APPLICATION NOTE

Comprehensive Test of Mixers and Frequency Converters

PNA-X Application Note 1408-23

KEYSIGHT

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Introduction

Frequency-translating devices translate a band of input frequencies to a different band, one that is either higher (up-converted) or lower (down-converted) in frequency. This crucial function makes them integral to most RF applications, including wireless-communications systems and in the transmit and receive portions of radar and electronic-warfare systems. This application note describes how a modern vector-network analyzer (VNA) like Keysight's PNA-X Series can be used for comprehensive testing of mixers and frequency converters.

Definition of mixers and converters

Mixers are the core elements for providing frequency translation. An input signal is mixed or multiplied with a local-oscillator (LO) signal that is used to switch diodes or transistors on and off, producing sum and difference products. In most cases, only one of these products is desired, and the other is suppressed by filtering. Mixers can be passive or active devices (Figure 1), and depending on the application, the input signal is provided to the IF port for up-conversion, or the RF port for down-conversion. While the frequency-translating or mixing process is inherently nonlinear, mixers are otherwise expected to behave linearly – for example, a 1 dB input-signal change should result in a 1 dB output-signal change, and a mixer's magnitude and phase responses versus frequency should be independent of input power. However, just like amplifiers, mixers have linear and nonlinear regions of operation, depending on the input drive level.

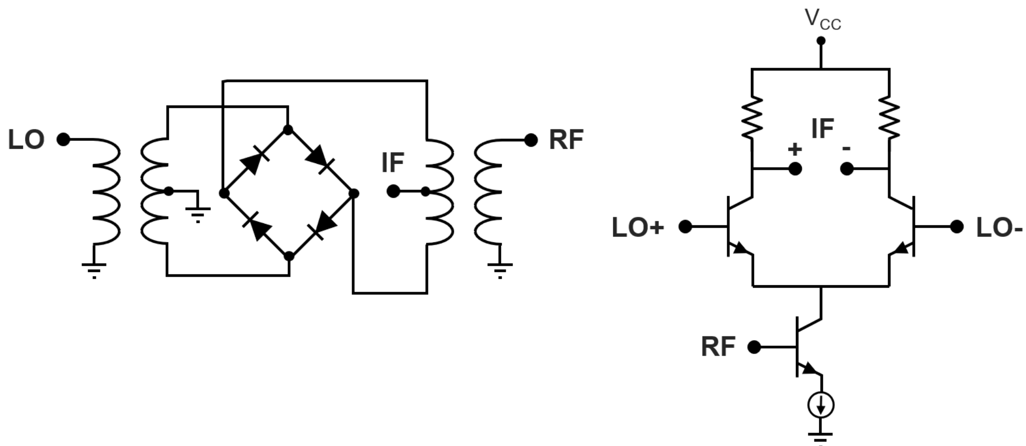


Figure 1. Examples of a passive diode-based mixer (left) and an active transistor-based mixer (right).

Frequency converters are more complex assemblies that contain one or more mixers, along with amplifiers and filters and possibly other signal-conditioning components like attenuators, isolators, limiters, and phase shifters. Each of the mixers within the frequency converter requires an LO signal, which for test purposes, is either supplied from the VNA, an external signal generator, or from an internal or embedded oscillator within the DUT.

Many of the measurements performed on mixers and frequency converters are familiar RF tests that are commonly performed on amplifiers, such as gain, gain flatness, group delay, gain and phase compression, intermodulation distortion (IMD), and noise figure. In this application note, each of these measurements and more will be reviewed in more detail.

Test approaches

Many frequency converters are still tested on legacy systems consisting of a stack or rack(s) full of RF test equipment. Compared to the modern methods covered in this application note, these test systems are significantly slower and less accurate as well as being more difficult and expensive to configure and maintain. The most common approach for testing frequency converters has been to use stand-alone signal generators for RF and LO signals, and a spectrum analyzer as a measurement receiver. This approach is conceptually simple, but in practice has many drawbacks:

- The test configuration is difficult to set up, as lots of external signal-conditioning components are needed, such as combiners, filters, and attenuators. It is often difficult to switch between different tests, as components need to be added or removed.
- Typically, only transmission-magnitude measurements like conversion gain are performed, requiring VNAs to characterize port matches, deviation from linear phase, and group delay.
- Error correction is limited to magnitude-response corrections, so attenuators are often required to reduce mismatch errors, at the expense of signal-to-noise ratio (SNR).
- Automated, swept-frequency or swept-power measurements require a computer and software to synchronize the various instruments, resulting in sweeps that are many times slower compared to those of a VNA

A modern VNA like the PNA-X is well suited for testing frequency converters due to flexible hardware coupled with many software measurement applications. This combination means that many different tests can be performed while maintaining a single set of connections to the device-under-test (DUT). Tests include both forward and reverse linear characterization with magnitude and phase, as well as nonlinear tests of compression, distortion, and noise figure. Since most of the needed hardware is contained entirely within the instrument (Figure 2), speed can be optimized, resulting in throughput improvements of up to 100 times compared to typical legacy systems. Advanced error-correction methods have been developed for all measurement applications for highly accurate results. For example, mismatch errors are removed for swept gain, delay, and noise figure measurements, and power meters are used to calibrate the test system to get high accuracy for both setting and measuring absolute power, which is essential for compression and IMD measurements. A lot of effort has also been put into designing simple, intuitive user interfaces and guided calibrations to make a complex instrument relatively simple to use.

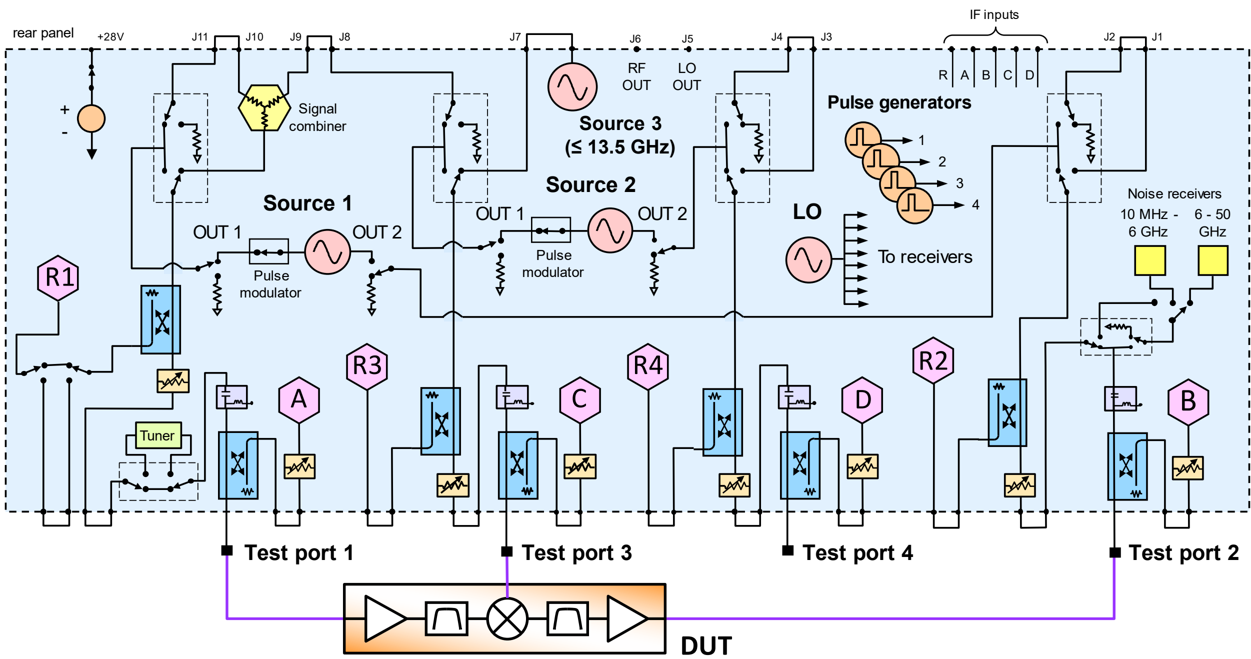


Figure 2. The PNA-X contains the main elements needed for frequency-converter characterization including a full S-parameter test set, multiple RF sources, a built-in signal combiner, pulse hardware, and a low-noise receiver. Shown is the configuration for intermodulation-distortion testing.

Challenges of converter test

When characterizing transmission parameters of frequency-translating devices, standard S-parameter methodology cannot be applied. S-parameters are defined as complex (magnitude/phase) ratios of incident (a_n) and test signals (b_n). When the stimulus and response frequencies are the same, which occurs for non-frequency-translating devices, complex ratios can be directly measured as receiver ratios like b_2/a_1, which is the core measurement of S_{21}, or b_1/a_1, which is the core measurement of S_{11}. Data acquisition for all the receivers is simultaneous, so a single sweep in each measurement direction (forward and reverse) is sufficient to gather the necessary data for a two-port set of error-corrected S-parameters. Testing port matches of mixers and frequency converters is similarly easy since for reflection measurements, the incident and test signals are at the same frequency, allowing use of normal S-parameters. However, for frequency-translating transmission measurements, the input and output frequencies are not the same, so direct receiver ratios cannot be used.

When testing transmission parameters of frequency-translating devices, two challenges must be overcome: offsetting the stimulus and response frequencies and calculating receiver ratios when the frequencies differ. With modern VNAs like the PNA-X, the frequency-offset challenge is easily overcome, since all the internal RF sources (including the one that provides the LO signal for the measurement receivers) are fully synthesized and can be independently set to any frequency within the range of the instrument. It has not always been this easy however, as earlier generations of VNAs required the insertion of an external mixer into a phase-locked loop to offset the source and receiver-LO frequencies. Due to the abundance of spurious signals with this arrangement, it was difficult to achieve phase lock over a broad frequency range, and spurious signals often showed up in the measurement results.

The second challenge involving receiver ratios is also not difficult to overcome for transmission-magnitude measurements, as direct receiver ratios are not necessary. At the expense of measurement speed, input power can be measured in one sweep, output power in another, and the individual receiver measurements can be ratioed. In receiver terms, the magnitude-conversion response $|SC_{21}| = |b_2|/|a_1|$, where $|a_1|$and $|b_2|$are measured with two sequential sweeps. This measurement is very accurate since the PNA-X performs match-corrected power measurements that are calibrated using a power meter and an S-parameter calibration kit or ECal electronic-calibration module.

Frequency-translating phase measurements

Transmission-phase measurements are particularly challenging for frequency-translating devices, since phase is generally measured between signals at the same frequency. With earlier VNAs, it was necessary to use a reference mixer whenever phase or group delay measurements were required. The reference mixer was either placed in series with the main signal path (often called the down/up or up/down-converting method) or was placed in a parallel path to the DUT (Figure 3). There were pros and cons to each approach, but either way, the reference mixer had to cover the frequency plan of the DUT, and it required its own LO signal. Often, a separate calibration mixer was also used to calibrate the test system.

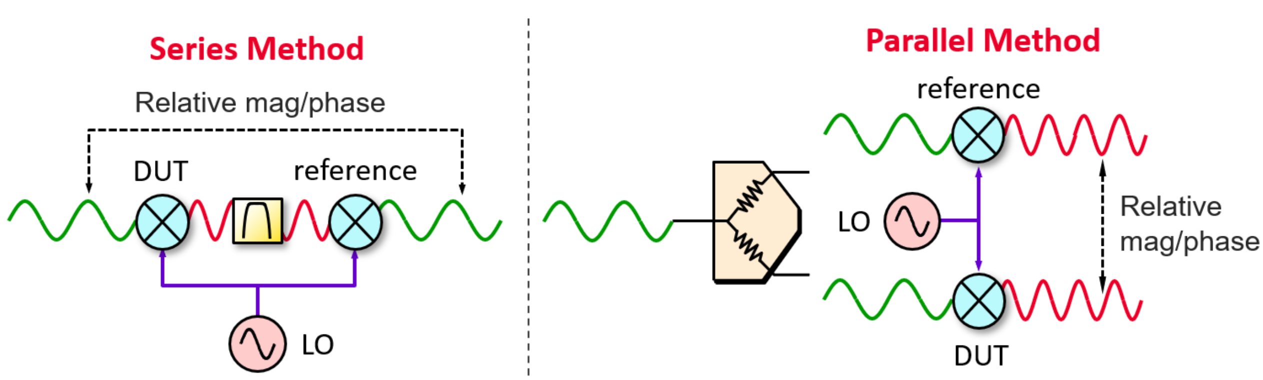


Figure 3. Traditional setups for measuring the phase and delay of mixers and frequency converters require series or parallel reference mixers.

Modern VNAs eliminate the need for a reference mixer for most frequency-converting measurements of phase and group delay by taking advantage of modern synthesizer hardware that supports phase-coherent frequency sweeps, something that earlier-generation VNAs could not do. With phase coherency, the phase response versus frequency of a single receiver is repeatable, allowing the phase responses of different receivers measured over different frequency ranges to be ratioed, similar to the previously discussed ratio of magnitude responses. This capability eliminates the need for a reference mixer, greatly simplifying test setups.

Core Measurements Using SMC, SMC+Phase

The Scalar Mixer/Converter (SMC, S93082B) and Scalar Mixer/Converter Plus Phase (SMC+Phase, S93083B – a superset of S93082B) are the core measurement classes for mixer and frequency-converter characterization using a PNA-X. A reference mixer is not needed and using frequency and power sweeps, measurements of output power, gain, gain flatness, gain compression, phase deviation, phase-versus-drive (AM-to-PM conversion), group delay, and port matches can be performed. Examples of these measurements will be covered in this section, using the down-converting DUT shown in Figure 4.

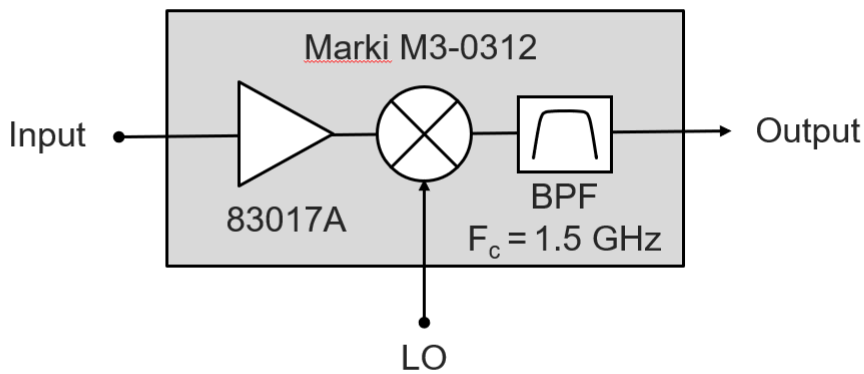


Figure 4. Block diagram of the down-converting DUT used for all example measurements unless otherwise noted.

Swept-IF example

In a typical frequency converter, the operating frequency ranges of the ports are quite different. The RF port (the input port of a down-converter or the output port of an up-converter) generally covers a much broader frequency range than the IF port, which is typically band-limited by a filter. Swept-IF measurements primarily show the IF-bandwidth response versus frequency and are commonly achieved in single-stage converters by sweeping the input port while the LO signal is fixed at a single frequency (as it often is during operation) and measuring the swept response at the output. Figure 5 shows the user-specified mixing plan and test results of such a swept-IF measurement of the example down-converter. From this measurement, which is a combination of the converter’s input and output frequency responses, narrowband gain and gain flatness can also be observed, as well as absolute output power.

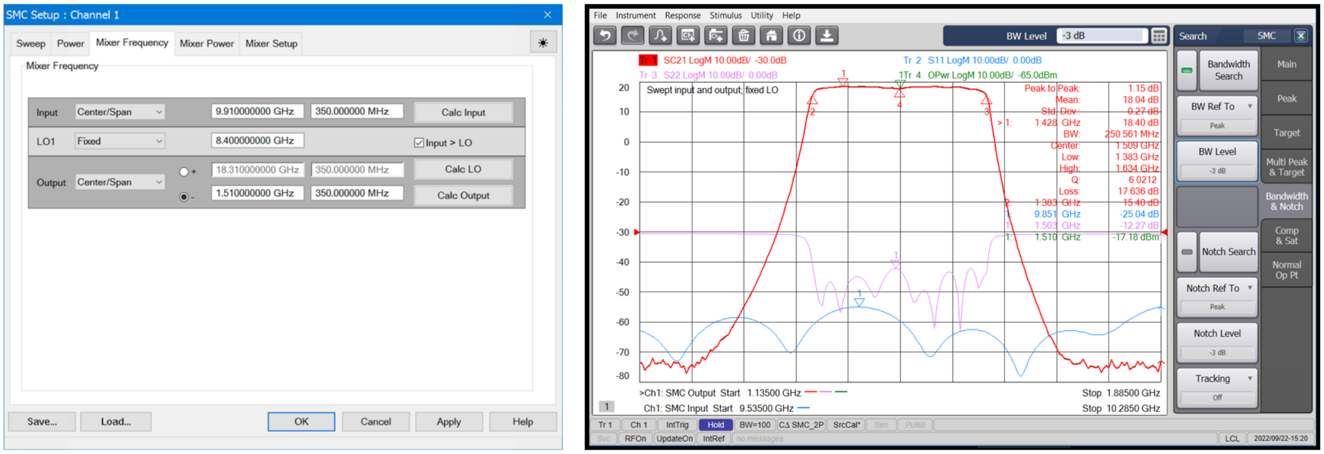


Figure 5. Mixing plan and transmission and reflection results from a swept-IF measurement of a down converter, with bandwidth and statistical data.

To isolate just the input or output response, swept LO measurements can be configured as shown in Figure 6.

Single-Stage Converter			
Input	**LO1**	**Output**	**Response**
Swept	Fixed	Swept	Input, Output
Swept	Swept	Fixed	Input
Fixed	Swept	Swept	Output

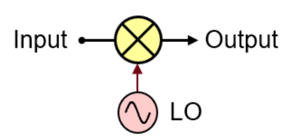


Figure 6. Single-stage converter-mixing configurations for swept-frequency responses.

Swept-LO example

Figure 7 shows the mixing plan and results of measuring the same down-converting DUT as above, but with simultaneous sweeps of the RF input and LO signal provided to the DUT. With this setup, the output frequency is fixed, and in this example, positioned in the center of the IF filter response. The results show the combined frequency response of the DUT's input amplifier and mixer, unaffected by frequency response of the IF filter. Broadband gain and gain flatness are observed with this measurement.

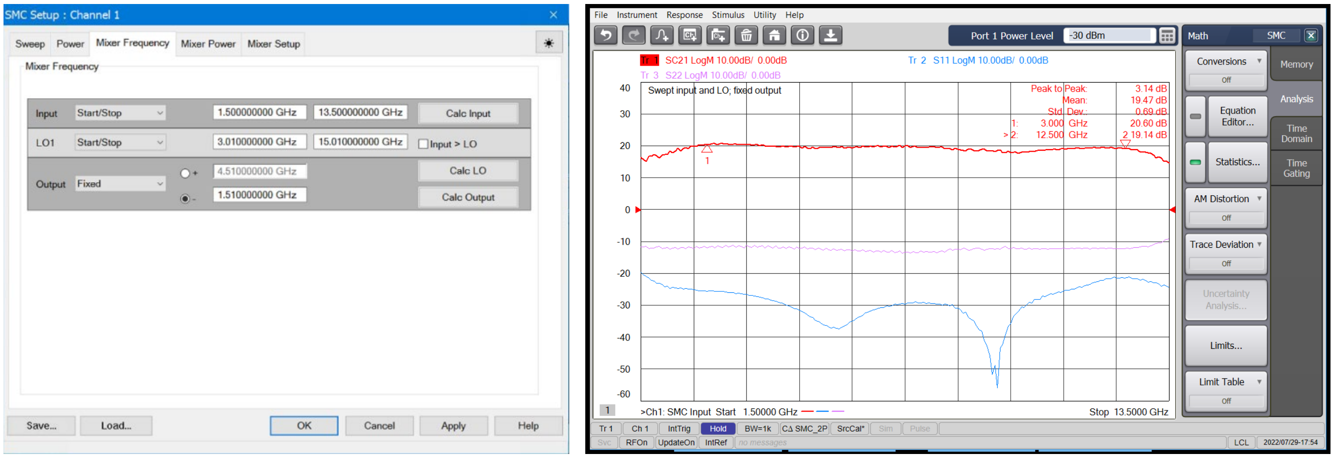


Figure 7. Mixing plan and transmission and reflection results of a swept-LO frequency sweep of a down-converter, with statistical data.

Dual-stage converters

SMC+Phase can easily configure single- and dual-stage converter setups. Dual-stage converters have two mixing stages and typically two IF filters – one at the input (for up-converters) or output (for down-converters), and a second filter in-between the two mixing stages. Figure 8 shows the user interface for configuring a dual-stage setup where both LOs are supplied externally.

