly the same since the CTS is a slice of the parent waveform. As a rule of thumb, it is generally recommended to match the CCDF of the CTS and parent waveforms to 0.1% probability. Typically this can be obtained by using a minimum of 1000 number of tones for the CTS.

Measurement throughput

It is desirable to reduce measurement time to increase throughput. Measurement time can be determined by several parameters.

- SA Span
- NBW
- CTS number of tones
- Reusing linear characteristics

Measurement throughput and measurement accuracy are generally a trade-off. It is important to have the right balance of speed and accuracy and to modify the parameters in the MOD setup depending on the target measurement value.

Equalization aperture

Channel response

Consider the transmission of data over a channel, which could be physical (like a cable) or a wireless medium. An ideal channel will have constant gain and linear phase, meaning 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 is transmitted through a non-ideal channel, the modulated signal will experience frequency dependent magnitude and phase changes due to the non-flat response of the channel, resulting in linear distortion.

The role of equalization

Equalization is used to compensate for this linear distortion due to the degrading effects of a transmission path. Receiver equalization is often used in communication systems to remove linear distortion by compensating for the nonideal effects of the channel. The equalization will have the inverse response of the channel to make the channel appear to be ideal (flat), a simplified illustration is shown in Figure 21.

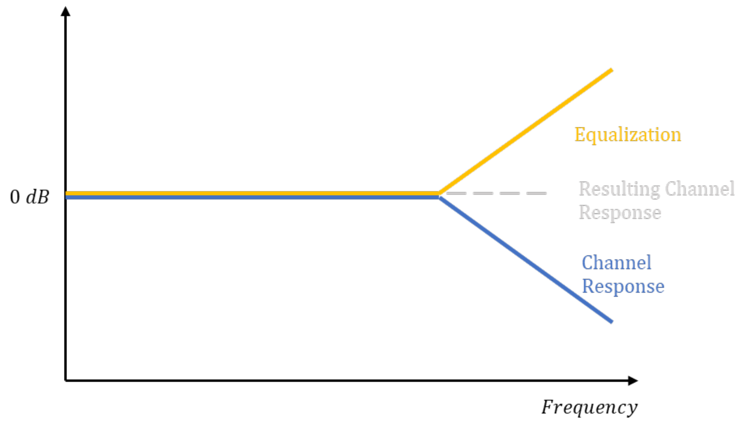


Figure 21. Channel equalization

For channel equalization, the problem then becomes finding the frequency response of the channel.

The role of frequency response in the modulation distortion application

Like the response of a practical channel, the response of an amplifier is not flat versus frequency. In the modulation distortion application, the concepts of frequency response identification are used to isolate the linear from non-linear 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

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

Here $H(f)_{actual}$ is used to indicate the true frequency response function of the DUT, whereas $H(f)_{measured}$ is used to indicate the frequency response function as measured by the modulation distortion application. The goal of the modulation distortion application is to minimize the difference between $H(f)_{measured}$ and $H(f)_{actual}$.

To find the frequency response function, the modulation distortion application uses the modulated input signal as the stimulus signal, instead of sweeping a constant amplitude CW stimulus signal over a large bandwidth to measure this response, like a typical s-parameter measurement. The input spectrum $X(f)$ and output spectrum $Y(f)$ is measured and $H(f)_{measured}$ is determined by linear regression. $H(f)_{measured}$ is reported as MGain21.

The distortion is found by

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

The role of equalization aperture

The equalization aperture is a setting in the MOD application that determines the ability of the linear regression algorithm to fit ripples in the DUT frequency response function, $H(f)_{actual}$. The equalization aperture is the frequency resolution for which $H(f)_{measured}$ is fit to $H(f)_{actual}$. If there is ripple at the output, a narrow equalization aperture will provide more detail of the ripple and a better fit, as illustrated in Figure 21.

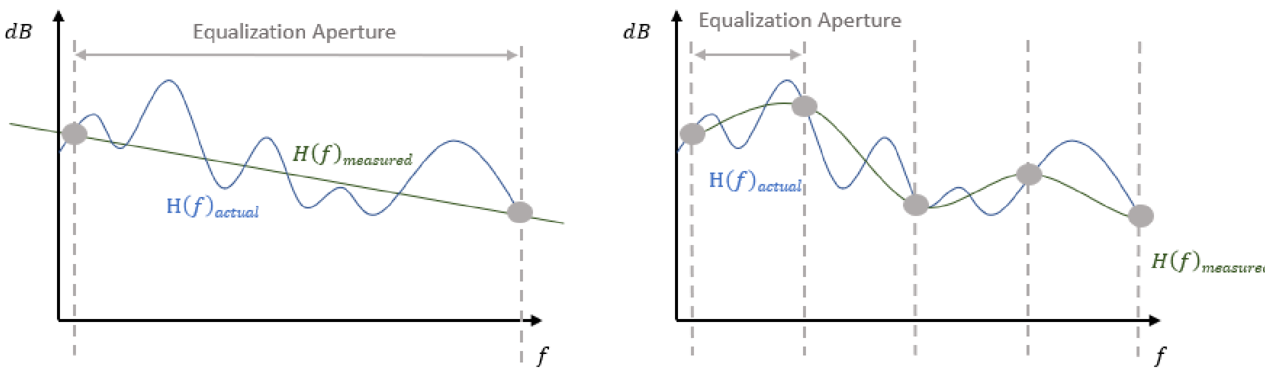


Figure 22. Equalization aperture width determines fit of linear regression algorithm to DUT frequency response function

In these figures, the dotted lines indicate the width of the equalization aperture.