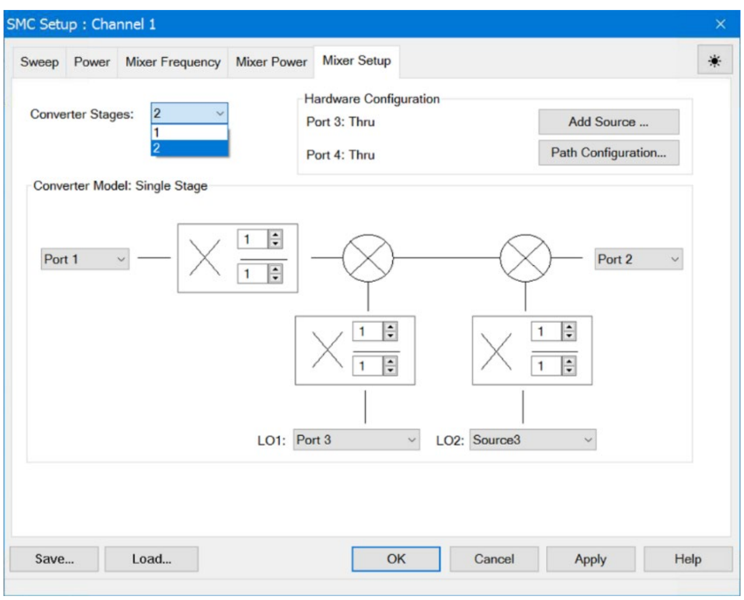


Figure 8. User interface for configuring a dual-stage converter setup.

Figure 9 shows which converter sections are measured as the result of various combinations of fixed and swept input and LO signals.

Dual-Stage Converter					
Input	**LO1**	**IF**	**LO2**	**Output**	**Response**
Swept	Fixed	Swept	Fixed	Swept	Input, IF, Output
Swept	Swept	Fixed	Fixed	Fixed	Input
Swept	Fixed	Swept	Swept	Fixed	Input, IF
Fixed	Swept	Swept	Fixed	Swept	IF, Output
Fixed	Swept	Swept	Swept	Fixed	IF
Fixed	Fixed	Fixed	Swept	Swept	Output

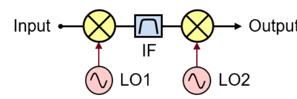


Figure 9. Dual-stage converter mixing configurations for swept-frequency responses.

Measuring converters with more than two mixing stages

SMC+Phase can be used for converters with more than two stages of conversion, with the limitation that only one LO can be controlled. While the user interface only shows one- and two-stage mixing plans, as long as the input, controlled LO, and output frequencies along with the direction of sweeps is consistent with the actual DUT's mixing plan, the internal frequencies in the user interface of each conversion stage are not important. Creating the proper mixing plan in the user interface is achieved by calculating the frequency of the uncontrolled LO that is a mathematical combination of the actual DUT LOs that are not defined or controlled. Care must be taken to select the proper high-side or low-side mixing condition so that when the PNA sweeps up, the receivers sweep in a direction that matches the output of the DUT. Figure 10 shows the actual DUT mixing plan and the calculated two-stage mixing plan for an example three-stage converter with one controlled LO at 9 GHz ($LO1_{DUT}$). LO2 in the PNA's mixing-plan is equal to $LO2_{DUT}$ – $LO3_{DUT}$ (5 GHz – 1.1 GHz = 3.9 GHz). As can be seen, the input and output frequencies of the three-stage DUT and the PNA-X's two-stage mixing plan are the same.

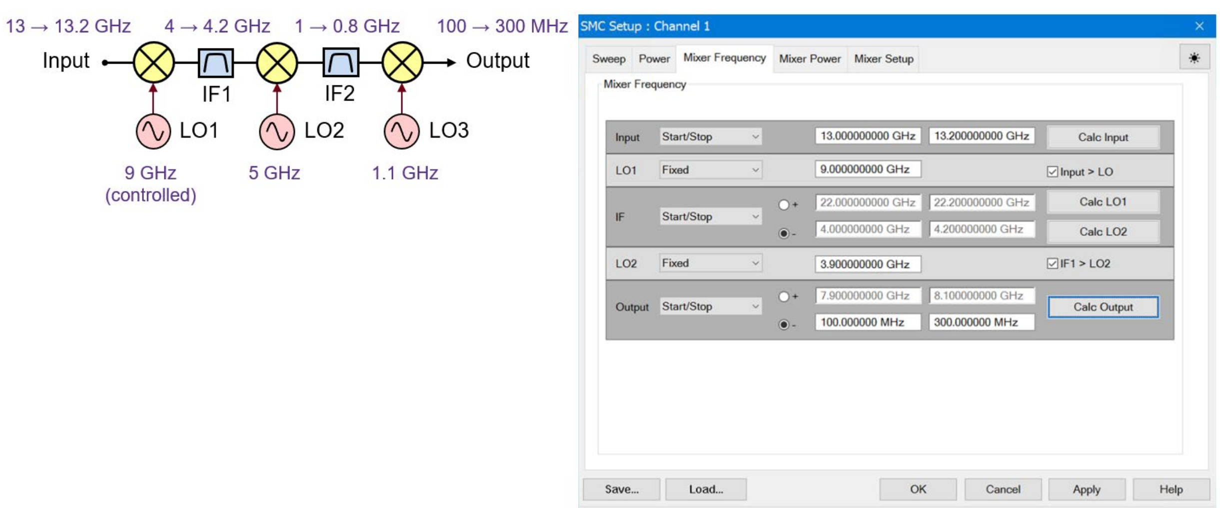


Figure 10. Actual and calculated mixing plans of an example three-stage converter with one controlled LO.

Converters with internal multipliers or dividers

Some converters have built-in multipliers or dividers in the main or LO paths, which become part of the overall mixing plan. The example in Figure 11 shows how an LO multiplier can be taken into account.

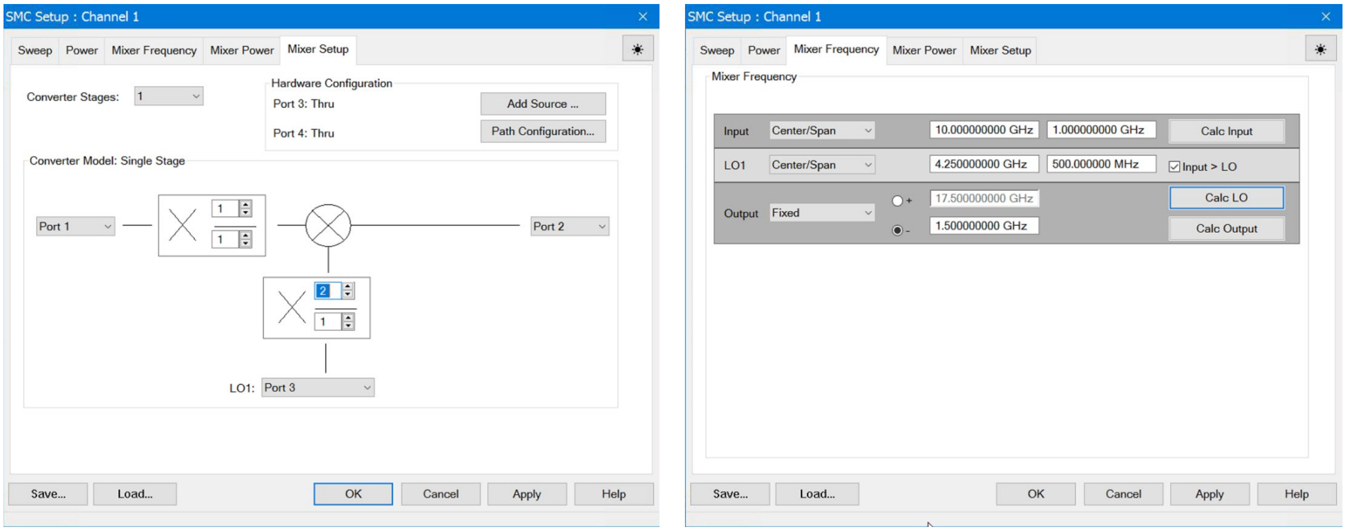


Figure 11. Example setup and mixing plan of a single-stage converter with an internal LO doubler.

Frequency multipliers

The ability to enter multipliers in the user interface can also be used to test frequency multipliers. With a single-stage mixing plan, setting the numerator at Port 1 to the desired harmonic number N and setting LO1 to zero results in an output frequency that is N times the input frequency.

Phase and delay measurements

To enable measurements of phase and group delay, the checkbox for Enable Phase on the Sweep tab of the SMC Setup dialog box must be checked. For instruments with fractional-N-based sources (synthesizer version 6), the starting phase of the sources cannot be controlled so phase traces are normalized to zero at one point in the phase response (user selectable). This eliminates the effect of a random starting phase for each sweep, giving stable, normalized phase traces. For this normalization, the user should select a measurement point that has good SNR, typically in the middle of the band of interest. Phase normalization does not impact deviation-from-linear-phase measurements, nor does it impact group delay measurements, since absolute phase is not a factor when calculating the slope of the phase-versus-frequency response. However, for these instruments, phase normalization means it is not possible to use SMC+Phase to compare phase offsets between multiple paths, multiple DUTs, or phase changes within a DUT. Figure 12 shows fixed-LO measurements of conversion gain and group delay through the example one-stage converter. The trace-analysis feature has been used to show the delay statistics over a 200 MHz span centered in the middle of the passband.

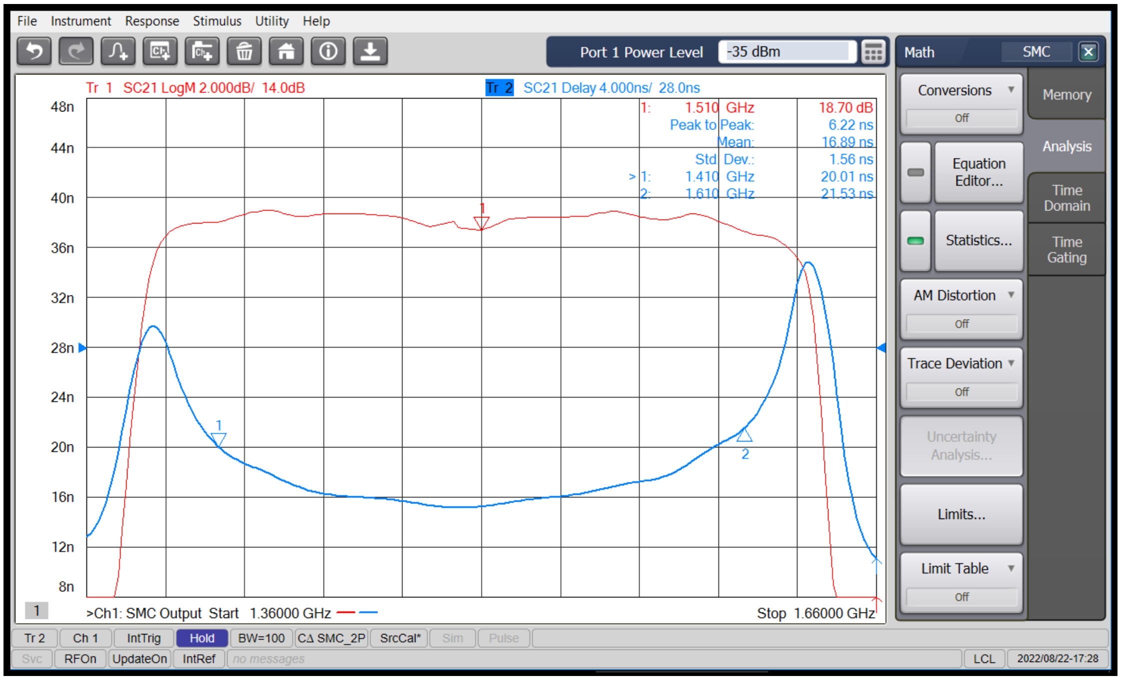


Figure 12. Fixed-LO measurements of conversion gain and group delay through the example one-stage converter.

Absolute phase

For instruments with direct-digital-synthesis (DDS)-based sources (synthesizer version 7), an additional choice is available called “Use Absolute Phase”. With this choice, the phase of the internal sources and internal LO start up with a repeatable phase at every point, giving stable and repeatable phase measurements sweep-to-sweep, without the need for phase normalization. This feature allows SMC+Phase to make measurements that previously required the Vector Mixer/Converter (VMC) measurement class and the associated reference and calibration mixers. One example is measuring phase changes that occur within a device, for example, due to an internal phase shifter in a transmit/receive module. Another example is measuring the absolute phase variation among various frequency converters or mixers due to internal path-length differences. Since the phase of the LO contributes to the converters absolute output phase, any difference in LO path lengths among multiple test stations will influence the test results. This issue can be overcome using the proper phase calibration method, described below in the calibration section.

The “Use Absolute Phase” choice can only be selected when LO1 (and LO2 if used) is set to an internal source that isn’t being used for the input stimulus. The LO can come from a test port or from the Option XSB rear-panel source (Source 3). The “Use Absolute Phase” choice is grayed out if either LO is set to “Not controlled” or to an external signal generator. Absolute phase does not work for converters with embedded LOs, since the phase of the DUT’S LO would not be synchronous with the sources within the PNA-X, even if they share a common frequency reference.

Deviation from linear phase

When measuring a device's phase response, it is generally not useful to display wrapped phase without electrical-delay compensation, as shown in Tr 1 of Figure 13. Measurements of deviation-from-linear phase are accomplished by removing the linear, negative slope from the phase-versus-frequency response, leaving the phase deviations. This process can be accomplished in one of two ways. The traditional approach is to place a marker in the center of the response and use the Marker → Delay feature, the results of which are shown in Tr 2. The firmware calculates an electrical delay value that corresponds to the linear portion of the phase response, and then mathematically removes that amount of electrical delay from the measured results. This effectively flattens the phase trace around the active marker. With this approach, the average response is not centered around zero, so the user must either calculate the phase-deviation values with delta markers, with use of the trace-statistics feature (with a user-defined span if desired), or with external software. The second approach is shown for Tr 3, where the Trace Deviation feature is used with the Linear choice. With this approach, the response is centered around zero degrees, so the phase deviation values can be directly read from the trace. Other deviation choices are Parabolic and Cubic.

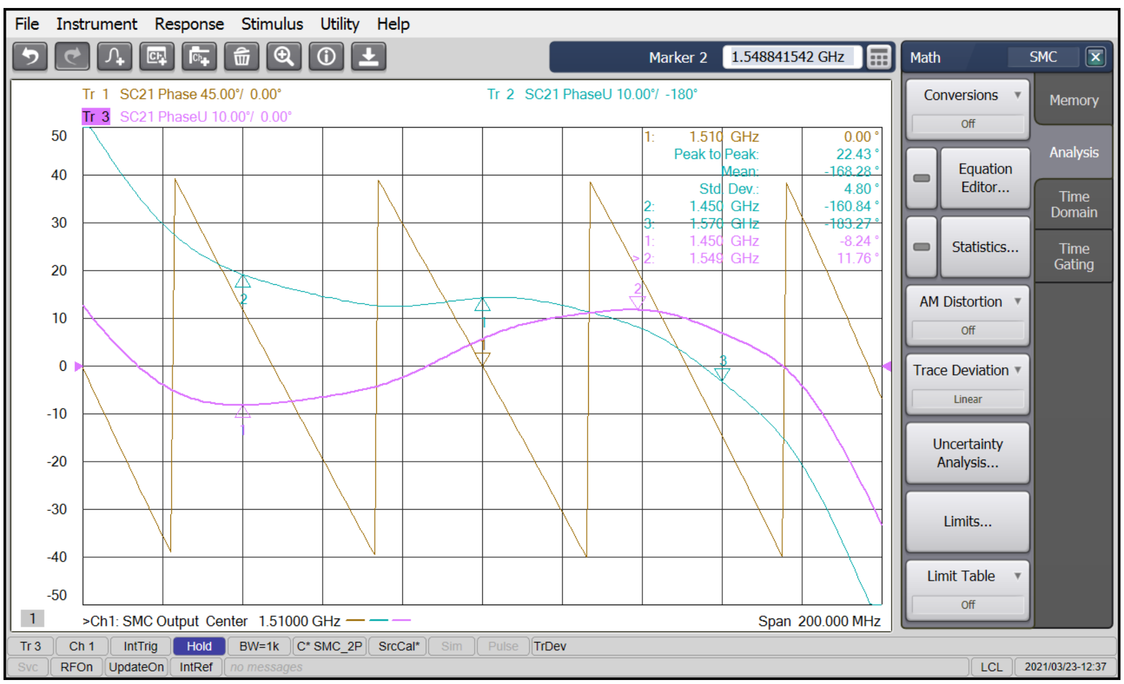


Figure 13. Phase measurements without (Tr 1) and with (Tr 2, Tr 3) electrical-delay compensation to remove the linear portion of the phase response to show deviation-from-linear phase.

Figure 14 shows the results of measuring deviation-from-linear phase of a converter with a 4.4 GHz LO and two different LO-path lengths, created in this example by adding an adapter with 94 ps of delay, shown in the bottom plot (Ch3). Both Ch1 and Ch2 have a fixed reference marker set for the case of the shorter LO path. In Ch1, the phase is normalized to the middle point, and after the adapter is added to create a longer LO path, the delta marker shows that the difference in phase between the two LO-path conditions cannot be measured. In Ch2, the “Use Absolute Phase” method was chosen, and in this case, the phase shift due to the additional LO-path length is directly measured with the delta marker as 151.8 degrees. The additional LO phase due to the adapter can be calculated as 94 ps x 4.4 GHz x 360 degrees = 149 degrees. This is very close to the measured value, with the small difference likely due to small match changes between the two LO-path conditions.

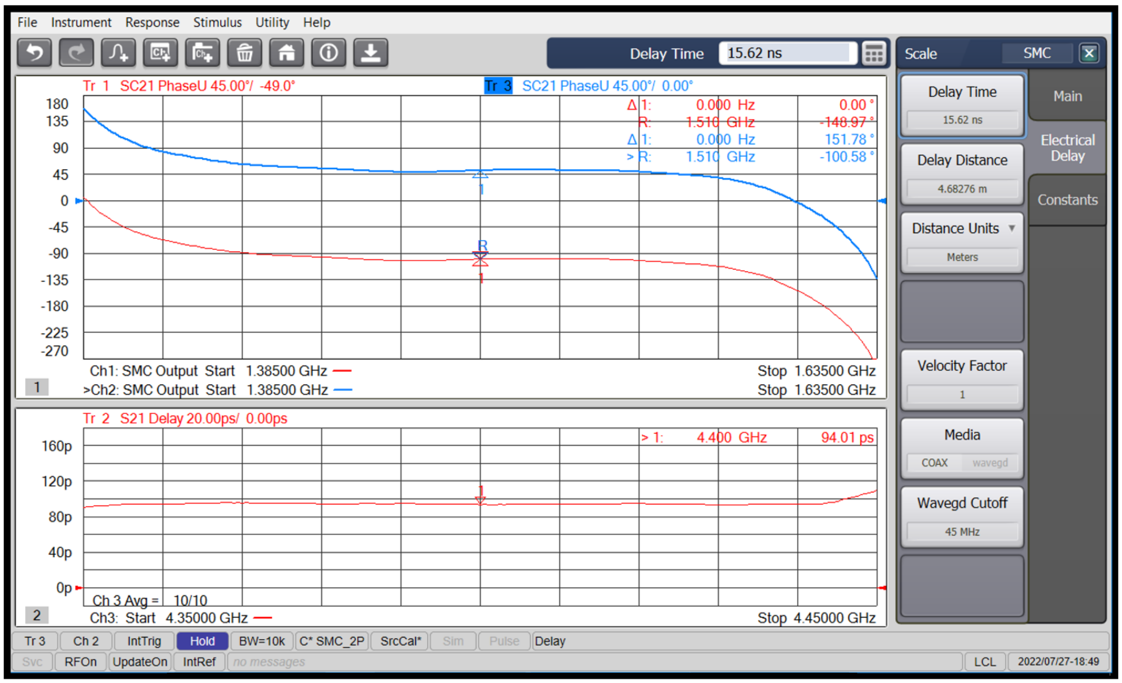


Figure 14. Selecting “Use Absolute Phase” allows measurements of phase shifts resulting from differences of LO-path lengths.

Power sweeps for gain and phase compression

Power sweeps at a fixed frequency are very useful for characterizing non-linear behavior such as gain compression and phase compression (also called phase versus drive). Power sweeps are also used for the related measurements of AM-to-AM and AM-to-PM conversion, where the slope of the gain and phase plots versus input power are calculated. Gain compression is one of the most common figures of merit for active devices and is generally considered as the approximate point where a device transitions from linear to nonlinear behavior. It is defined as the input or output power where the device’s gain drops by x dB (compared to the linear or small-signal gain), where x is defined by the user. Typically, x is 1 dB, and the compression point is commonly known as P1dB. Gain compression is easily measured on a frequency converter by performing a power sweep at the input of the device while measuring conversion gain (SC21). This can be repeated with different mixing plans, for example, with different input and LO frequencies, or a fixed input frequency and different LO frequencies. The PNA-X also has a gain-compression application that automates the frequency and power sweeps to conveniently display gain (and phase) compression versus frequency, which will be discussed in a later section. Also, it can be very beneficial to sweep the power of an LO signal, a topic that will be covered in the section on intermodulation distortion.