How equalization aperture affects the distortion measurement

If the equalization aperture is too narrow, there will not be enough samples and there will be a large uncertainty in $H(f)_{measured}$. Although a wider aperture will provide less detail of the output ripple, the measurement will be less susceptible to noise. There is 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	Wide aperture
Provides more detail of DUT output ripple	Provides less detail of DUT output ripple
Measurement uncertainty is higher	Measurement uncertainty is lower

Recommendations

In general, the recommendation for the equalization aperture is 5% of the BW, with a minimum of 150 tones. It is typical to use 3000 tones. The appropriate equalization aperture in the frequency domain can be calculated from

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

In the time domain, the appropriate equalization aperture is calculated from

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

The aperture settings denote a “tap equivalent”, that is useful when comparing EVM results from the modulation distortion application and a vector signal analyzer. For receiver equalization, the taps are the values of the discrete filter applied to the received signal to correct for a channel response that is not flat. The number of taps corresponds to the number of ripples within the signal bandwidth and as such corresponds with the

OFDM

When comparing the EVM results from the modulation distortion application and a vector signal analyzer, OFDM results between the two methods are expected to match very well. The reason is that the OFDM equalization in VSA is based on measuring pilot tones and performing the equalization in the frequency domain. This results in a near perfect equalization, just as it is calculated by the modulation distortion application.

Single Carrier

When comparing the EVM results from the modulation distortion application and a vector signal analyzer for single carrier signals, the user will set the equalization aperture in the modulation distortion application to the equivalent equalization filter in the signal analyzer.

An S21 measurement of the DUT in the standard channel will help determine what the equalization aperture should be.

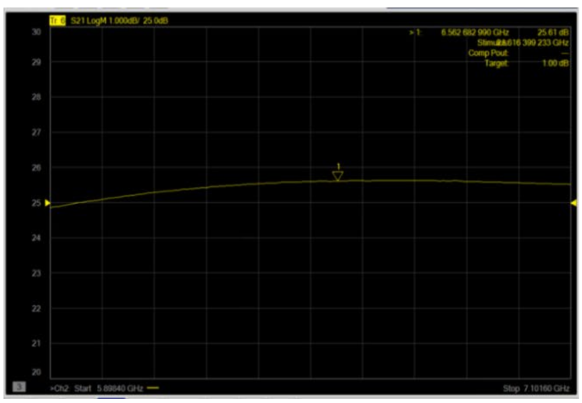


Figure 23. For an amplifier with a relatively flat frequency response function, the equalization aperture can be set to a wider value.

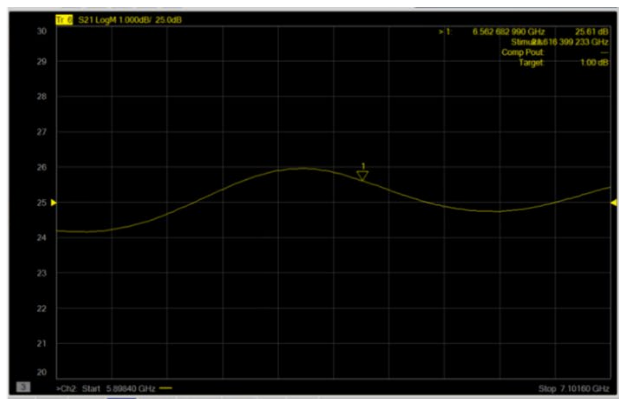


Figure 24. A narrower equalization aperture can be used to fit the ripples.

Limits: Equalization aperture = Too wide

If the equalization aperture is very wide and there is ripple at the output, the frequency response function MGain21, will not fit $H(f)_{actual}$. PGain21 is the tone-by-tone modulation gain, Pout2/Pin1, and is shown in the measurement screen capture below. The equalization aperture setting for this measurement would underestimate the linear contributions to $Y(f)$. This causes a non-ideal equalization and leads to part of the linear dispersion to be included in the reported distortion EVM value. As a result, the distortion EVM value will be slightly higher.

$H(f)_{measured}$ is missing the ripple in $H(f)_{actual}$

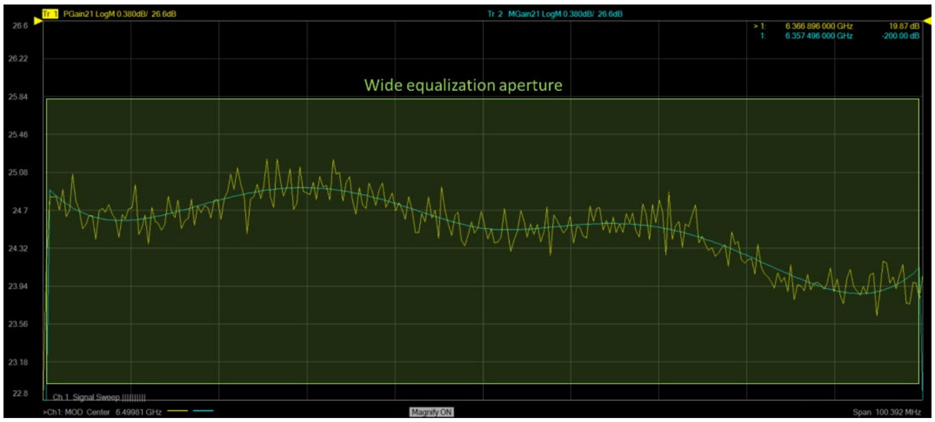


Figure 25. Equalization aperture width too wide and missing ripples in DUT output

Limits: Equalization aperture = Too narrow

The measurement screen capture shows an equalization aperture that is set too narrow. The MGain21 trace appears the same as the PGain21 trace and looks "noisy"; this is due to the reduction in the number of tones used to calculate the gain. When there are not enough tones in the regression interval (equalization aperture) for a meaningful gain measurement there is high uncertainty in the measurement.

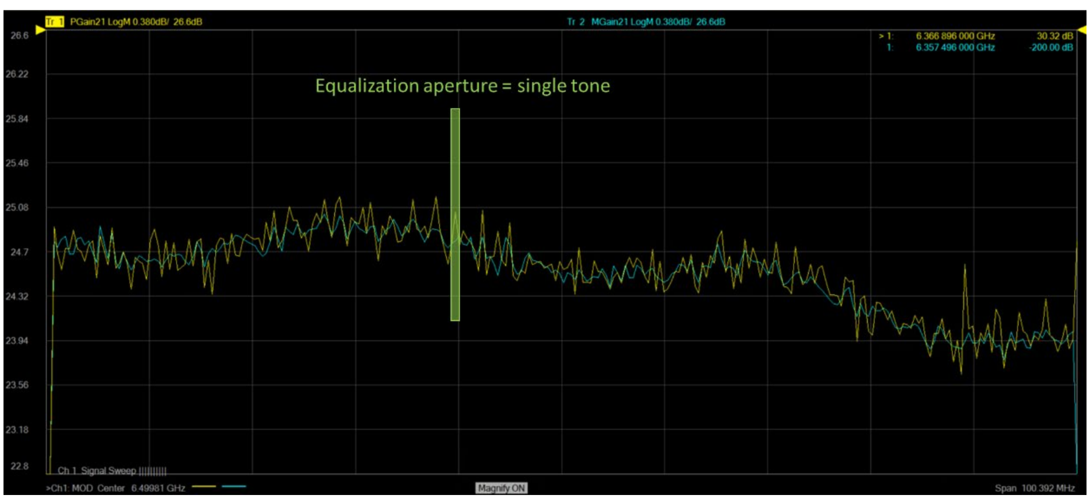


Figure 26. Equalization aperture width too narrow resulting in high measurement uncertainty

Equalization aperture = Reasonable

With an appropriately set equalization aperture, MGain21 follows the ripples in $H(f)_{actual}$.

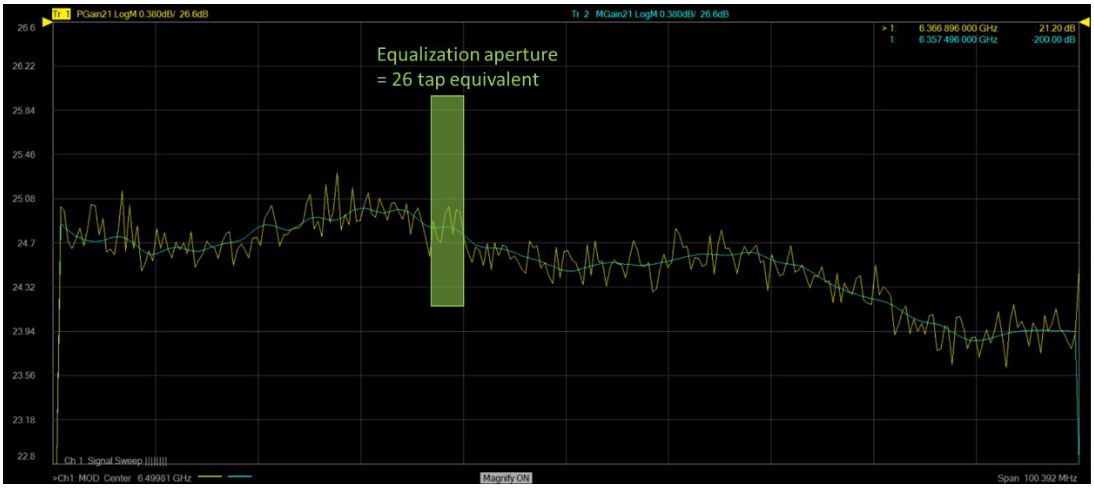
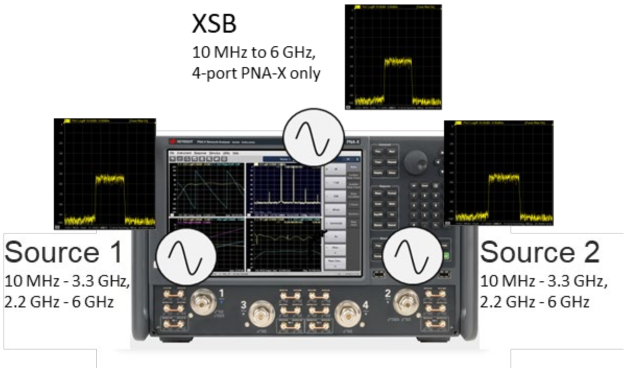