In Figure 15, Tr 1 shows a measurement of gain compression, where the input was swept from -40 dBm to -20 dBm. A compression marker (a marker-search function) shows the gain and input and output power associated with the 1 dB gain compression point. Using the AM Distortion feature (AM-AM), Tr 2 shows the compression response normalized to zero at the start of the power sweep, which is assumed to be in the DUT's linear region of operation. By enabling the checkbox for "Y-axis = Slope Calculated Over Aperture" in the AM Distortion feature, the derivative of the gain compression trace is directly displayed (Tr 3), which shows the dB change in gain per dB change in input power. At the 1 dB compression point, the slope of the compression curve is -0.48 dB/dB.

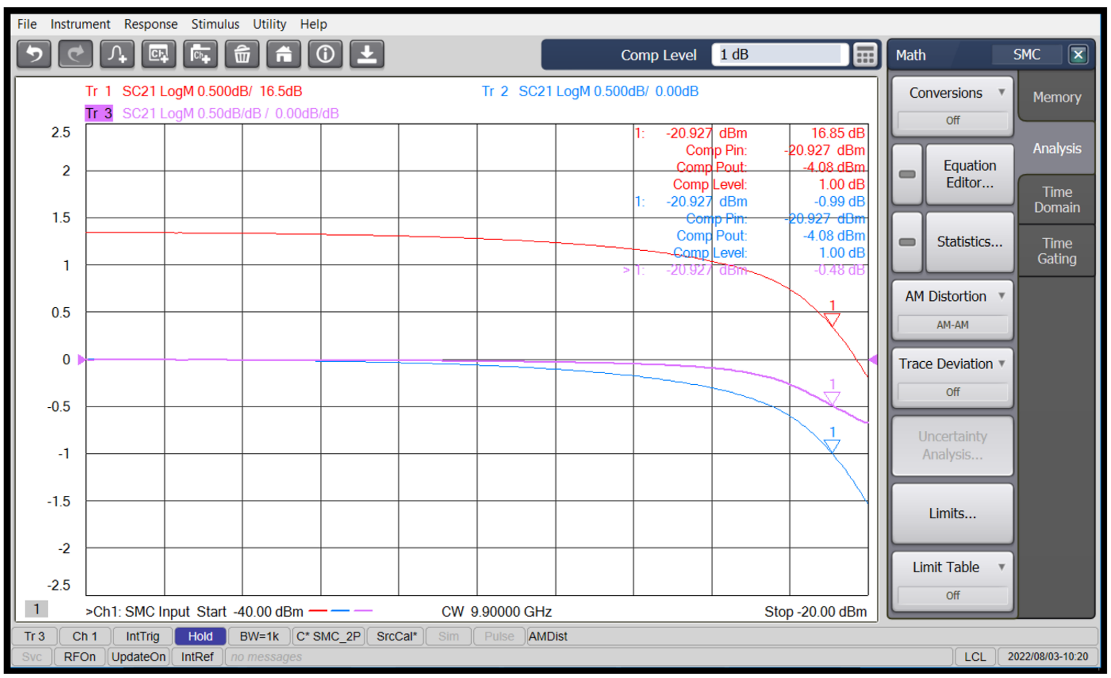


Figure 15. Gain-compression and AM-to-AM-conversion measurements using a power sweep.

Figure 16 shows a similar plot, but with the traces formatted as phase. Tr 1 shows phase versus drive, and Tr 2 is the same response but with the AM Distortion feature (AM-PM) enabled, which normalizes the phase at the start of the power sweep to zero. The phase compression marker on this trace shows that 1 degree of phase compression occurs with considerably lower power than the 1 dB gain compression point. With the the Y-axis aperature checkbox enabled (Tr 3), the phase change per dB change in input power at the 1 degree compression point is shown as -0.33 degree/dB.

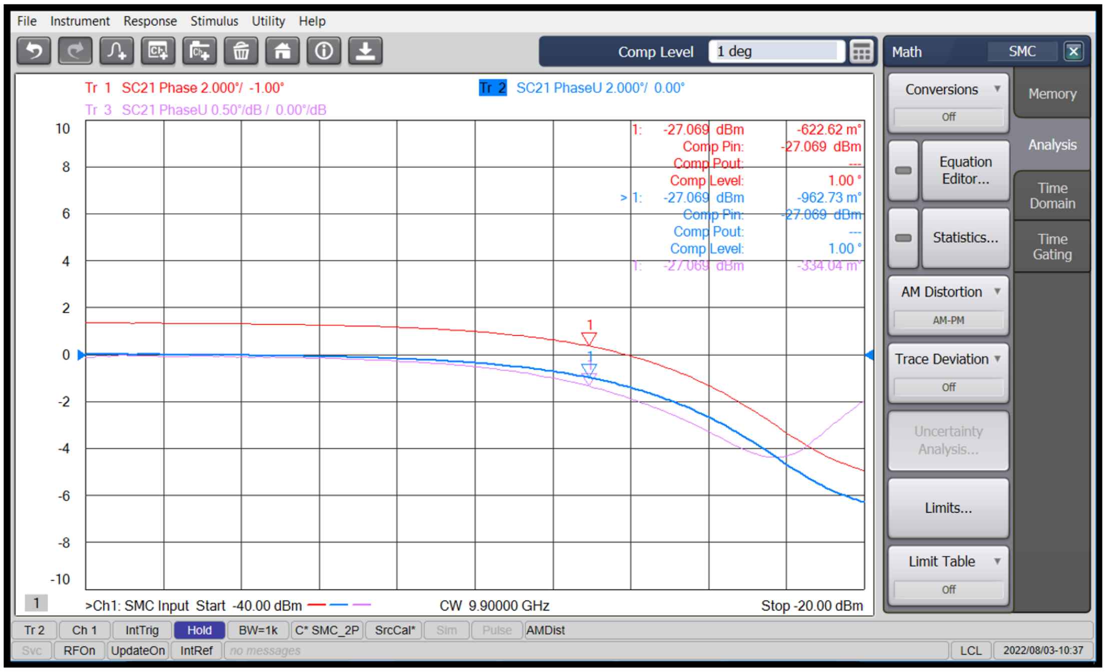


Figure 16. Phase-versus-drive and AM-to-PM-conversion measurements using a power sweep.

Phase transfer

A related measurement to phase compression or phase versus drive is a measurement often called phase transfer. In this case, a large out-of-band signal is swept in power while the phase response of a smaller in-band signal is measured. This measurement must be done outside of SMC+Phase, using the frequency-offset mode (FOM) of a standard channel.

The setup takes advantage of the PNA-X's built-in signal combiner which routes source 2 through the combiner and out port 1, as shown in the Path Configuration dialog in Figure 17. In the FOM setup dialog shown in Figure 17, the main source supplies the swept-power out-of-band signal at 6.41 GHz, while the second source supplies a constant-power in-band signal at 5.91 GHz. The receivers are tuned to the in-band output frequency (1.51 GHz), and source 3 (available with Option XSB) provides the 4.4 GHz DUT LO from the rear panel to port 3. Alternately, an external signal generator could be used for the LO signal. The Power and Attenuators dialog is used to configure the power levels. Here, Port 1 (the out-of-band signal) is swept from -50 dBm to -25 dBm, while Port 1 Src2 (the in-band signal) is kept at a constant -35 dBm. The LO power (Source3) is set to 18 dBm which overcomes some system loss.

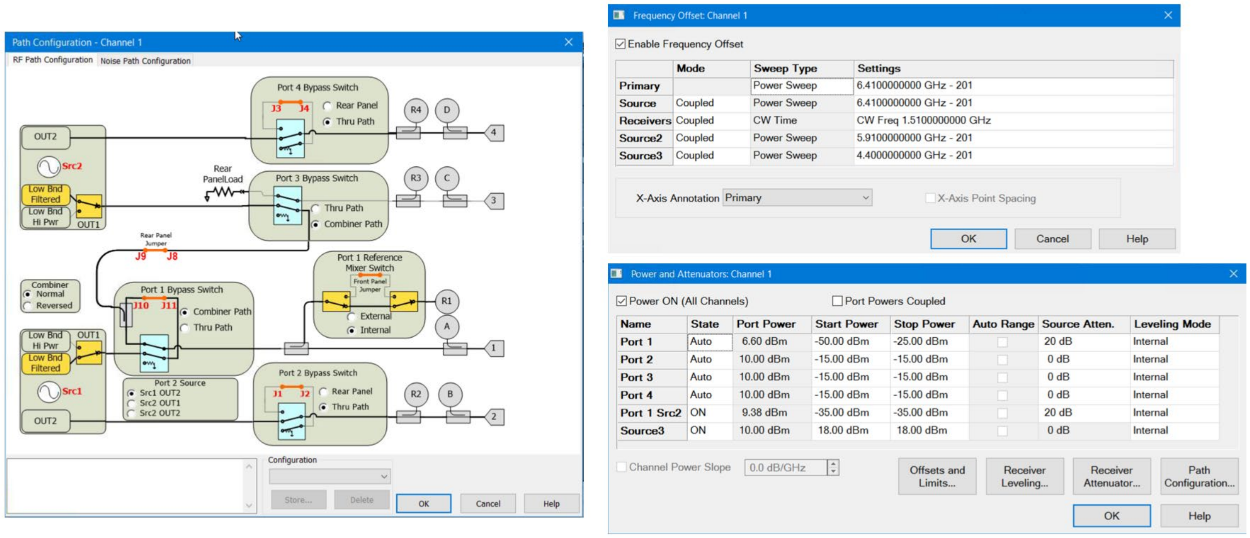


Figure 17. Setup dialog boxes for phase-transfer measurements using frequency-offset mode and the internal signal combiner.

The results of the measurement are shown in Figure 18. The measurement is done with a single receiver (b2 in this case) measuring ouput power and relative phase. Tr 1 shows gain compression of the constant-power in-band signal as a result of sweeping the out-of-band signal, and Tr 2 shows the phase compression or phase transfer.

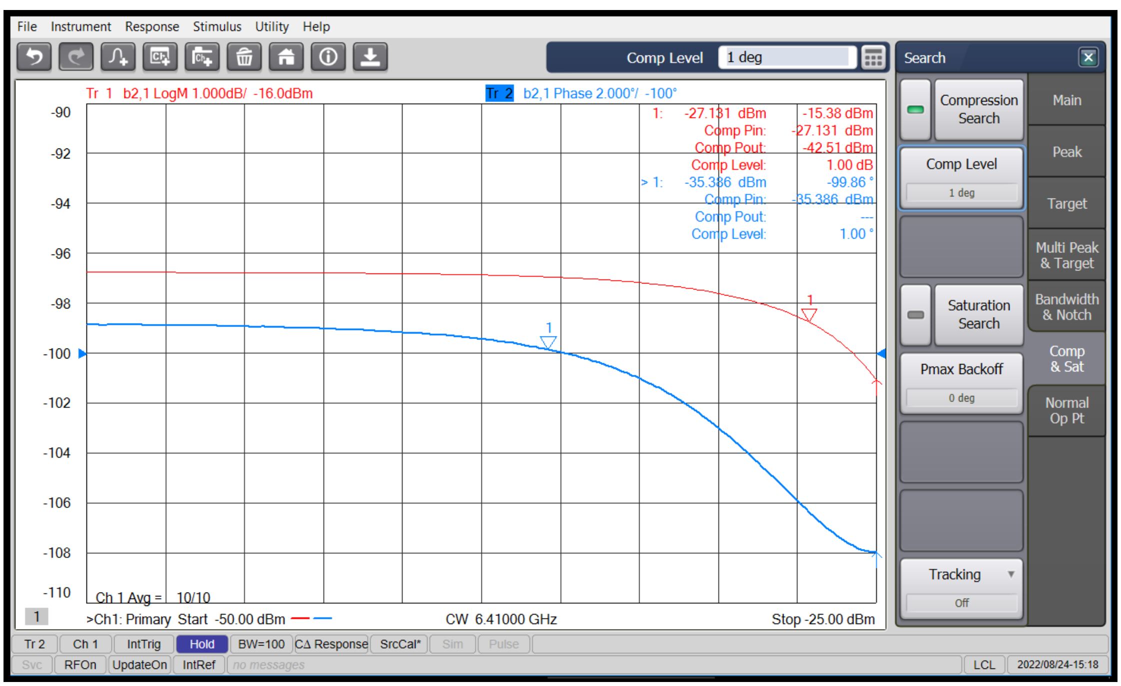


Figure 18. Example phase-transfer measurement resulting from a power sweep of an out-of-band signal.

Because calibrating a standard channel cannot be done while using FOM (it is automatically turned off during the calibration), it is convenient to create a second channel for the calibration that includes all the frequencies used in the phase-transfer measurement. The easist way to accomplish this is to set up the second channel in segment-sweep mode, where three single-point segments are defined corresponding to the input in-band frequency, the input out-of-band frequency, and the in-band output frequency. Next, Cal All should be used to calibrate the second channel, and Enable Extra Power Cals should be configured to calibrate the output power of Port 1 Src2, and if desired, the LO power. Upon completing the calibration, chose to Save As User Calset. Next, apply the saved calset to the phase-transfer measurement channel using the Cal Set Selection dialog – when selecting the calset, use the “Do not change the active channel’s stimulus settings” choice, which will preserve the measurement settings and use interpolation as needed. Because of the way error coefficients are calculated, averaging must be used for the phase-transfer measurement when calibration is applied. The number of averages can be set to one if measurement speed is critical.

Segment sweeps

Characterizing group delay versus LO frequency of broadband frequency converters that have narrowband IF filters is difficult because the IF filter prevents broadband frequency sweeps and swept-LO/fixed-IF phase measurements of delay are not possible with SMC+Phase. One way around this issue is to use segment sweeps, in which each segment has a frequency span consistent with the IF response, but different RF and LO frequencies that can cover the full range of the DUT. An example of using segment sweeps is shown in Figure 19, where the segment spans are 315 MHz, and the LO is stepped in frequency from 4 GHz to 9 GHz in 500 MHz steps using 11 segments. In the top window, delay and magnitude traces are shown, where the segments are all plotted over the common output-frequency span of the down-converter. In the bottom plot, a mode called “X-Axis Point Spacing” is used, where the data in each segment is concatenated instead of overlaid, and the x-axis units is points instead of frequency.

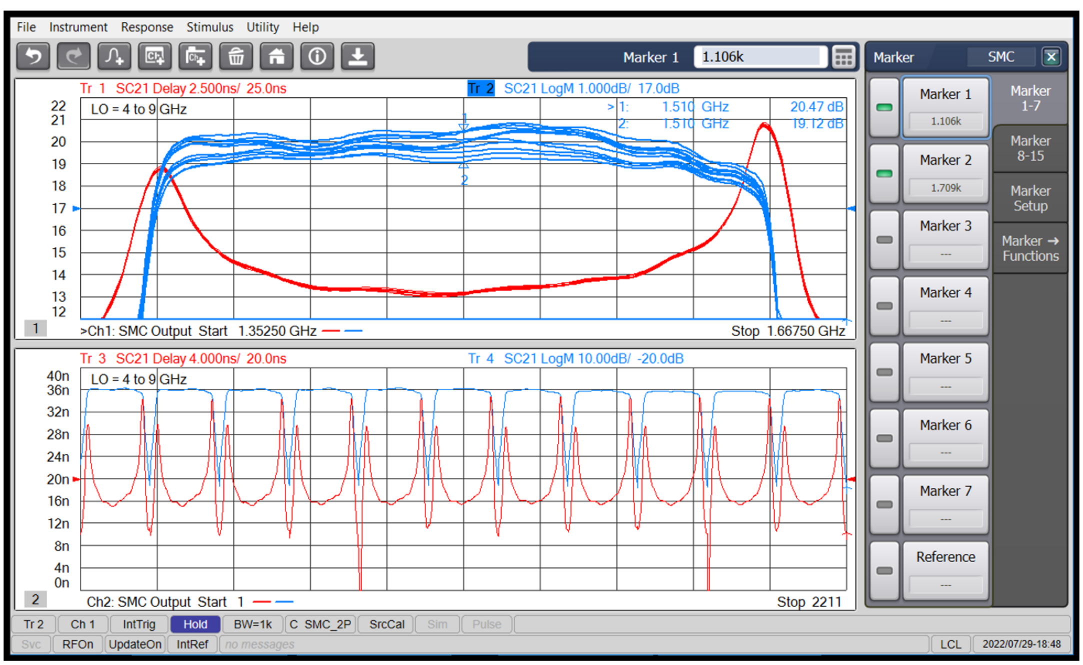


Figure 19. Segment sweeps of conversion gain and group delay, showing overlay mode (top) and X-axis point spacing (bottom).

Calibration

For magnitude measurements, calibration consists of normal S-parameter steps combined with a power-meter step where the absolute power of the source driving the DUT's input port is measured. The absolute-power measurement in conjunction with the S-parameter error coefficients are used to transfer the power calibration to the measurement receivers. This combination allows measurements of vector-error-corrected input and output match, as well as match-corrected power measurements of input and output power, which when combined, is used to calculate forward and reverse conversion gain or loss.

When phase and delay measurements are desired, and the "Enable Phase Correction" checkbox is enabled, three phase-correction choices are presented, as shown in Figure 20.

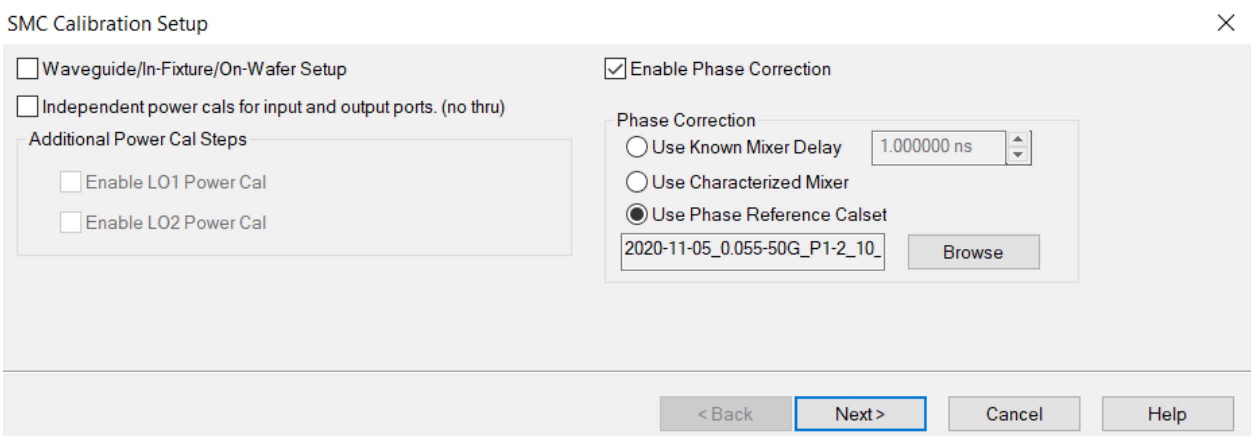


Figure 20. Phase-correction calibration choices.

The first is "Use Known Mixer Delay", a method where a calibration mixer is used as a through-calibration standard with a user-specified value of electrical delay, typically on the order of a few nanoseconds. The second choice is "Use Characterized Mixer", where data is loaded for a reciprocal-mixer/filter pair that has been previously characterized using reflection measurements with different impedance terminations (usually done with an ECal module or with mechanical open, short, and load standards). A calibration mixer (either with the first or second choice) should be used to get consistent absolute-phase measurements between multiple test stations, where the test-station LO cable lengths might be different. In this scenario, the same calibration mixer must be used among the test stations. The absolute phase of the DUT will be relative to the calibration mixer and the results will be the same as those obtained using a VMC channel (described below), but without the need for a reference mixer.

The third phase-calibration choice is "Use Phase Reference Calset". This choice is suitable for all phase and delay measurements where calibrated absolute phase of the LO is not needed, for example, for group delay and deviation-from-linear-phase measurements, or for measuring phase changes in a converter with a phase shifter. This choice uses a two-tier calibration approach, where the first tier is referenced to the test-port connectors of the PNA-X itself. This first tier provides all the receivers with a magnitude and phase calibration over the full frequency range of the instrument, as well as providing power-calibrated sources at each test port. This calibration is done ahead of the DUT measurement, using the Phase Reference Wizard. The second-tier calibration, which is typically done over a subset of the first-tier calibration frequencies, is performed at the desired reference plane and typically only requires

S-parameter calibration standards using an ECal module or a mechanical calibration kit. The second-tier calibration removes the effects of cables, adapters, switches, and any other components required to connect to the DUT.

A big advantage of using the phase-reference approach is that a calibration mixer is not required. Instead, the first-tier calibration is done with a set of broadband calibration standards: an S-parameter calibration kit, a power sensor, and a Keysight U9391C/F/G comb-generator driven from the 10 MHz reference-out signal of the PNA-X. The comb generator provides a set of phase-aligned spectral lines spaced 10 MHz apart that are used to characterize the broadband phase-versus frequency response of the receivers. The power meter is used for a broadband receiver-power calibration. The power meter and comb generator only need to be connected to one port each, and the S-parameter correction terms are used to extend the power and phase calibration planes to the other ports and receivers. This first-tier cal is saved as a user calset, and because the PNA-X hardware is very stable (assuming a relatively stable temperature in the operating environment), it does not need to be repeated on a frequent basis. The time between phase-reference characterizations is left to the user to determine, but the interval is typically many months.

Measuring devices with embedded-LOs

Special considerations are required when measuring phase and group delay on frequency converters with embedded (internal) LOs that are inaccessible and without means to provide or receive a frequency reference to ensure frequency synchronization between the DUT and the PNA-X. This situation is common with satellite transponders, where size, weight, and power limitations eliminate easy access to the local oscillators aboard the satellite. Historically, this challenge meant that VNAs were not used for transponder phase and delay measurements, dramatically hindering efforts to improve the speed of transponder characterization. With a modern instrument like the PNA-X with advanced measurement applications such as the one described in this section, the VNA is now the main tool for transponder characterization.

SMC+Phase requires tuning the PNA-X receivers to frequencies that precisely match the output frequencies of the DUT. However, without a way to establish frequency locking between the DUT and the PNA-X, the measurement receivers will not be tuned to the actual output frequencies of the DUT unless extra measurements are taken to characterize and compensate for LO-frequency offsets. The frequency offset for most embedded-LO converters ranges from a few kHz to several tens of kHz, but any offset is enough to cause rapid changes in the measured output phase, obscuring the actual response of the DUT and making it impossible to correctly measure group delay. The solution for this class of devices is the S93084B Embedded-LO Application. This software application performs background sweeps to measure the DUT's output-frequency offset compared to the nominal mixing plan (within a defined frequency tolerance). The measured offset is then added to the nominal LO value in the mixing plan. This modifies the output frequency of the mixing plan, which then shifts the tuning of the PNA-X receivers to the correct DUT output frequency. From then on, the SMC+Phase measurements are performed as described above. The embedded-LO feature works on DUTs with multiple mixing stages, as described in the previous section "Measuring converters with more than two stages". It also works in other converter-measurement classes such as gain compression, intermodulation distortion, and noise figure.

LO-offset determination

The algorithm used to determine the frequency offset of the DUT's internal LO(s) is split in to two parts. For the first part, a CW signal within the measurement span (default is center) is applied to the input of the DUT, and the DUT output is measured with a broadband receiver sweep over a user-selectable frequency span (default is 3 MHz), centered at the nominal output frequency based on the mixing plan. A peak search is done to find the frequency of the output signal, and the difference between the peak and nominal output frequency is used as the coarse value of the LO offset. This is a fast process but does not provide the necessary frequency resolution for a successful SMC+Phase phase or delay measurement.

The second part of the LO-offset determination is called the precise sweep. The same CW signal is applied to the DUT input, but instead of performing a broadband frequency sweep, the receiver measuring the DUT output is tuned to the output frequency determined from the broadband sweep, and a phase-versus time sweep is taken. The slope of the trace is calculated to determine the fine LO-frequency offset which is then combined with the coarse value to create a new estimate of the overall LO-frequency offset and a new estimate of the output frequency. The precise sweep is repeated as necessary until either the estimated output frequency is within a defined tolerance value (default is 1 Hz), or the maximum number of iterations is reached. The combination of coarse and precise sweeps to determine the actual LO value is fast while providing the necessary frequency resolution for successful phase and delay measurements. The mixing-plan graphical-user interface always shows the nominal frequency values, but the measured LO-offset frequency is displayed in the embedded-LO setup dialog box and on the lower left of the display where is updated by default for each sweep (Figure 21).

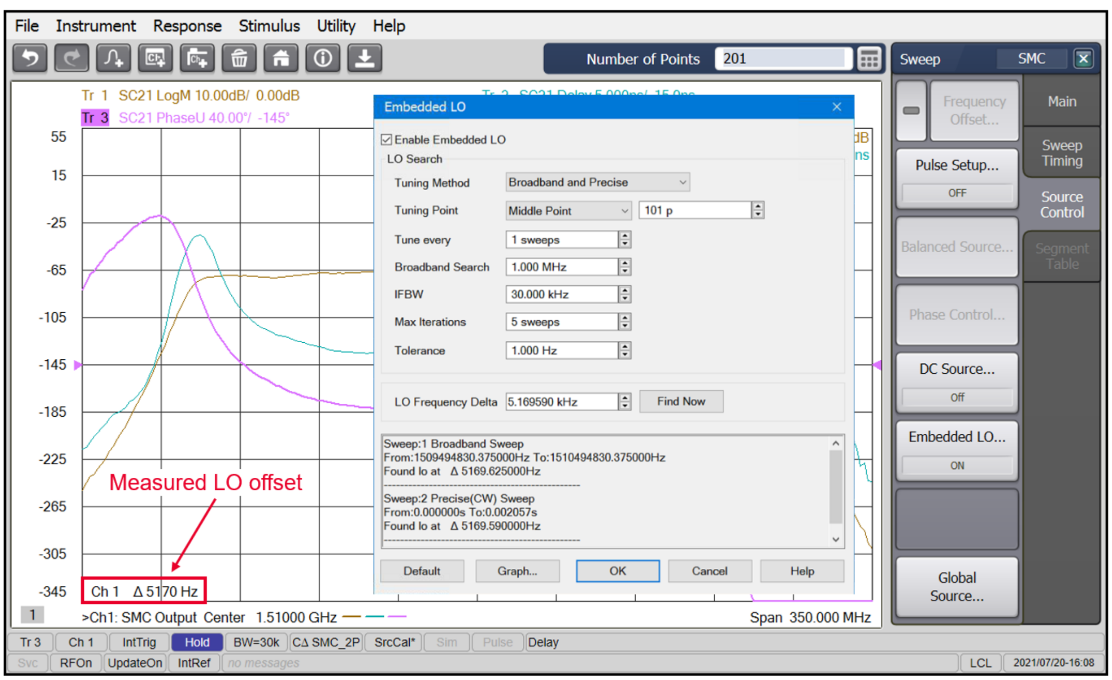


Figure 21. The embedded-LO frequency-offset value is shown in the setup dialog box and displayed on the lower-left of the display.

LO stability

For high-accuracy measurements, the phase noise of both the PNA-X and the LOs in the DUT must be sufficiently low so that errors due to non-ratioed receiver-phase measurements are not excessive. This is not an issue for most satellite transponders, as their LOs are typically locked to crystal oscillators that are very stable and have low phase noise. The phase-noise performance of the DDS sources in the PNA-X is also quite good, which allows the use of wide-IF bandwidths for speed, with minimal or no sweep averaging. For older PNA-Xs with fractional-N-based sources, an IF bandwidth between 10 kHz and 30 kHz is recommended, with at least 10 sweep averages. Figure 22 shows the effect that LO phase noise has on group delay measurements. The top trace serves as a reference measurement using port 3 of the PNA-X for the LO signal. The other traces used an external signal generator with a phase-noise impairment feature which allows the level of the phase-noise pedestal to set (in dBc/Hz) over a defined offset range, 1 kHz to 30 kHz in this example. The added noise on the delay trace starts to become noticeable at approximately -80 dBc/Hz.

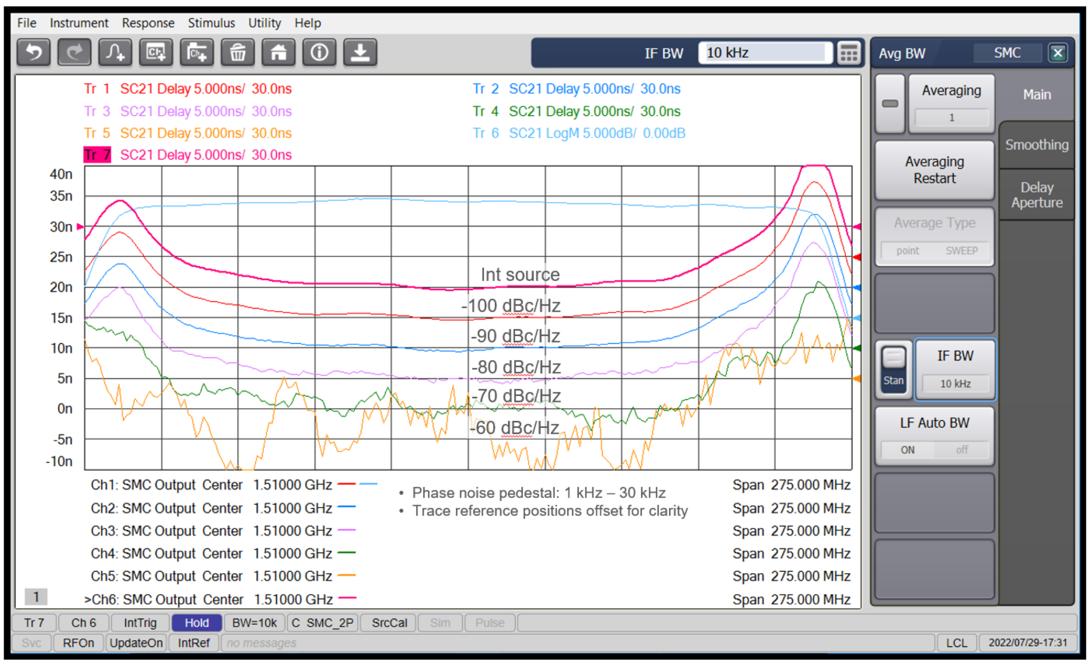


Figure 22. Effect of LO phase noise on group delay measurements.

VMC

Prior to the release of SMC+Phase, the vector mixer/converter (VMC) measurement class was the only way to measure phase and group delay of frequency converters. VMC requires both a reference mixer and a calibration mixer, both of which have been eliminated with SMC+Phase. VMC is still useful for older PNA-X instruments that have fractional-N-based sources. VMC can be used to measure absolute group delay, deviation from linear phase, and phase changes within a device, for example, from a phase-shifter in a transmit/receive module or a beam-forming integrated circuit. It can also be used for adjusting the LO-path length within a single-path converter to match a reference or “golden” device.

The setup diagram of a VMC measurement on a single-stage frequency converter using a two-port PNA-X is shown in Figure 23. VMC uses the parallel-test method, where the reference mixer is connected to the port 1 reference receiver using front-panel jumpers at port 1. The mixer is switched in during VC21 forward-transmission measurements to provide a phase reference for the DUT's output signal, and switched out for S11 reflection measurements.

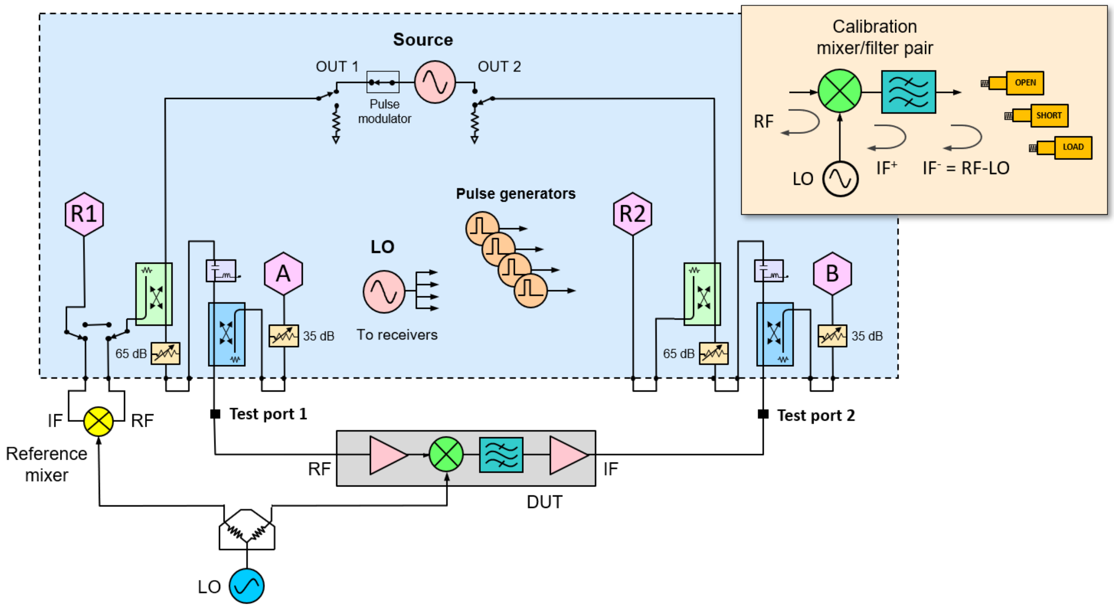


Figure 23. VMC-measurement setup for a single-stage converter and calibration standards used for characterizing the calibration mixer/filter pair.

For calibration, a characterized, reciprocal mixer that covers the DUT's frequency plan is used in conjunction with a filter that selects the proper mixing response. The mixer/filter pair is characterized with three reflection measurements using three different calibration standards connected to the mixer/filter's output as shown in the upper right of Figure 23. Either mechanical standards or an ECal module can be used. From the three reflection measurements, the input and output match of the calibration mixer/filter pair can be calculated, as well as the two-way transmission response. Since the mixer is presumed to be reciprocal (which means the conversion response is the same in both directions), the one-way response is calculated as half of the two-way response. To calibrate the test system for absolute magnitude and delay, the calibration/filter pair is used as a characterized-through standard. The absolute phase of the DUT will be relative to the calibration mixer, and as long as the same calibration mixer/filter pair is used, any differences between the LO path lengths of the reference mixer and DUT among multiple test setups will be removed.

Absolute Phase Offsets of Multipath Converters

Many frequency converters have multiple paths driven from common, internal LOs. For fixed LOs, differences in LO-path lengths adds phase offsets between the converter outputs, which must be characterized. Since SMC+Phase and VMC normalize the phase response of embedded-LO converters, the absolute phase response of individual converter paths cannot be measured. However, since the goal is characterizing the difference in phase between paths, one path can serve as the reference and the remaining paths can be measured relative to that path. Figure 24 shows an example two-path converter with embedded LOs, and the test setup for measuring the relative phase difference between paths. This setup uses the S93089B Differential and I/Q Device Measurement application to set the relative phase between the RF stimulus from ports one and three to zero degrees at the input to the converter. This phase coherency can be maintained across a frequency sweep. The phase difference at the DUT output ports is measured as a ratio of the test receivers at ports two and four, for example B/D or b2/b4. Using S93089B eliminates the need for a phase-matched splitter and equal-length cables at the input of the DUT or having to compensate measurement results due to phase differences between paths of an input splitter and associated cables.

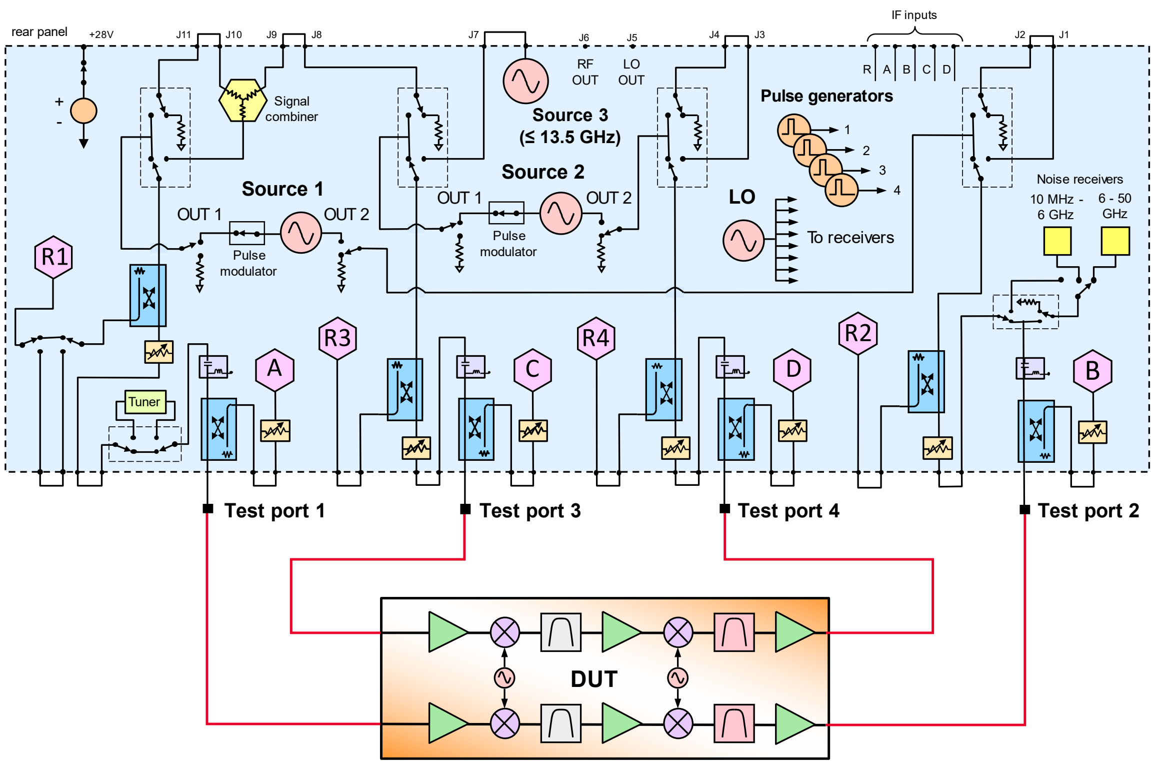


Figure 24. Test setup for measuring the relative phase difference between paths of an example two-path converter with embedded LOs.

When the LOs in the converter use a frequency reference that can be locked to the PNA-X or can provide a reference to the PNA-X, the receivers in the PNA-X will be tuned to the proper frequencies as defined in the measurement setup. However, if the embedded LOs are not locked to the PNA-X, then the output signals may fall on the skirts of the PNA-X's IF filters, causing SNR degradation. One way to mitigate this effect is to simply increase the measurement IF bandwidth so that loss is minimized. However, for large LO offsets, increasing the IF bandwidth by a large amount causes a large rise in the noise floor. A better way would be to determine the overall LO offset by other means, in order to adjust the receiver frequencies to account for the offset. One way to accomplish this is by using the embedded-LO feature in an SMC or SMC+Phase channel to find the frequency offset, and then use the offset value to adjust the DUT's output-frequency range in the Differential I/Q channel.

Isolation and Leakage Measurements

Isolation and leakage measurements characterize signals that do not undergo frequency translation through the mixer or converter. Common measurements are input-to-output isolation (for example, RF to IF for down-converters or IF to RF for up-converters), and LO leakage out the input and output ports. These measurements are especially common for mixers since there isn't any filtering to suppress the responses. Isolation and leakage measurements are typically done in the standard measurement class. Isolation measurements use transmission S-parameters like S21, which looks at the output signal at port 2 resulting from an input signal at port one, both at the same frequency. Typically, LO signals are supplied to the DUT's mixers during the test so that they are operating as they would in normal use. If the LOs are supplied by the PNA-X, then frequency-offset mode must be used to set them to the appropriate frequency, and the Powers and Attenuators dialog is used to set the appropriate power levels.

LO leakage is typically measured as absolute power. This can be accomplished by using a single (non-ratioed) receiver that has been properly calibrated. The LO is set to the appropriate sweep range and power level, and usually there is no signal applied to the DUT's input. In Figure 25, Tr 1 measures the RF to IF leakage of a down-converter where the input sweeps from 3 GHz to 12 GHz with a fixed LO at 7.50001 GHz. The LO was slightly offset in frequency to avoid LO leakage at the output falling on a measured isolation point. Tr 2 shows the leakage in dBm resulting from sweeping the LO from 3 GHz to 12 GHz. Since there are no other signals applied to the converter, there is no need for frequency-offset mode.