Figure 27. Appropriately set equalization aperture shows MGain trace following ripples in DUT output

XSB

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



The Internal 6 GHz arbitrary waveform generation application uses the Direct Digital Synthesizer (DDS) sources of the PNA/PNA-X and works with MOD or the spectrum analysis application. The waveforms are limited to 6.2866 us (131k samples at 19.2 GSa/s), must be on a frequency grid of (M/N) * 600 MHz and the minimum tone spacing range is 150 kHz

DPD

Modulated signals have a non-constant envelope and a high peak to average power ratio and thus require linear PAs with a wide dynamic range. There are two ways to avoid the distortion caused by the high-power peaks in the modulated signals. The first is to operate the PA with high input back off, so that when there is a signal that has a high power, the PA will not compress. This method means that for a significant percentage of time, the PA is operating at wasted capacity. The second method is to operate the PA for efficiency and figure out what to do about the distortion. There is an inherent tradeoff between linearity and efficiency. Since operating the PA with high input back off reduces the efficiency, reduces the range in signal transmission and shortens battery life, it is not the preferred method to reduce distortion in a PA.

Digital predistortion (DPD) is a method that linearizes the PA by predistorting the input signal with the inverse of the distortion contributed by the PA, resulting in a linearized signal at the output of the PA. The new DPD analysis is within the MOD application and can be used to generate direct DPD and model DPD. The DPD analysis software on the VNA allows PA designers to find the minimum achievable EVM and DPD cost by comparing characteristics of the different DPD models.

Load-Pull

PAs are typically designed for a fixed output load, such as 50 Ω, and their performance can be severely degraded when operating with varying loading conditions in terms of conversion efficiency, linearity, output power capability, and even reliability. Load pull is the process of varying (pulling) the load impedance presented to the DUT so sensitivity or changes in performance, like distortion parameters, across non-50Ω load impedance conditions can be measured.

The modulation distortion application now provides integrated load control for distortion measurements, known as Arbitrary Load Control (ALC) for MOD. This powerful feature computes FOMs such as EVM and ACPR for varying load conditions with a modulated input signal.

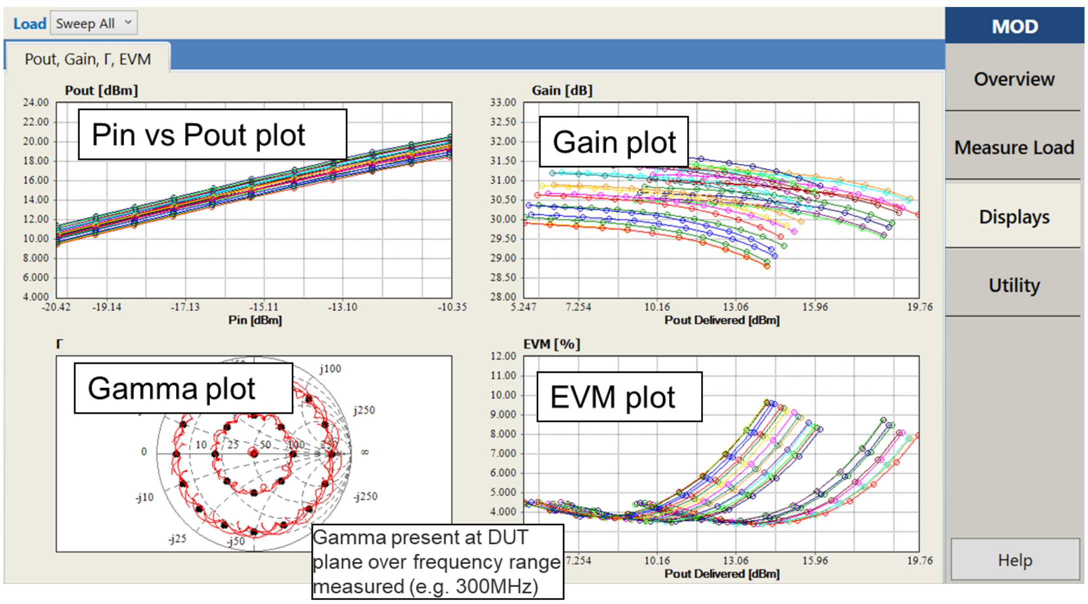


Figure 28. Arbitrary load control for MOD distortion measures power, gain and EVM for varying gamma

Additional challenges in device characterization under wideband modulated input signals

In mobile communication devices, the load impedance presented to the PA varies because mobile devices are rarely stationary and depending on the habits of the user and the location of usage, the range of load conditions may vary a great deal. In 5G 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 ohms, the power amplifier in the front-end of the beamformer IC can exhibit degraded performance. As phased arrays become ubiquitous, load sensitivity tests for power amplifiers are increasing in use and importance.