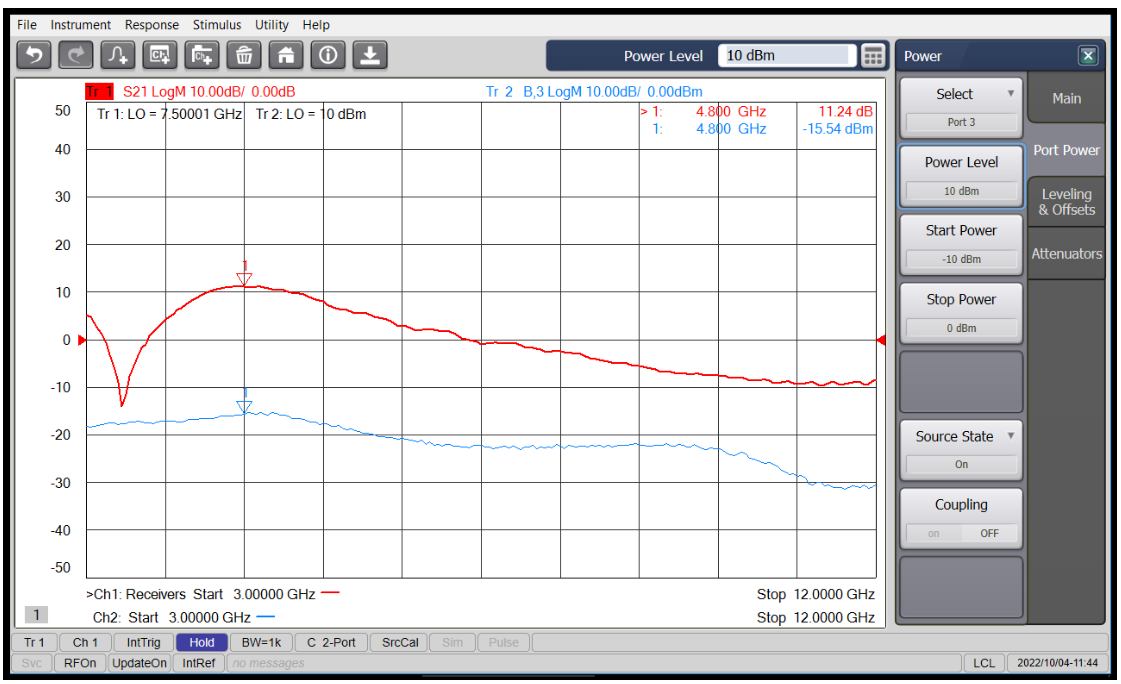


Figure 25. Measurements of RF-to-IF isolation (Tr 1) and output LO leakage (Tr 2).

Gain Compression

While VNA power sweeps provide an intuitive way to measure gain compression at a fixed frequency (as discussed earlier in the SMC+Phase section) and provide significantly faster measurements than the traditional approach of using an external signal generator and a spectrum analyzer or power meter, further speed improvements can be made when measurements of gain compression versus frequency are required. The gain-compression measurement application (S93086B, covering amplifiers and frequency converters) was developed to optimize the speed of these measurements, while providing high flexibility for measurement definition and setup.

The problem of using power sweeps with equal power steps is that a lot of unnecessary data is taken. For reasonable power resolution, the number of points in the power sweep is typically between 50 and 100. However, it only takes two data points to measure gain reduction: one reference point in the linear or small-signal region (where gain is not a function of power), and one point in the compression region corresponding to the desired drop in gain (typically 1 dB). With this approach, 96% to 98% of the data is not used for the calculation of gain compression, although we need the data to ensure that the two necessary power points are measured across the measurement frequencies and DUT-to-DUT variations are accounted for.

SMART Sweep

A smarter algorithm for setting input powers can significantly reduce the amount of unnecessary measurement data. The gain-compression application uses an algorithm called SMART Sweep that utilizes unequal power steps in an iterative loop to minimize the number of measurements required to determine the gain compression point, at each frequency of interest. In most instances, only 3 to 7 iterations are needed, producing results much faster than using normal power sweeps (Figure 26). Mismatch-corrected power measurements are used to enhance accuracy. The algorithm works like this: First, DUT gain is measured at the user-specified linear (i.e. small-signal) power level. Next, the power is increased to a value within the user-defined power range, and gain is measured again. If there is a drop in gain that exceeds the specified compression level, the power is reduced. If the gain drop is smaller than the specified value, the power is increased. This process is repeated until the desired gain drop is achieved, within a user-specified tolerance. During the iteration process, a compression curve-fitting algorithm is used to estimate what the next input-power value should be, which decreases the number of iterations compared to simpler algorithms. Once the compression point has been found, the power at compression can be displayed as either input power or output power. This process occurs for each frequency point in the setup.

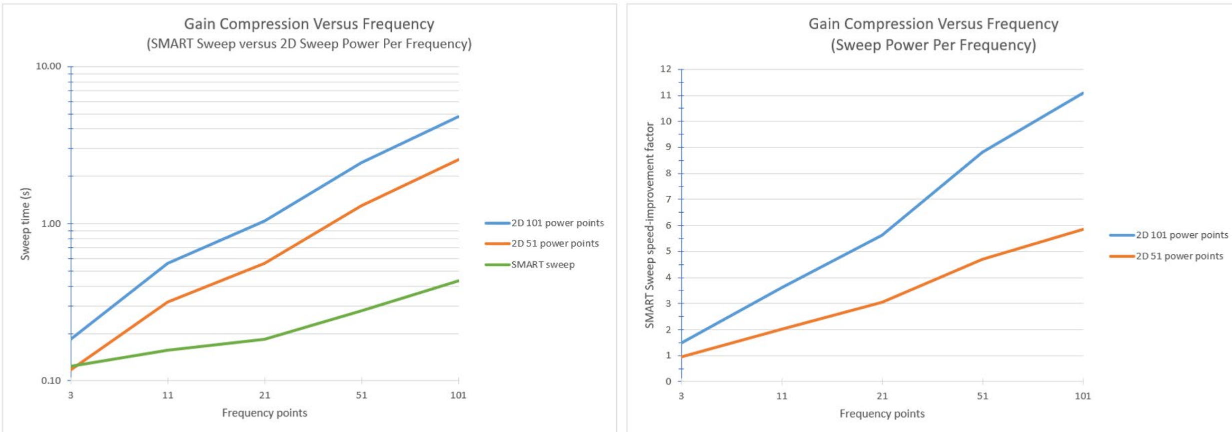


Figure 26. Test Swept-frequency gain-compression measurements using the SMART Sweep algorithm are usually much faster compared to traditional stepped-frequency power sweeps (left). The speed-up factor is shown at right.

While the power-setting algorithm used with SMART Sweep is the fasted method, at some frequencies the DUT is compressed well beyond the desired value. This overshoot may not be acceptable for high-power or other particularly sensitive DUTs than cannot tolerate power levels significantly above the compression point. An alternative “safe-mode” power-setting method is available that increases the power in a more controlled way, minimizing the power overshoot that can occur. When Safe Mode is on, the power is increased by user-specified coarse steps until a user-specified gain drop is reached. Beyond that point, the power is increased in user-specified fine steps until the target compression point is reached. This means the input power is never higher than the fine-power setting. The trade off when using Safe Mode is that more power points are needed to reach compression, which increases the overall test time.

2D sweeps and phase compression

For users that want a more comprehensive data set of power and frequency, the application also offers two “2D” sweep types: sweep power per frequency (i.e. the traditional VNA method) or sweep frequency per power. With either choice, the user can choose to check for magnitude and/or phase compression. When checking for both magnitude and phase compression, the power at compression is reported for whichever limit (dB or degree) is encountered first. The full data set of S-parameters, power, and frequency for both SMART Sweeps and 2D sweeps can be saved in comma-separated-variables (csv) format for analysis or importing into RF simulation software.

Compression methods

The gain-compression application includes several different methods for determining the DUT’s compression point (Figure 27). The most common method is Compression from Linear Gain. This method uses the user-specified linear-input power to establish the gain reference and is the best choice when using SMART Sweep. The other methods are best used with 2D sweeps. Compression from Max Gain is sometimes used for converters that exhibit gain expansion (i.e. an increase in gain) prior to compression. With this method, the maximum gain is used as the reference instead of the linear gain. This method results in a lower compression point compared to using the linear gain as the reference.

Compression from Back Off examines the gain versus input-power data and uses the user-specified back-off level to determine the gain reference. At each frequency, the algorithm searches for the first pair of data points (starting with the highest input power) where the difference in input power equal to the back-off level results in the desired amount of compression. The compression point is the higher-power value of the pair. The X/Y compression method is similar, except the algorithm examines the output-power versus input-power data. Again, pairs of data points are examined to find the first pair where the user-specified delta-X change in input power results in the user-specified delta-Y change in output power.

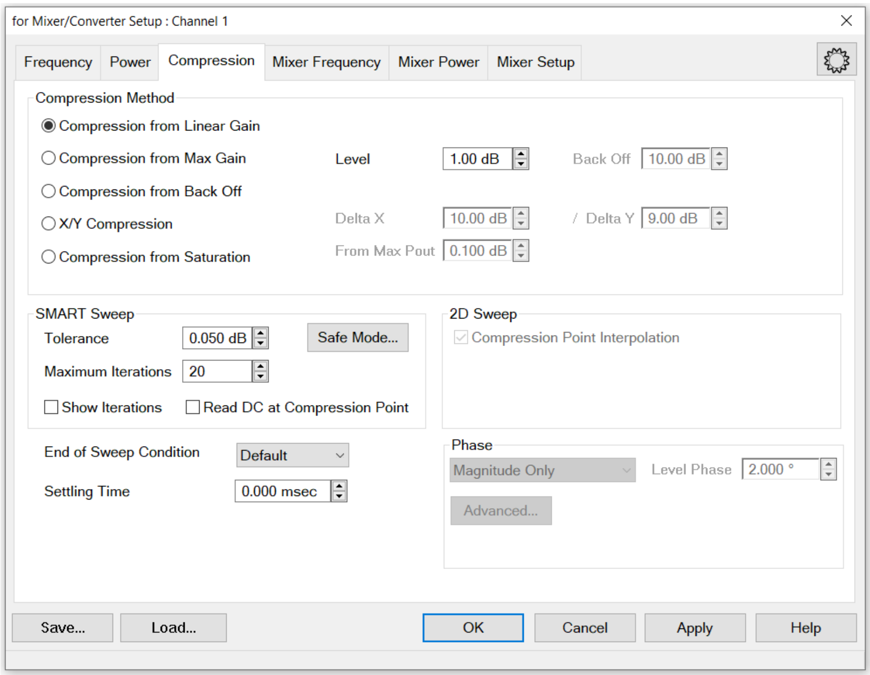


Figure 27. The gain-compression application offers multiple methods to define the compression point.

For converters that operate with fully saturated output amplifiers (for example, satellite transponders using TWT amplifiers and phase-modulated carriers), it is desirable to know the power-saturation level. TWT amplifiers exhibit unique behavior where increased input power beyond the saturation point results in decreased output power. The output-power peak is usually not very sharp, so small amounts of noise can result in large sweep-to-sweep variation of the input power corresponding to the maximum output power. To avoid this problem, at each frequency, the Compression from Saturation method looks for the maximum output-power value over the user-specified power-sweep range, and then finds the input power where the output power has decreased from the maximum by the user-specified “From Max Pout” value, typically a few tenths of a dB. This method gives repeatable saturation-power values for both input and output powers.

Intermodulation Distortion

Two-tone intermodulation distortion (IMD) is the most widely used measurement of in-band distortion for microwave devices. Two closely spaced signals are applied at power levels that cause the DUT to behave nonlinearly, which creates higher-order mixing products on either side of the two main signals. Since the PNA-X has two internal, well-filtered RF sources and a built-in signal combiner, it can easily create two-tone IMD stimuli. Other in-band distortion measurements such as NPR and EVM are discussed with the spectrum-analyzer and modulation-distortion applications.

The IMD application S93087B includes two measurement classes for converters: Swept IMD Converters and IM Spectrum Converters. The swept-IMD class is the primary measurement tool, providing swept measurements of center frequency, tone spacing, tone power, or LO power (Figure 28). The application measures the main signals and IMD products of order 2, 3, 5, 7, or 9. IMD parameters can be displayed as absolute power, relative to the main-signal power (dBc), or as input- or output-referred intercept points.

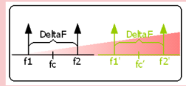
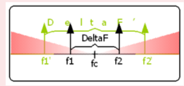
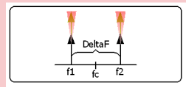
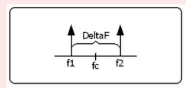
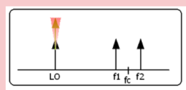
Sweep Type	Center Freq	Tone Spacing	Tone Powers	Diagram
Swept Fc	Swept	Fixed	Fixed	
Sweep Delta	Fixed	Swept	Fixed	
Power Sweep	Fixed	Fixed	Swept	
CW	Fixed	Fixed	Fixed	
LO Power Sweep	Fixed	Fixed	Fixed	

Figure 28. The swept-IMD application's sweep types include center frequency, tone spacing, tone power, LO power, or fixed (CW).

Figure 29 shows measurements of a converter configured as shown in Figure 2 (where the converter's LO is supplied by the optional third-internal source of the PNA-X), showing a swept-center-frequency sweep (left), and a swept-input-power sweep (right).

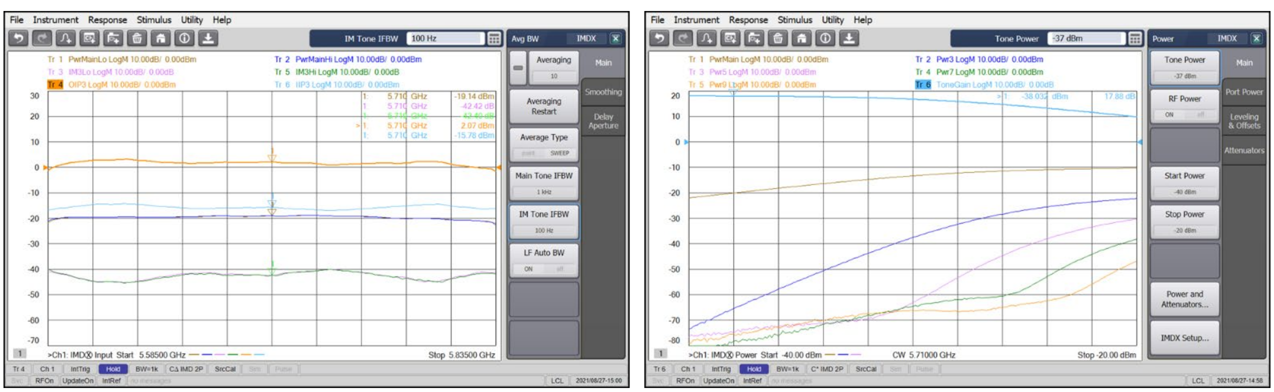


Figure 29. Examples of swept-center-frequency (left) and swept-power (right) IMD sweeps.

LO power sweeps

LO power sweeps are a very useful way for converter designers to find the optimum LO power that balances distortion performance with power consumption. Too little LO power starves the mixer and increases IMD. An LO with more power than needed for proper mixing adds needless components and wastes power, which is highly undesirable for devices with a limited energy supply such as battery-operated mobile phones or solar-powered satellites. This test is difficult and time consuming to do with the traditional approach using external signal generators and a spectrum analyzer, but quite easy with a PNA-X. Figure 30 shows an example of sweeping the power of an external LO signal while measuring the converter’s gain and IMD performance. Marker 1 has been placed in a reasonable spot where the mixer is operating away from starvation (i.e., in the region where the tone gain is constant). Doubling the LO power from 10 to 13 dBm (a 3 dB increase) only results in about 0.36 dB increase in the input-referred intercept point IIP3 (from -14.87 dBm to -14.51 dBm), which is probably not a worthwhile trade off.

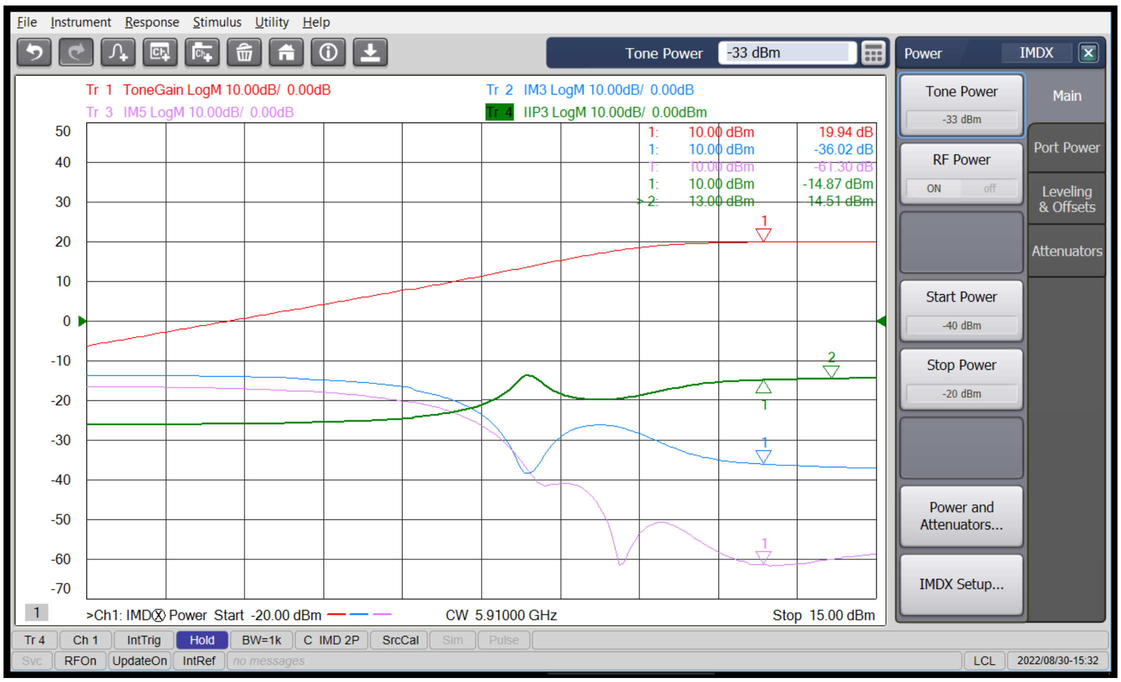


Figure 30. Example of sweeping the power of an external LO signal while measuring the converter’s gain and IMD performance.

IM Spectrum Converters

As mentioned above, S93087B includes a measurement class that provides a display of the frequency spectrum of the IMD signals. This tool provides a quick way to check for IMD imbalance between lower and upper sideband pairs and for looking for higher-order products. The marker feature “Marker -> IM Spectrum” opens an IM Spectrum channel in which the stimulus is set to the same frequency and power values corresponding the marker’s position in the Swept-IMD channel. Figure 31 shows an example of the spectrum display after using Marker -> Spectrum and adding markers. Markers 2, 3, 4, and 5 show IMD products of order 3, 5, 7, and 9 respectively. For example, marker 3 shows the amplitude of the fifth-order product generated as 3 * 509.5 MHz – 2 * 510.5 MHz = 507.5 MHz.

While useful for troubleshooting IMD setups, the IM Spectrum measurement class is not intended as a general-purpose spectrum analyzer, as it is considerably slower than the S93090xB spectrum analyzer application which has been optimized for speed.

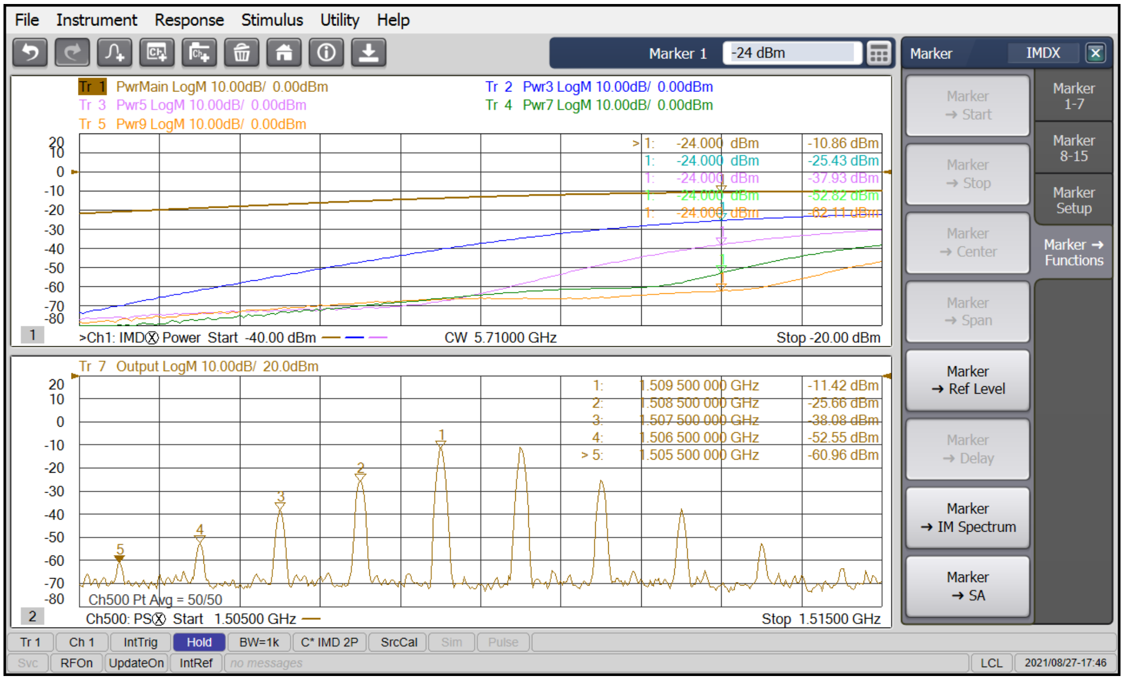


Figure 31. Example of IM Spectrum (bottom) after using Marker -> Spectrum.

Noise Figure

Noise figure (NF) measurements, expressed in dB, are the most common way to characterize how components or sub-systems degrade the SNR of signals passing through the device. SNR degradation can be caused by loss, extra noise generated by active devices, or both. Noise figure (or noise factor in linear terms) is defined as the input SNR divided by the output SNR, and it applies equally well to amplifiers and frequency converters. If an amplifier or frequency converter were perfect, the output noise would be equal to the input noise multiplied by the gain of the device, resulting in the same SNR at both the input and output. However, for all real-world active devices, the output noise is larger than the input noise multiplied by the gain of the DUT, so the SNR at the output will be smaller than that at the input, resulting in NF > 0 dB. It is important to note that when measuring and comparing noise figures, the test system is assumed to be 50 ohms. Deviations from non-50-ohm source match, if not compensated for, can cause significant measurement errors.

The PNA-X noise figure application S93029B, typically used in conjunction with an internal low-noise receiver (Option 029), uses the cold-source NF measurement method. Instead of using a calibrated noise source to provide two levels of input noise-power from which DUT gain and NF can be calculated (the Y-factor method), the cold-source method first measures DUT gain with S-parameters or their equivalent for converters, followed by a measurement of output-noise power with a single room-temperature termination at the DUT's input. From these two sets of measurements, NF can be calculated. This method, along with a source-impedance tuner, lends itself to advanced error correction that can overcome the accuracy degradation of a non-50-ohm source impedance, making the PNA-X the industry's most accurate NF measurement solution for amplifier and converter test. This approach also makes it easy to measure NF along with other essential converter measurements like conversion gain, compression, and IMD, all with a single set of connections (Figure 32). More details about the cold-source method and the PNA-X's unique vector-noise calibration can be found in application note "High-Accuracy Noise Figure Measurements Using the PNA-X Series Network Analyzer".

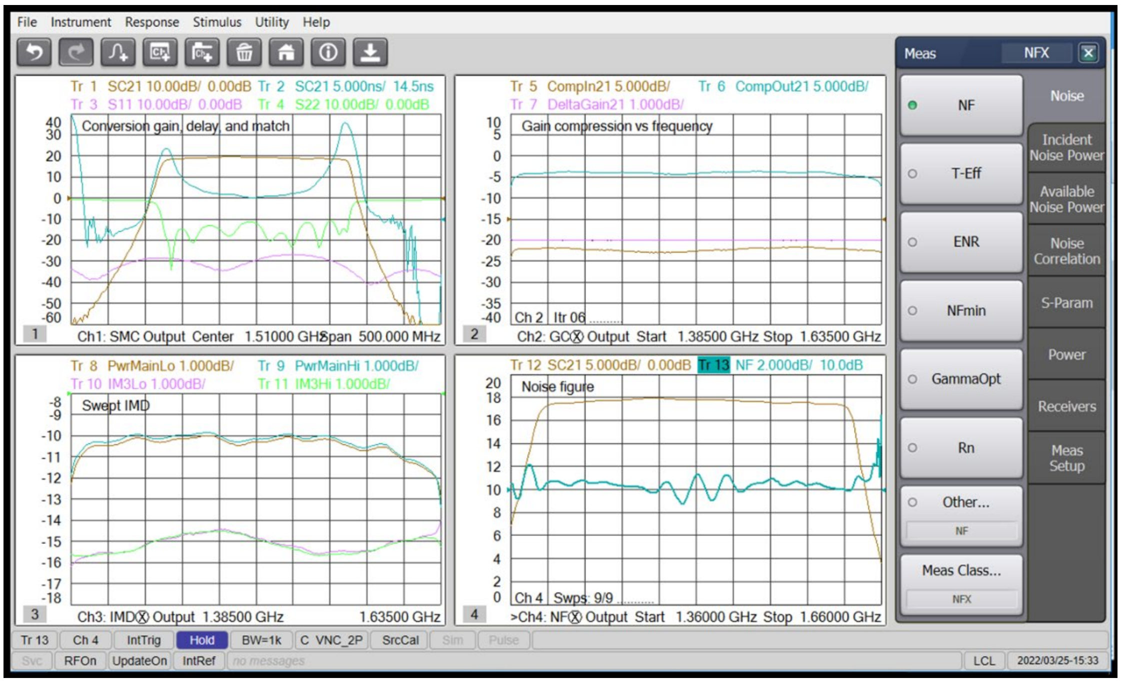


Figure 32. With a single set of connections, noise figure is easily included with other essential converter measurements like conversion gain, compression, and IMD.

Single- and double-sideband converters

One advantage of using the cold-source method is it correctly measures both single-sideband (SSB) and double-sideband (DSB) down-converters (Figure 33). DSB converters typically provide the most broadband frequency coverage, but the lack of a filter in front of the mixer means they have more down-converted noise than the equivalent SSB converter. Without the filter, input noise (typically from a front-end LNA) enters the mixer and conversion occurs at sidebands both above and below the LO frequency, resulting in both sidebands getting mixed to the IF output of the converter. The noise contributions of the two sidebands may not be equal, as it depends on the frequency response of the converter's front-end. If the front-end is flat between the upper and lower sideband, then the DSB-converter has 3 dB more noise than the filtered SSB equivalent. If the response is not flat, then the difference can be larger if the desired response is around the sideband with lower noise conversion. When using the Y-factor method, the measured NF would be the same for both DUTs, since a ratio of noise-power measurements is made, and the excess noise of the DSB converter (relative to the SSB converter) is ratioed out. For most DSB converters, Y-factor-based NF measurements typically read between 0 and 4 dB better (lower) than the actual NF value. Since the cold-source method measures noise power only once, there is no ratio effect, and the DSB and SSB converters will measure differently, and each will have the correct value for noise figure.

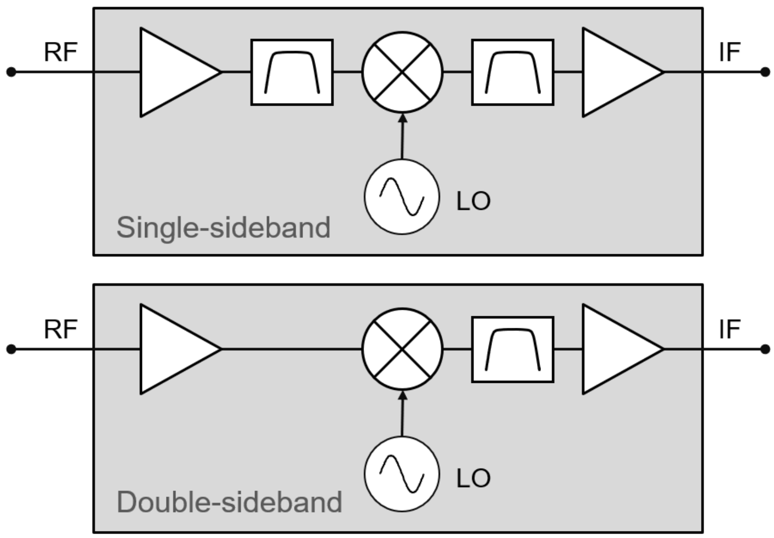


Figure 33. Single-sideband (top) and double-sideband (bottom) down-converters.

LO-noise contribution

A special consideration when measuring the NF of down converters concerns broadband noise on the LO signal. Noise on either side of the LO that is offset by the IF can mix within the first mixing stage of the DUT to add noise to the converted output. For example, consider a satellite-communications down converter with an IF of 140 MHz. LO noise that is offset from the nominal LO frequency by +/- 140 MHz converts to the IF and adds to the noise coming from the converter's front end. Depending on the mixer's conversion sensitivity to LO variations, the relative level of the LO noise, and the noise contributed by the converter's input path, the effect of LO-noise contribution varies from negligible to a significant noise contribution. The more gain present in the input path (and therefore more noise going into the first mixer), the less that LO noise affects the overall NF. If the LO is embedded within the DUT, then its noise contribution is not influenced by the test system. However, for test systems where the LO is provided externally, LO-noise contribution can be a significant source of measurement error.

This effect can easily be seen in Figure 34, showing measurement results from a down-converter with a 1.5 GHz LO provide by the PNA-X and an ouput of 321.4 MHz. In the lower-left window, the DUT consisted of just a mixer and an output filter. The orange trace (Tr 8-mem) shows the NF without any LO filtering. The red trace is the NF with a 1.5 GHZ bandpass filter added to the LO signal to remove noise. The difference in NF between the two conditions is shown by the delta-marker as 13.1 dB, a huge difference. In the upper-left window, an input amplifier was added with 15.3 dB of gain which reduced the difference between a filtered and unfiltered LO to 1 dB. In the upper-right window, the gain of the input amplifier was increased by 20 dB to 35.4 dB, reducing the difference between a filtered and unfiltered LO to just 0.1 dB.

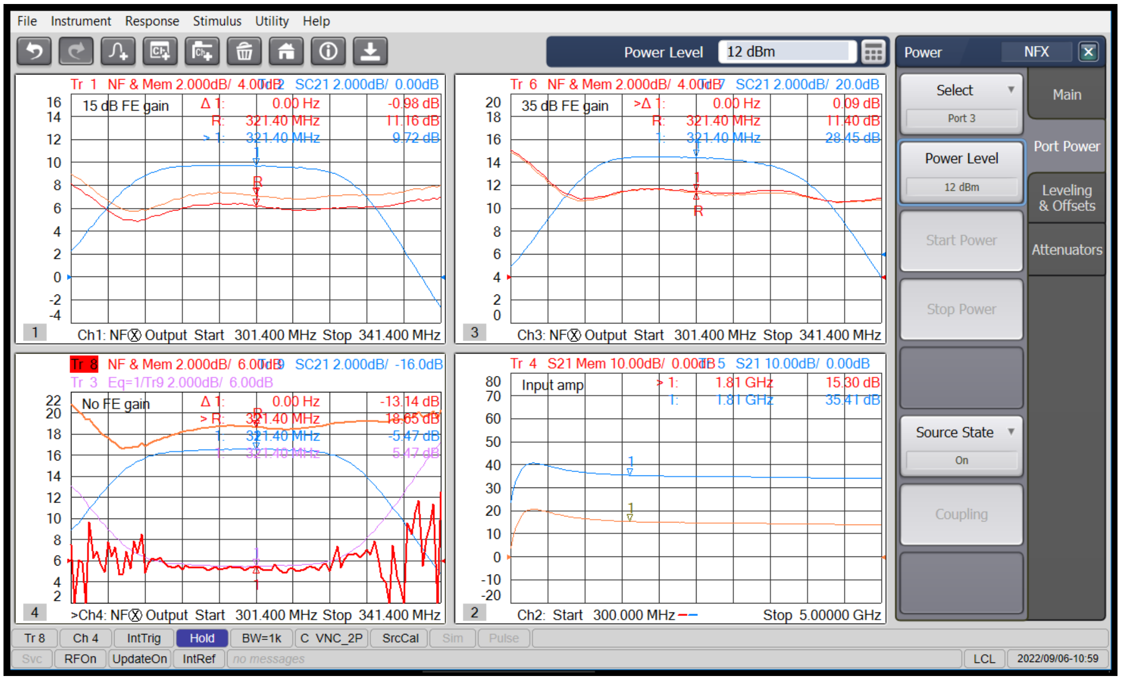


Figure 34. LO-noise contributions to noise figure with 0 dB (lower left), 15 dB (upper left), and 35 dB (upper right) of front-end gain.

Spurious Tests

Characterizing harmonic and non-harmonic spurious signals that emanate from the ports of a frequency converter is a critical piece of the test suite, since these signals can cause in-band and out-of-band interference when operating in their intended environments. Despite liberal use of filters within a typical converter, spurious signals generated by internal mixers, amplifiers, and frequency synthesizers can leak out of any of the RF ports. This is especially problematic for devices with a direct antenna port, as these spurs would then be broadcast to the operating environment. Spurious testing using legacy methods with standalone sources and spectrum analyzers typically takes up the largest portion of the overall test time.

Using the S93090xB Spectrum Analysis application, any of the test receivers within the PNA-X can be used as a spectrum analyzer, allowing spurious test at all the converter's RF ports (Figure 35). Within the spectrum analysis (SA) measurement class, any of the internal RF sources can be set to the desired frequency and power to provide the necessary RF, IF, and LO stimuli needed during spurious tests. And by taking advantage of VNA error-correction methods, calibrated SA measurements can be made at any desired reference plane, whether coaxial, waveguide, or on-wafer.

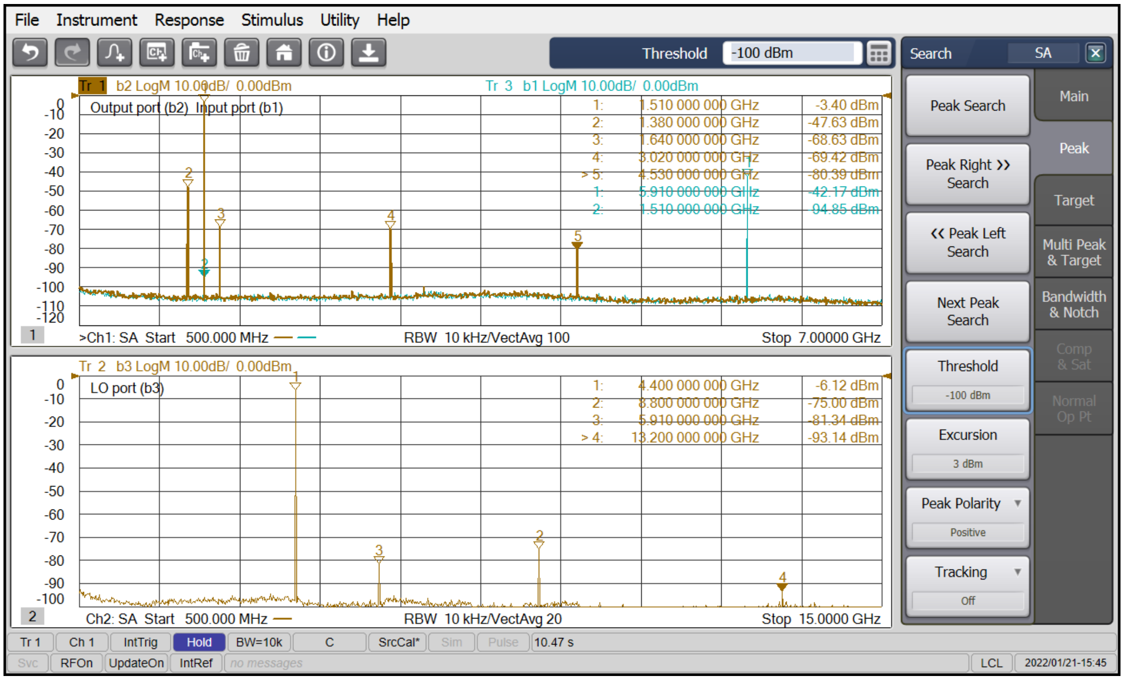


Figure 35. Spectrum measurements of spurious signal emanating from the input, output, and LO ports of a converter.

Since the PNA-X does not have preselector filters in front of its receivers and because the SA application uses stepped Fast Fourier-Transform (FFT) sweeps with optimized data processing, measurements are very fast, often ten to hundreds of times faster than conventional swept-LO benchtop spectrum analyzers (Figure 36).

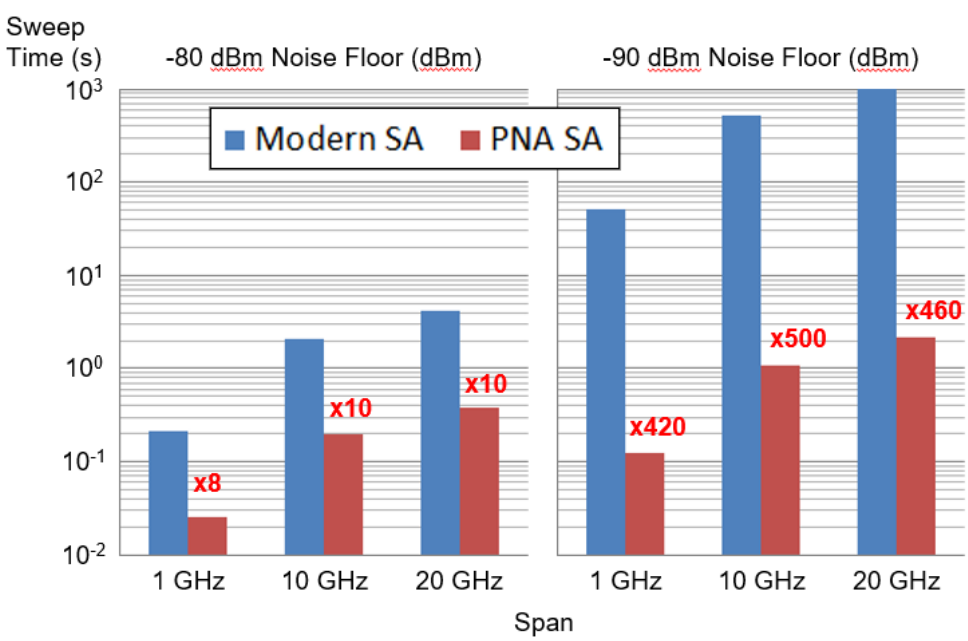


Figure 36. Spectrum measurements with the PNA-X are very fast, often ten to hundreds of times faster than conventional swept-LO benchtop spectrum analyzers.

Image rejection

Since there are no filters ahead of the mixers in the PNA-X receivers, the display would show image signals if image-rejection methods were not employed. Images occur in swept measurements when an input signal is mixed into the receiver's IF when the LO is both below and above the signal, creating both a desired and a false response. To overcome this problem, a software preselection algorithm is used that borrows from a method that has been employed with unpreselected microwave spectrum analyzers for many decades, often called "signal identification" or sig-ID. At its simplest form, two sweeps are taken, with one sweep using low-side mixing and the other high-side mixing. Real signals will remain in the same position on the display, while false signals will move to a different position. By using the smallest value of the two sweeps for any trace point, the image signal is not displayed. However, in dense signal environments, two LO frequencies may not be enough to eliminate image responses. The SA application employed in the PNA-X defaults to four LO acquisitions for each displayed data point. The user can decrease it to one LO acquisition when measurement speed is most important or increase it to six or eight acquisitions when better image rejection is desired. For even better rejection, the LO frequencies are randomized by default to make it even less likely that an erroneous signal will be displayed. Figure 37 shows a display with normal image rejection applied (top window), and a special diagnostic mode that shows the image signals (bottom window).

Figure 37. Normal image rejection (top) and special diagnostic mode showing the image signals (bottom).

Filter shape

The digital filters used for S-parameter measurements are optimized for speed and don't require high selectivity, since stimulus/response testing always places the measured signal in the center of the IF bandwidth. However, for spectrum analysis, high selectivity is very important to distinguish between closely spaced signals. This requires Gaussian-shaped digital filters, which typically require more filter taps than an equivalent-bandwidth filter used for network analysis. Figure 38 shows the difference in selectivity of a 3 kHz filter for network-analysis (Tr 1) and spectrum-analysis (Tr 2) modes.

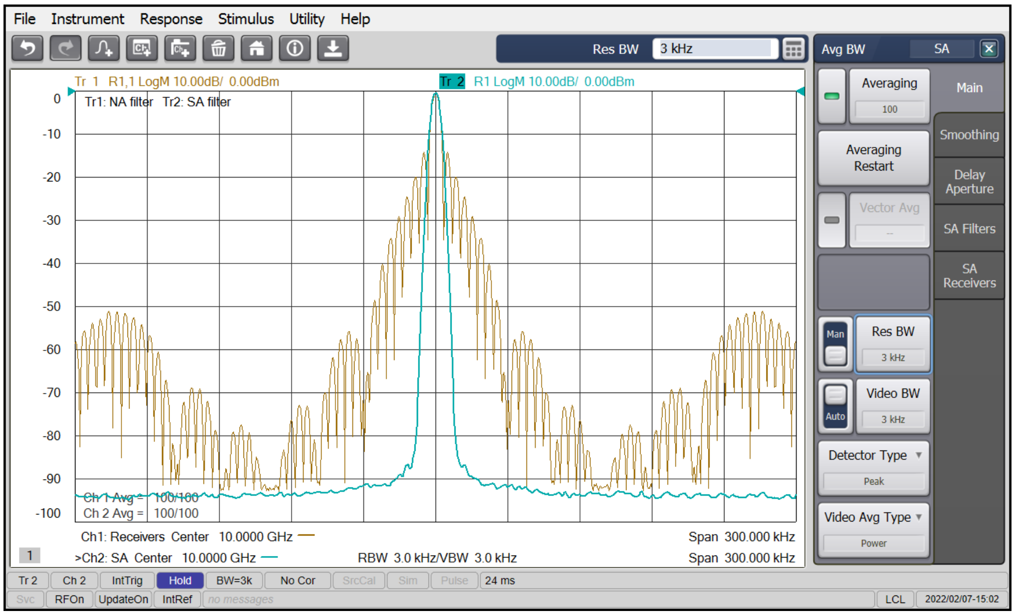


Figure 38. Filter-selectivity comparison between network-analysis (Tr 1) and spectrum-analysis (Tr 2) modes.