Performing load-pull measurements

Using ALC for MOD to perform a load pull measurement on a PA under modulated input signal conditions is illustrated in the configuration.

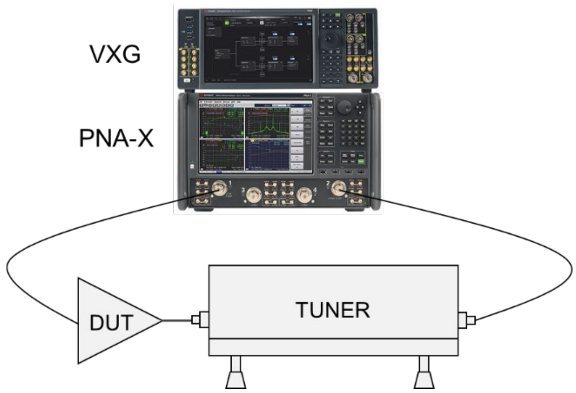


Figure 29. Arbitrary load control for MOD distortion measurement setup

The modulated input signal is provided by the VXG and the tuner is connected as a load to the DUT. The PNA controls the X-Y positions of the tuner, thus varying the load impedance presented to the DUT, at a specific frequency point. Once the impedance at a specific frequency point is set, the gamma is measured for the bandwidth of interest by the PNA and the EVM, NPR and ACPR can be measured as a function of that load condition. ALC for MOD works with Focus Microwaves and Maury Microwave tuners.

About passive tuners

Passive electromechanical tuners typically have two parallel plates, a precision 50 Ω slabline center conductor and a metallic probe that can be moved in the X and Y directions. The probe position of an electromechanical tuner is precisely controlled by a motor. Probe movement in the Y direction will disrupt the electric field of the signal entering the tuner and will change the magnitude of reflection (gamma). Probe movement in the X direction (towards or away from the DUT) changes the phase of the reflection. In this way, nearly any impedance can be presented by the tuner to the DUT. The range of impedance values that the tuner can present to the DUT is limited by the type of tuner used and any loss in the test set. Passive tuners reflect the signal back to the DUT and the reflected wave is always less than the incident wave.

Calibrating the tuner

Tuner calibration maps the X and Y positions of the probe in the tuner to s-parameters. The mapping is used by the PNA-X during load pull measurements to present user specified impedances to the DUT. Using a standard channel on the PNA, the user is guided through tuner calibration for a carrier frequency of interest. Fundamental tuners, such as those used in the ALC for MOD application, set one specific frequency gamma, but are uncontrolled at harmonics.

Setting the load grid

The load grid is the user defined set of impedance values that will be presented to the DUT during the MOD measurement. The ALC for MOD software makes generating the load grid simple by creating a grid from magnitude and phase values or custom values entered by the user.

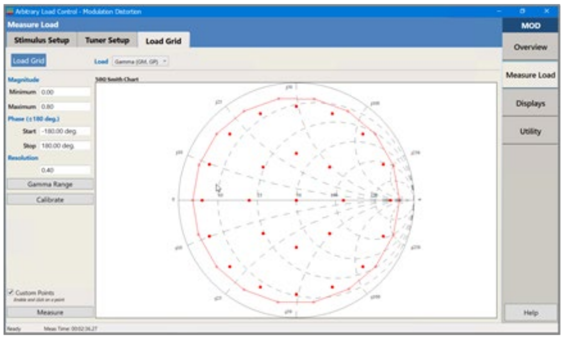


Figure 30. User defined gamma load values

Calibrating the load grid

For accurate load pull measurements, the gamma presented at the DUT reference plane must be precisely measured. Since the MOD channel is calibrated with the tuner in the initialized state (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. By creating delta s2p data at each gamma point during the load grid calibration for use during the measurement, accurate gamma values are presented at the DUT reference plane. This maintains the validity of the PNA calibration for not only the 50 Ω gamma value, but 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, shown below.

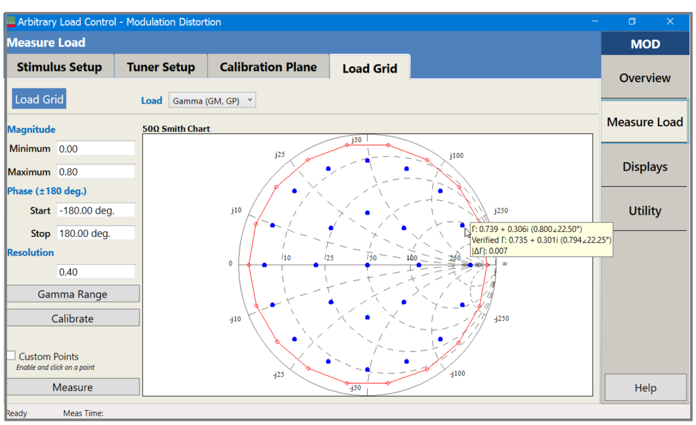


Figure 31. Load grid calibration for precision measurements

Optimizing the load-pull measurement

Meticulous care must be taken to present the specified load impedance values at the DUT reference plane. This means that during calibration of the MOD channel, tuner and load grid the user needs to know where the reference planes are for the DUT and the tuner, and also needs to account for any additions to the test set that are added after tuner calibration. The ALC software guides the user through including any front block, components that are between the DUT reference plane and the tuner calibration plane, so that the correct gamma is presented to the DUT.