Coherence and vector averaging

The SA application has a unique and very useful feature that extends the dynamic range of spectrum measurements by taking advantage of repetitive waveforms and the tight coupling between the sources and receivers. This feature is called vector averaging and is accomplished by averaging successive data records in the time domain, prior to the FFT process. As long as there is coherence from one data record to the next, the averaging process reduces noise while preserving the underlying waveform. The reduction in noise in the time domain causes a drop in the noise floor of spectrum measurements, at the expense of longer measurement times. However, time-domain averaging before the FFT is much faster than normal SA-channel averaging, which is done on successive FFT-processed sweeps.

When the SA application is in multitone mode (enabled under the Coherence tab of the SA Setup dialog box), the application ensures that all stimulus signals, whether CW or multitone, line up on the underlying FFT grid used to create the spectrum plot. In the time domain, this means that the sampled-data record is N times the repetition rate of the stimulus, allowing use of vector averaging. Figure 39 shows an example of a two-tone IMD measurement with (red trace) and without (orange trace) vector averaging. Using 1000 vector averages, the noise floor dropped by 30 dB, significantly improving the SNR for the IMD measurements and revealing a 7^{th}-order IMD product that was well below the non-vector-averaged noise floor.

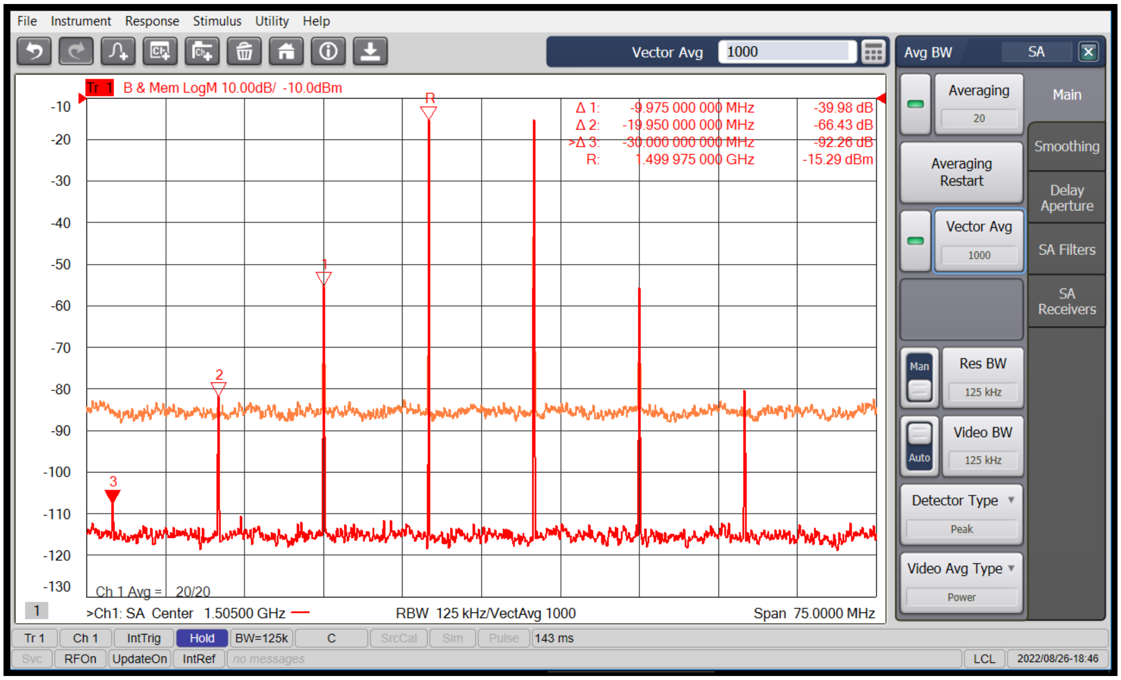


Figure 39. Example two-tone IMD measurements with (red trace) and without (orange trace) vector averaging. Using 1000 vector averages, the noise floor dropped by 30 dB.

Mixer Spurs

One key requirement of frequency-converter design is knowing the level of spurious signals coming from the internal mixers. This information is necessary to design filters with sufficient stopband rejection to meet the overall spurious-signal specifications of the converter. All mixers generate signals at $m*F_{RF} \pm n*F_{LO}$ where m and n are integers. Usually, only one set of m and n are used for the desired mixing product (e.g. m = n = 1, or m =1 and n = 3), while the other sets represent spurious mixing products of various orders.

While the Spectrum Analysis application can be used to characterize unfiltered mixers, a better choice (especially for swept measurements) is the S93089B Differential and I/Q Devices application, which is suitable for both single-ended and differential mixers. For a mixer used as a down-converter, the sum spurious products (m*FRF + nFLO) are often way out of band and therefore easily filtered, but some of the difference products (m*FRF – nFLO) often fall close to the desired band of operation. Figure 40 shows the frequency-range definitions used to test a single-ended down-converting mixer where both the RF and LO signals are swept and only the difference products are measured. Frequency ranges were first defined for the RF and LO signals, and then for the second and third harmonics of the LO using multipliers. With this core set of frequency ranges, it is then easy to use the multiplier and offset fields to define additional frequency ranges for the desired signal (1.5 GHz in this case) and for the spurious mixing products, where in this example, m and n are 1, 2, and 3. An example of the frequency-setup dialog box is shown for F13. This approach makes it easy to add ranges for higher order products beyond those shown in the example, or for including the sum products (checking the “Up” box will add the coupled and offset frequencies).

Looking at the frequency range F12 (which represents $2{*}F_{RF} – 3{*}F_{LO}$), it is clear from the start and stop frequencies that a spur will cross the desired IF of 1.5 GHz during the sweep (this will occur when the RF is at 6 GHz and the LO is at 4.5 GHz: 2*6 GHz – 3*4.5 GHz = -1.5 GHz). Although the frequency range shows mathematically correct negative values, the actual signals will show up as positive frequencies.

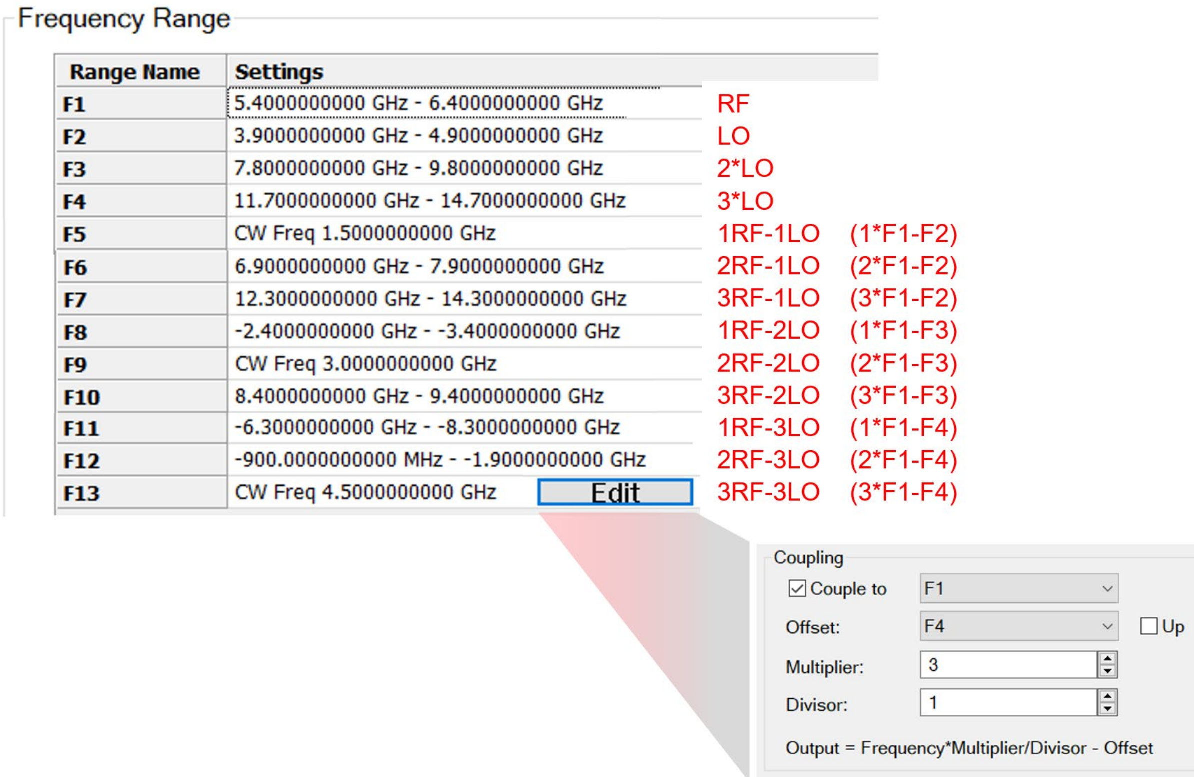


Figure 40. Example set of frequency ranges to measure the desired and spurious signals of $m{*}F_{RF} – nF_{LO}$, where m and n are 1, 2, and 3.

The results of the measurement are shown in Figure 41. All the mixing products are relative to the input signal power. The marker on Tr 3 shows the mixer's conversion loss for the desired RF – LO product as -6.0 dB. The marker on Tr 10 shows the crossing spur at -1.5 GHz as -62 dB, which is 56 dB below the desired IF product. Since spurious products are often a function of input power, this set of measurements can be repeated in the same channel or duplicated in other channels with different input-power levels. The frequency axis for each parameter can be set by the user, so all of the products can share a common frequency axis if desired.

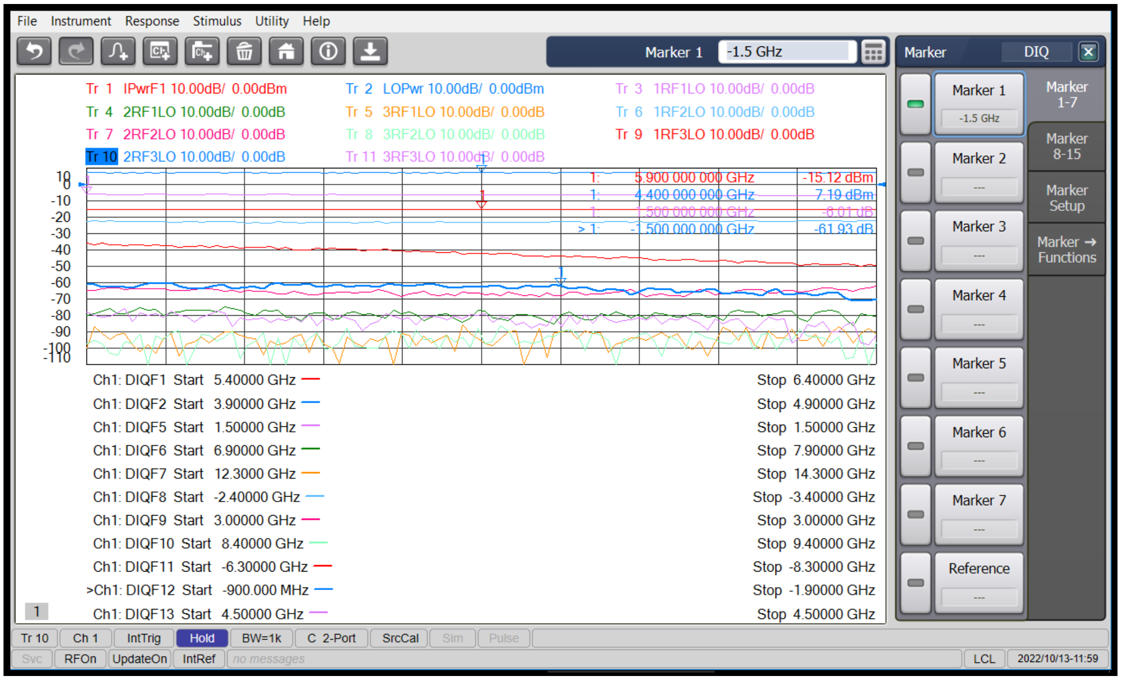


Figure 41. Example swept measurements of the desired and spurious mixing products for a down-converting mixer using the Differential and IQ Devices application.

Pulsed-RF Testing

The measurements described so far can be made using pulsed-RF stimulus by taking advantage of the PNA-X's built-in pulse modulators and set of five pulse generators. Pulsed stimulus is useful for high-power devices that might be damaged with continuous-wave stimulus, or to simulate real-world operating conditions for devices like transmit/receive modules. Pulsed-RF stimulus can be combined with CW sources used as LO signals. For devices that require pulsed bias, the signals from the internal pulse generators can be routed out the rear panel to control external pulsed DC-power supplies. An external pulse generator can also supply a pulse trigger to the PNA-X to synchronize external and internal pulse generators. The dialog boxes used to set up pulse timing are shown in Figure 42.

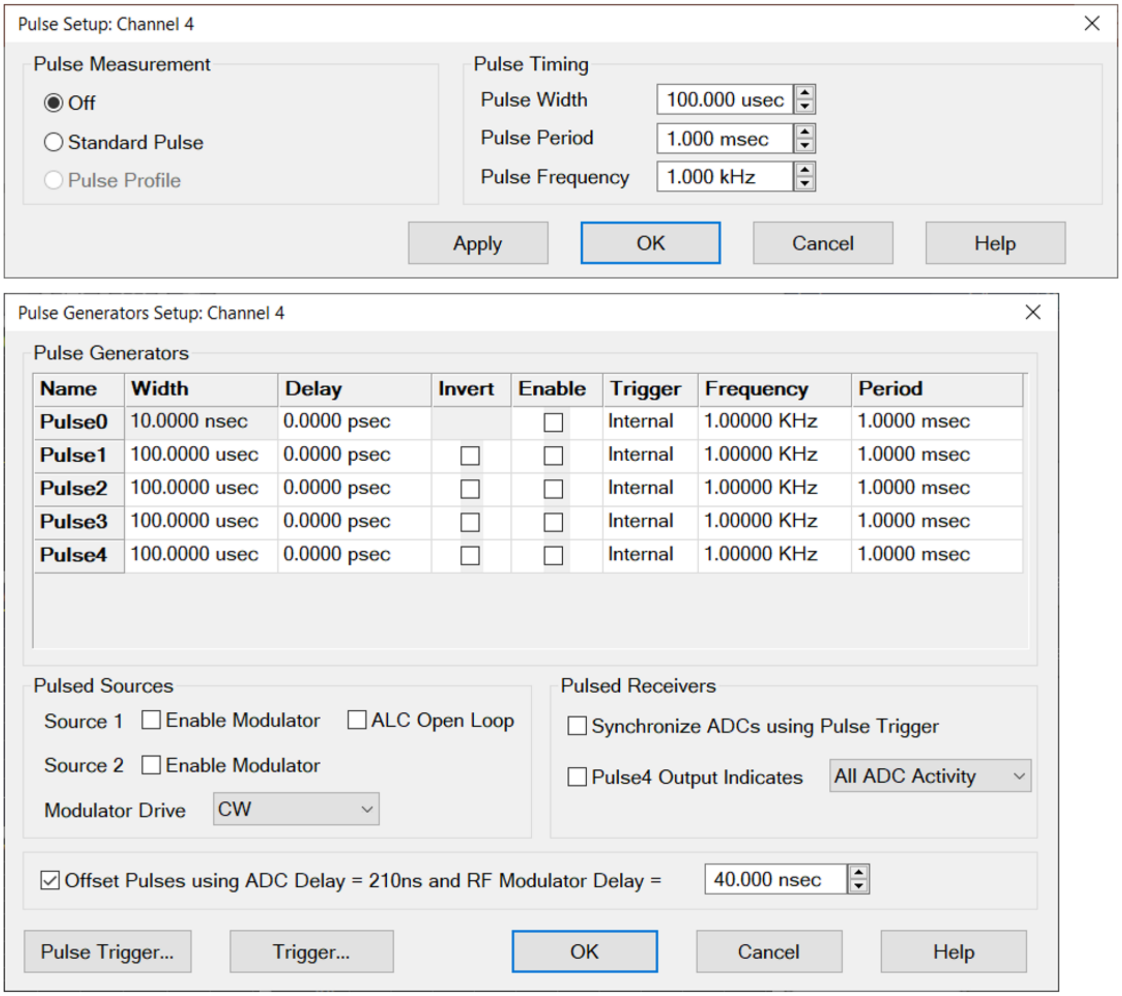


Figure 42. Basic dialog box to turn on and set up pulse timing (top) and advanced dialog box for full pulse-generator and modulator control (bottom).

Receiver leveling

When the internal RF sources are pulsed, the firmware automatically turns off the automatic-level-control (ALC) circuitry to prevent unleveled errors, and sets the leveling mode to Open Loop. This mode has less stimulus-power accuracy than the internal ALC mode but is compatible with pulsed-RF operation. When measuring S-parameters or transmission-conversion parameters, which assume linear operation of the DUT, precise stimulus-power control is not important. For measurements of nonlinear behavior like compression and IMD, stimulus-power accuracy is very important. For those measurements, receiver leveling is recommended. Receiver leveling, coupled with power-meter calibration, provides accurate and precise power control over a broad range of pulsed-RF stimulus powers due to the high linearity of the measurement receivers.

More pulsed-RF resources

Two other application notes are available that go into more detail on the theory and setup of pulsed-RF PNA-X measurements:

- Wideband and Narrowband Detection for Pulsed-RF Component Testing
- Active-Device Characterization in Pulsed Operation

Phase-Noise Measurements

The change to DDS sources in the PNA family from the earlier fractional-N-based design improved the phase noise performance of the internal sources by approximately 30 dB in the critical 1 kHz to 100 kHz carrier-offset region (with enhanced low-phase-noise option UNY). This improvement makes it practical to measure the phase noise of internal LOs embedded inside frequency converters. To take advantage of the improved phase-noise performance, the S930317B Phase Noise Measurements application was created for measurements of phase noise, AM noise, residual noise, and integrated noise on LOs, with carrier offsets from 0.1 Hz to 10 MHz. The PNA-X phase-noise-measurement noise floor is comparable to that of single-channel, high-performance signal analyzers with phase-noise-measurement capability, like the UXA signal analyzer. While single-channel instruments like the PNA-X and UXA do not utilize cross-correlation methods as provided by dedicated phase-noise analyzers like the Keysight N5511A phase noise test system (PNTS) which offers the lowest noise floor for phase-noise measurements, the performance is adequate for most embedded-LOs. Figure 43 shows example measurements of phase noise and amplitude noise of the output from an embedded-LO frequency converter.

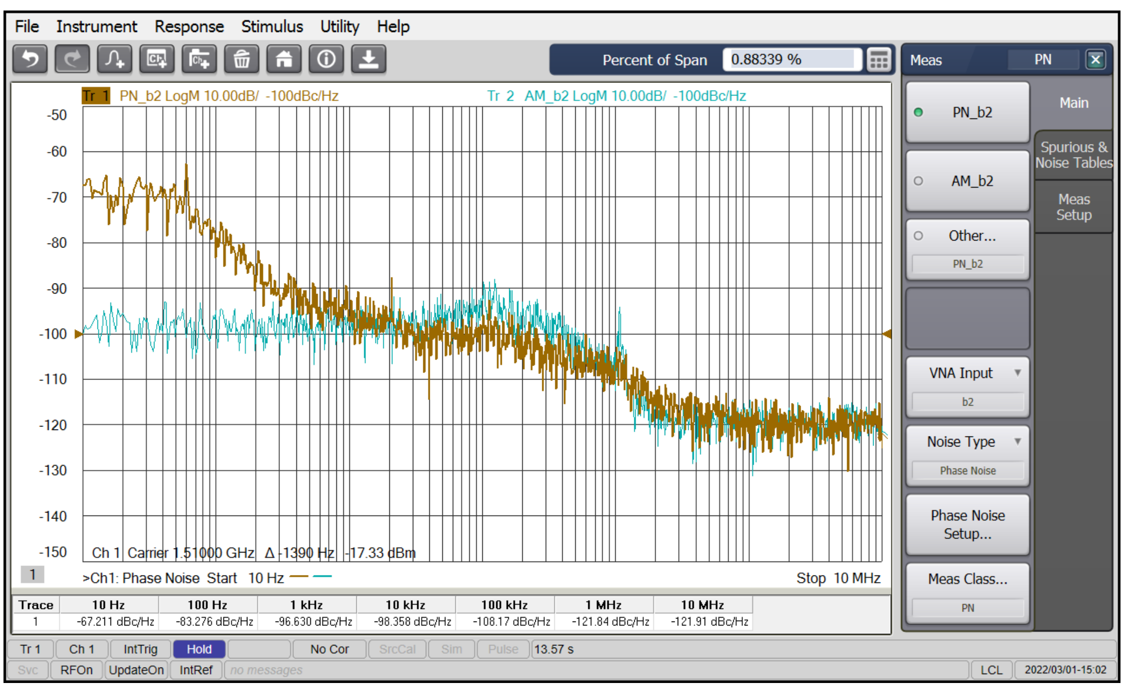


Figure 43. Phase-noise (Tr 1) and amplitude-noise (Tr 2) measurements of an embedded-LO frequency converter.