The goal of a load-pull measurement is to accurately present varying load impedances at the DUT reference plane (shown as gray dotted line) and measure DUT performance. The gamma presented by the tuner is at the tuner calibration plane (shown as a red dotted line) and for accurate measurements in the test set shown, would need to be moved to the DUT reference plane. Note that since this is a load-pull measurements, only the calibration plane on the load side is of interest.

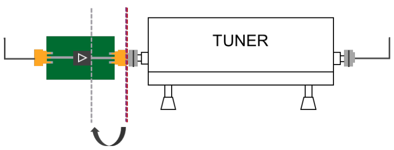


Figure 32. Aligning tuner and VNA calibration planes

The ALC software shows the possible locations of the DUT, tuner and PNA calibration planes. The user is guided through the steps needed to align the planes, if applicable for the user's test set configuration.

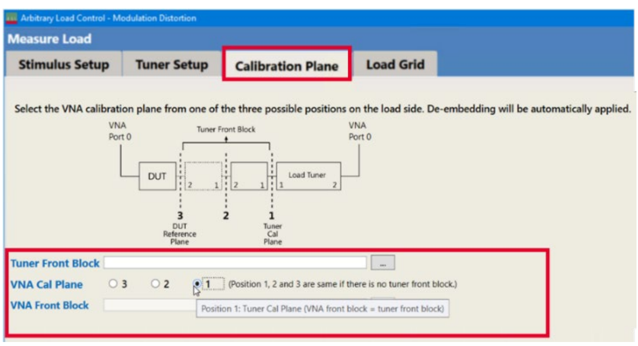


Figure 33. Calibration Plane tab in ALC guides user to correctly de-embed the setup

Another consideration for load-pull measurements is to minimize any loss in the test set to achieve the maximum gamma range. The smith chart illustrates how small amounts of loss can limit the maximum gamma that can be presented to the DUT.

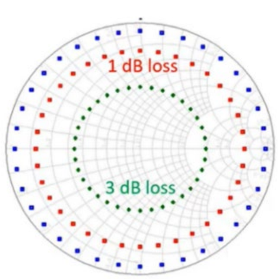


Figure 34. Test set loss limiting maximum gamma range

VSA-Link

The PathWave Vector Signal Analysis (VSA) software is a comprehensive set of tools for demodulation and vector signal analysis. VSA-link allows for the IQ data from the PNA-X to be sent to the VSA where the IQ data can be demodulated and the EVM calculated in the time domain.

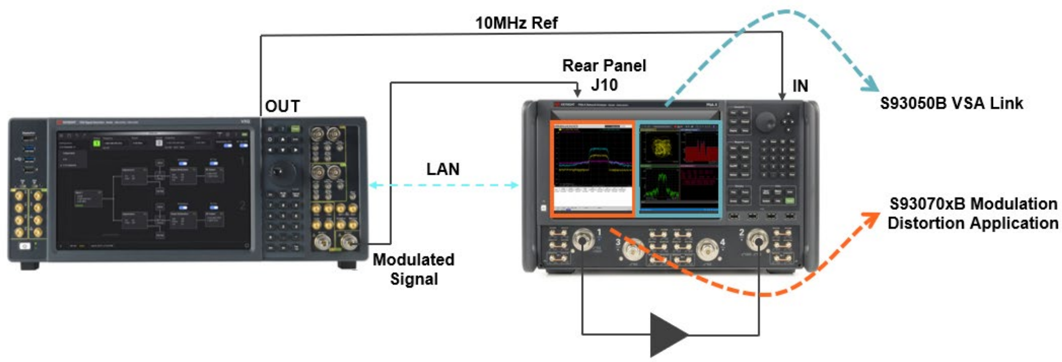
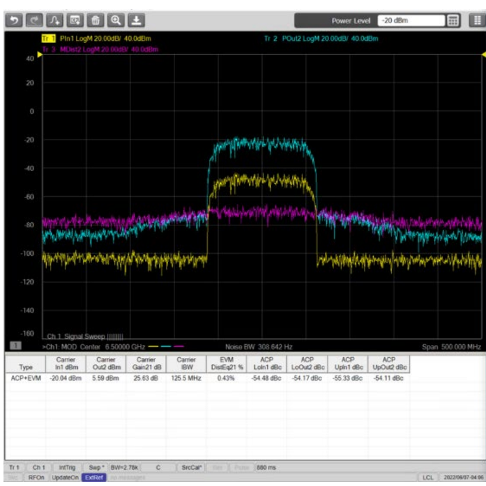


Figure 35. Modulation distortion measurement with VSA link for signal demodulation

After the PNA-X acquires the data and performs the IFFT, the IQ data is sent to the VSA for AM/AM, and AM/PM measurements to support ultra-wideband analysis plus full EVM demodulation.

For PA designers who may need EVM under a specific modulation scheme such as 5G NR, Wi-Fi, DBSX2, etc, VSA-link is particularly useful. In addition to viewing memory affects and gain compression, the correlation of EVM measured in the time domain to the EVM measured in the frequency domain can be performed.



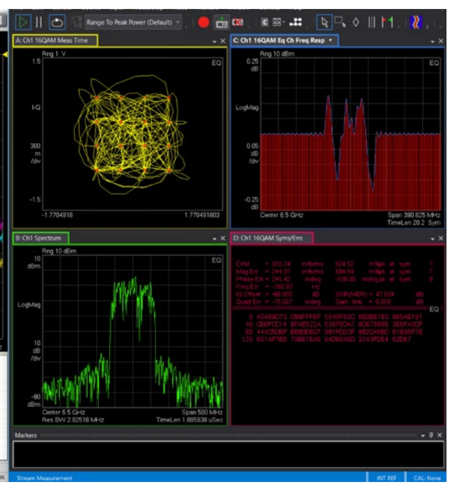


Figure 36. MOD distortion measurement EVM results

Figure 37. MOD distortion measurement IQ data streamed to VSA for demodulation and EVM calculation

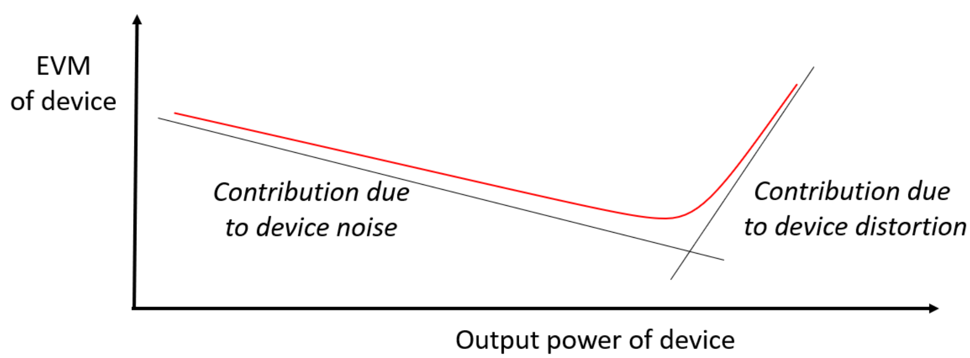
Appendix A: Technique Comparison for Measuring EVM for Amplifiers

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

For system level and transmitter testing, EVM measurements are very important since the aggregate EVM contains contributions from non-linear distortion, added noise, phase noise of the carrier and potential assimilation onto subcarriers, and IQ imbalances such as quadrature, gain differences, and DC offsets. These errors all 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, which help designers understand the individual contributions.

Now consider the EVM contribution of a component such as a power amplifier used in a system or transmitter. Contributions from distortion or noise will dominate the EVM of a component based on the amplifiers operating point while 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. It's true that the amplifiers do generate residual phase noise, but typically the contribution from residual phase noise can be disregarded in comparison to the noise or distortion contribution.

Summarizing the error contribution of EVM for a component can be shown using a typical "bath-tub" curve for an active component (see Figure). The "bath-tub" curve represents the measured EVM vs. 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 and at high power, the EVM is dominated by non-linear distortion. 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.



An additional error that has not been 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 error from the ideal signal, that will set the minimum EVM measurable by the receiver.

For Modulation Distortion on the PNA-X, the distortion EVM measurement focuses on the key contributions of the device and is immune to input signal EVM since it is measured simultaneously with the output signal. The measured output signal is then separated into a linear portion which yields gain and the non-linear distortion contribution EVM. The measurements are done with very high precision and high signal-to-noise measurement of the PNA-X receiver. By eliminating the contribution of the input signal and the good sensitivity of the PNA-X receiver the residual EVM noise floor is improved vs traditional VSA approach for devices.

	89600 VSA and X-Apps	**S93070xB modulation distortion**
Supported instruments	Keysight signal analyzers, oscilloscopes, PXI VSAs	Keysight PNA-X
Benefits	Standard compliant algorithms Flexible views of EVM (time, frequency, subcarrier, etc) Constellations diagrams Measures all system and transmitter EVM contributors	Very wide measurements limited only by signal generator bandwidth Lower residual EVM enabled by very good sensitivity and eliminating contributions of the input signal Measures contribution of the device only

Appendix B: Spectral Correlation Theory

The Modulation Distortion is a frequency domain analysis method. Nevertheless, as will be shown in Appendix C, the total amount of distortion measured in the frequency domain is equal to the total amount of distortion 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, 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 this document is based on 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 as

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

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 referred to as auto-correlation). 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 referred to as 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 is not linearly correlated with $X(f)$. The linear uncorrelatedness is mathematically expressed as

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

Note that $c_G(f)$ is a complex valued continuous frequency response function, equal to 1 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.

Equations **Error! Reference source not found.** and **Error! Reference source not found.** can be considered as the constitutive set of equations defining the quantities $c_G(f)$ and $D(f)$. It is assumed in the following that, for a given amplifier, $c_G(f)$ and $S_{DD}(f)$ are uniquely and solely determined by 2 things: the power spectral density of the input signal $S_{XX}(f)$, on one hand, and the statistical distribution of the amplitude of the corresponding time domain complex envelope $x(t)$ on the other hand. This statistical amplitude distribution is commonly expressed by the complementary cumulative distribution function

(CCDF) of the instantaneous power levels. The CCDF is a function of power and expresses the percentage of time the instantaneous output power is higher than any given power level.

The modulation distortion measurement problem is defined as follows. Given $X(f)$ corresponding to a given CCDF and PSD, decompose $Y(f)$ into the component $c_G(f)g_{SM}(f)X(f)$, which is linearly correlated with $X(f)$, and the remaining part $D(f)$, which contains the nonlinear distortion. The measurement procedure is explained in the following.

One starts by measuring $g_{SM}(f)$ using a VNA. Next, one excites the DUT with a signal $X(f)$ that has a CCDF and PSD corresponding to the modulation format and power level for which one wants to know $c_G(f)$ and $D(f)$. Next, one uses 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. Next 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:

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

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

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

With a dense enough tone spacing the expectation operator $E[.]$ 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 **Error! Reference source not found.** in to **Error! Reference source not found.** 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

$S_{DD}(f) = \left(1 - \gamma_{XY}^2(f)\right) S_{YY}(f)$, 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 **Error! Reference source not found.**,**Error! Reference source not found.** and **Error! Reference source not found.** are derived as follows. For the sake of compactness, the frequency dependency will be omitted in the following equations. Multiplying both sides of **Error! Reference source not found.** by X^* and calculating the expectation results in

$E[X^*Y] = E[X^*(c_G g_{SM} X + D)]$, which can be written as

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

Substitution of **Error! Reference source not found.** into **Error! Reference source not found.** and solving for c_G results in **Error! Reference source not found.**.

Multiplying both sides of **Error! Reference source not found.** 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 **Error! Reference source not found.** into **Error! Reference source not found.** 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

$S_{DD} = \left(1 - \frac{|S_{XY}|^2}{S_{XX} S_{YY}}\right) S_{YY}$, thereby proving **Error! Reference source not found.** and **Error! Reference source not found.**.

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 both the CCDF and the PSD of a given modulation format, e.g. 5G NR. Because the VNA can only detect discrete tones, the input signal is required to have a discrete spectrum, or in other words, the input signal needs to be periodic. In the following it will be explained how such a periodic input signal is created.

One starts by generating a time-domain “parent IQ-waveform” that is assumed to be representative for the modulation format one wants to investigate. This could, for example, be an IQ-waveform corresponding to a 10ms long radio-frame of a 5G NR signal with a modulation bandwidth of 800MHz. One option would be to simply repeat the radio-frame; that would result in a signal with a discrete spectrum with a tone spacing equal to the reciprocal of the period, in this case, 100Hz. The total number of tones within the signal bandwidth would be equal to 8,000,000. As one typically measures across a span equal to 3 times the bandwidth of the input signal with the goal of characterizing 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 not efficient and takes a prohibitively long time. The option we have chosen 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 to reduce the number of tones has been automated and is explained in the following.

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□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 a 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, referred to as a compact test signal (CTS), matches the CCDF and PSD of the parent waveform and is used as the input signal.

Appendix D: Relationship to existing EVM and ACPR

In this appendix, the relationship between the existing methods for measuring EVM and ACPR and the new method are explained.

With EVM one measures the output signal $y(t)$ using a VSA. For modulation formats based on orthogonal frequency domain multiplexing (OFDM), common for 4G, 5G and Wifi, EVM can be defined as the NMSE between the bandpass filtered $y(t)$, compensated for complex gain, group delay and frequency dispersion, and the ideal input signal $x(t)$, whereby $x(t)$ is derived through demodulation of $y(t)$. Note that EVM is typically expressed in % or in dB. EVM is given by the following equation:

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

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. Both the numerator and denominator present in the righthand side of **Error! Reference source not found.** are sums of amplitudes squared and are as such equal to their frequency domain equivalents, in accordance with Parseval`s theorem. This can be expressed as follows:

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

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 $Y(f)$:

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

whereby the frequencies f_i correspond to the in-band frequencies of the compact test signal. Considering **Error! Reference source not found.** and **Error! Reference source not found.** one concludes that the classic measure of EVM is identical to DEVM$_e$ under the condition that the equalization filter that is 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 $E(f_i)$ simply represents a group delay τ and a complex gain G. In that case **Error! Reference source not found.** and **Error! Reference source not found.** become

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

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, as well as the nonlinear contributions, whereas DEVM$_e$ only contains the nonlinear contributions. It is important to note that our method enables to accurately determine EVM without the need for demodulation, which greatly simplifies and speeds up the data processing.

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

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