Modulated-Carrier Measurements

All the distortion measurements discussed so far have either used a single-sinusoidal test input (not counting the LO signals), or in the case of IMD, two sinusoidal inputs. Modulated-carrier measurements use a broadband stimulus meant to fill a channel or frequency band of interest in order to simulate real-world operation conditions. Creating modulated-carrier test signals is easily accomplished using modern vector-signal generators or the internal DDS sources of the PNA-X. Specific modulation formats can be used (for example, QPSK or 16QAM), or the signal can be purely noise-like. Depending on the use case, the bandwidth of the test stimulus can vary from a few tens of megahertz to several gigahertz. Figure 44 shows the various vector-modulation types that can be generated directly by the PNA-X.

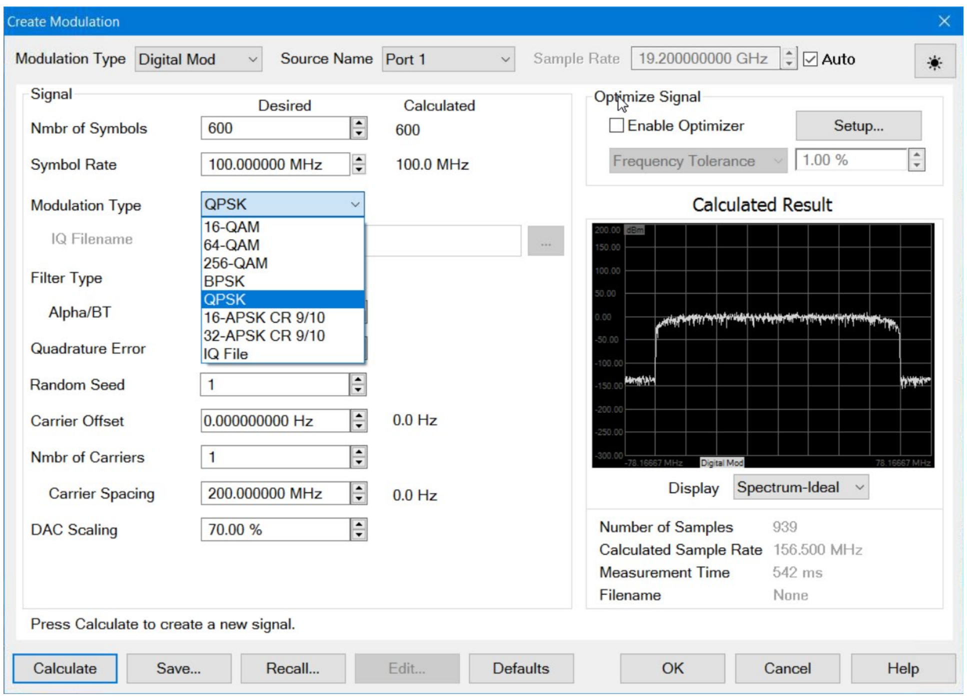


Figure 44. Examples of vector-modulation types that can be generated directly by the PNA-X.

When using modulated carriers for testing devices, a repetitive test signal is used. This means that in the frequency domain it looks like a multitone signal, typically composed of between several hundred to many thousands of individual sinusoids. When the converter is driven into its nonlinear region, intermodulation distortion will be generated from interactions between all the individual signals, creating a broadband distortion spectrum. Measurements done with a modulated carrier can be described as being either in-band or out-of-band. In-band measurements are done within the same frequency band as the test stimulus and include noise-power ratio (NPR) and error-vector magnitude (EVM). Out-of-band measurements characterize the distortion spectrum created on either side of the test-signal spectrum and include adjacent-channel power ratio (ACPR).

Adjacent-channel power ratio (ACPR)

One common out-of-band measurement using a modulated carrier is ACPR, where the powers in the lower and upper adjacent channels are compared to the power within the main band. The adjacent channels are typically one or two bands on either side of the center band. ACPR is easily measured using the S93090xB Spectrum Analysis application, as no special processing is needed to see the power spectrum of all the bands of interest. The test waveform can be noise-like or use a common modulation format. An example using a 100 Msymbols/s QPSK test signal generated from port 1 of the PNA-X is shown in Figure 45, where band markers are used to calculate ACPR of the lower and upper adjacent channels. At an input band-power level of -25 dBm, the input ACPR measured with Tr 1 is about -48 dBc, and the output ACPR measured with Tr 2 is about -36 dBc, a degradation of 12 dB. ACPR can also be measured using the S93070xB Modulation Distortion application, discussed below in the EVM section.

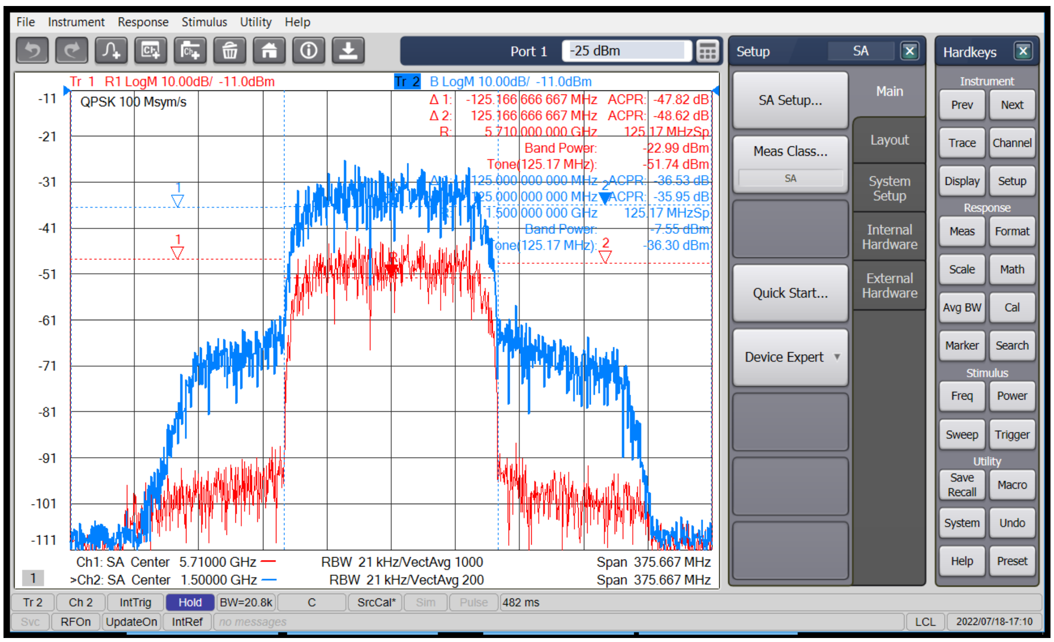


Figure 45. Example ACPR measurements using band markers to show input (Tr 1) and output (Tr 2) ACPR.

Noise power ratio (NPR)

Traditionally, in-band distortion of microwave frequency converters has been characterized by two-tone IMD measurements, a topic covered earlier. While straightforward to set up and measure, for many devices, a two-tone stimulus is only an approximation of the wideband signals seen during actual device operation. For example, satellite transponders typically carry wideband signals that fill the entire operational band, generating many more intermodulation products compared to the two-tone case. One way to characterize the in-band distortion is to create a deep, in-band notch in the spectrum of a wideband noise-like stimulus signal that fills the band of interest (the notch is usually in the middle of the band). Then, the amount of noise that has filled the notch (i.e. raised the noise floor) is measured at the output of the converter. The commonly used figure of merit is noise-power ratio (NPR), which is the ratio (or difference in dB) of the average in-band power level away from the notch to the band-power level within the notch, as shown in Figure 46. The notch width is typically between 1% and 10% of the channel bandwidth, and the notch depth ideally should be at least 10 dB below the expected output-distortion floor.

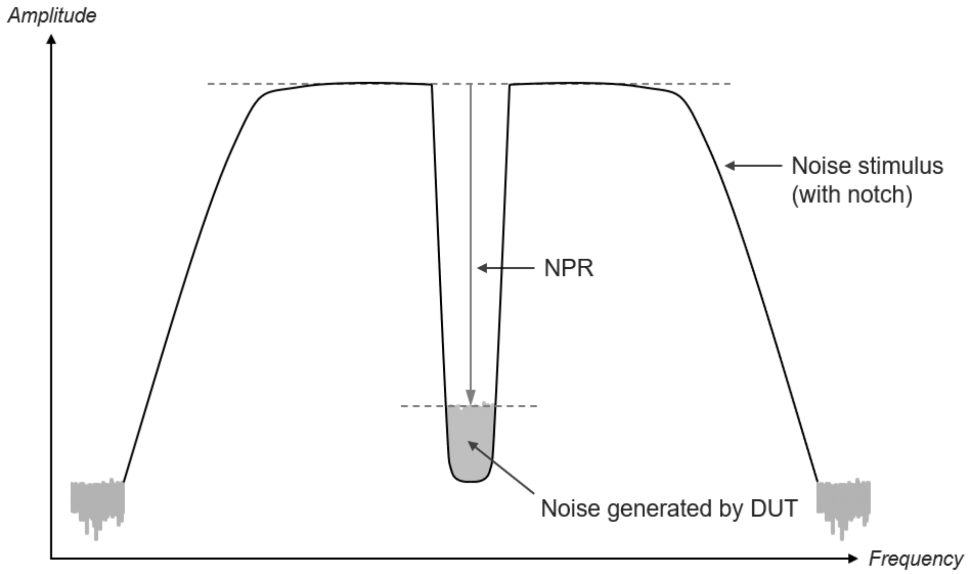


Figure 46. Noise-power ratio (NPR) is the ratio (or difference in dB) of the average in-band power level away from the notch to the band-power level within the notch.

While NPR originated long-ago in the world of analog telephone signals, where one voice channel was removed from a group of voice carriers to create the notch, the RF and microwave community historically created the NPR stimulus with a noise source followed by an amplifier followed by a notch filter. In recent years, as vector-signal generators have become both common and affordable, the wideband stimulus can easily be created as a multitone signal, with multiple sinusoids (typically 500 or more) spaced closely in frequency to fill the desired frequency band. To simulate wideband additive white Gaussian noise (AWGN), the signals typically have equal magnitudes and random phase, with amplitude levels set to create the desired overall band power. Any arbitrary number of the signals can be removed to create the notch by proper definition of the digital waveform used to modulate the RF carrier. While this used to require an external arbitrary waveform generator driving analog I/Q inputs of a vector RF source, most modern vector sources can directly use a digital waveform file to create the NPR test signal. In some cases, upconverters are used to translate the NPR signal to frequency bands above the upper range of the RF source. Keysight's vector signal generators can create wideband NPR signal covering RF, microwave and millimeter-wave frequencies, and the internal DDS sources in the PNA-X can be used to create NPR signals with carrier frequencies up to 6 GHz and waveform periods up to 6.28 us.

NPR measurements can be made with the S93090xB Spectrum Analysis application, or the S93070xB Modulation Distortion application. This section covers use of the SA application which can both control a vector source and set up the receivers for the NPR measurement. Creation of the NPR signal is accomplished using the Create Modulation dialog box shown in Figure 47. In this example, the internal DDS source at port 1 of the PNA-X was used to create the 250 MHz-wide NPR signal. The tone spacing of 161.3 kHz results from a waveform period of 6.2 us, which gives 1551 tones equally spaced across the 250 MHz bandwidth.

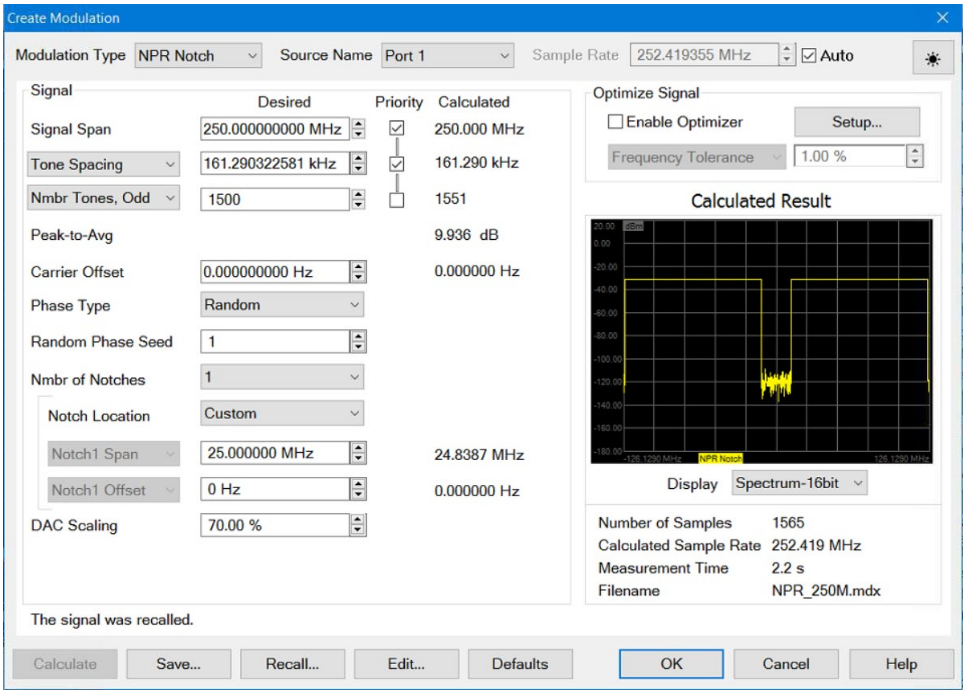


Figure 47. The Create Modulation setup dialog box is used to define the parameters of the NPR signal.

Since the input NPR signal can be measured with the R1 (a1) receiver, the firmware can correct the non-flatness of the original signal resulting from the frequency-response of the test system. Figure 48 shows the input signal both before (orange trace) and after (red trace) source correction. In addition to flattening the input tones, the distortion in the notch caused by nonlinearities of the source can be lowered as part of the source-correction process, which also can be seen on the input signal in Figure 48. After source correction was applied, the input NPR was -56.7 dB. The output signal (blue trace) shows that the notch has filled up due the converter's nonlinearity, with an output NRP of -37.2 dB. In addition, some distortion-caused spectral leakage into the adjacent channels can be seen.

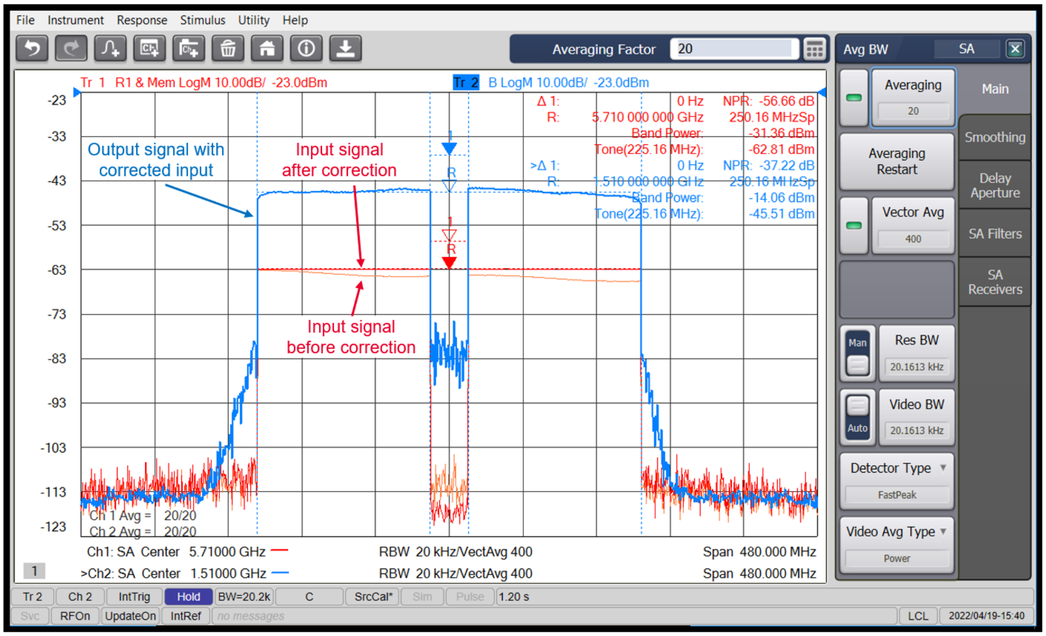


Figure 48. NPR measurement example showing the input signal (Tr 1) with and without source correction and the output signal (Tr 2).

Error-vector magnitude (EVM)

EVM is a method used to characterize in-band distortion with a wideband stimulus that does not require a notch within the band, as is done with NPR. It is a commonly used figure of merit for modulation quality in the sub-6 GHz commercial-communications ecosystem, but traditionally has not been used for other industries which instead have relied on two-tone IMD or NPR testing. With the advent of 5G and associated millimeter-wave bands, EVM is becoming more common in the microwave-test community.

The traditional way to measure EVM is to use vector-signal-analysis software like Keysight's 89600 PathWave VSA to demodulate the DUT's output signal in the time domain. The test waveform consists of a carrier modulated with some number of symbols in a particular modulation format. For each demodulated symbol, the magnitude of the error from the ideal symbol is calculated. The error is normalized to the amplitude of the outer-most symbol or to the square-root of the average symbol power. All the symbol errors are then combined using a root-mean-square summation. The overall EVM value can be converted into an equivalent distortion-power level and expressed as a percentage or dBc value of the total in-band power. However, EVM can also be calculated entirely within the frequency domain, without performing any time-domain symbol demodulation. This unique approach developed by Keysight is used in the S93070xB Modulation Distortion application. The method uses spectral correlation between the input and output spectrums to calculate the distortion spectrum, so the EVM measurement floor is not limited by the modulation quality of the vector source, unlike the case when doing symbol demodulation. This means the PNA-X method has the lowest residual EVM in the region where EVM is not limited by system noise. As was shown in the NPR section, source-correction can also be applied to EVM test signals to create more ideal test waveforms. And since demodulation is not done, non-symbol waveforms can be used as the test stimulus, such as flat tones with random phases which simulates additive white Gaussian noise (AWGN). More information about EVM using the PNA-X can be found in the application note "Create Accurate EVM Measurements with the PNA-X Series Network Analyzer".

Figure 49 shows the RF Path and Mixer tabs of the Modulation Distortion Converters setup dialog box for the example measurement shown in Figure 50 which uses a 16QAM test signal consisting of 600 symbols at 150 Msymbol/s. In this measurement, Tr 1 shows the input signal both before (orange trace) and after a source correction (red trace) was performed. The source correction improved the ACPR of the input signal by about 20 dB. Tr 2 shows the output of the converter, while Tr 3 shows the in-band distortion, from which EVM is calculated. An SC21 log-mag trace (Tr 5) is included to compare the converter's bandwidth to the modulated-signal bandwidth. The measured gain in the SMC channel closely matches the MGain21 (Tr 4) measured in the modulation-distortion channel as shown by the markers.

The table at the bottom of the window contains a lot of useful information, including the 2.42% equalized EVM of the converter (EVM DistEq21), where the contribution of the input EVM has been removed. The 3.06% single-ended equalized output EVM (EVM OutEq2) is what the symbol-demodulation method would report. This value is higher than the frequency-domain-based EVM as it includes the EVM of the input signal.

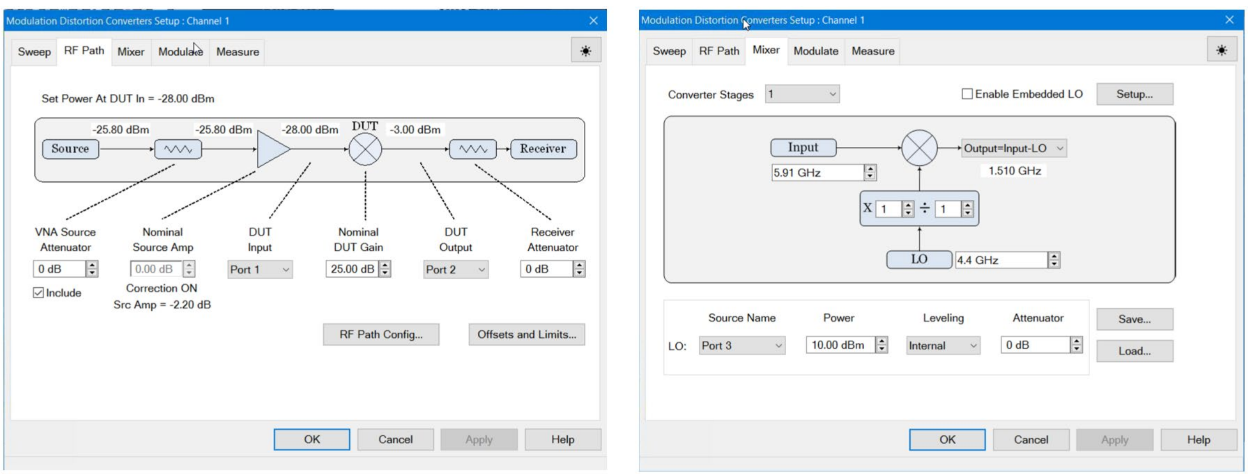


Figure 49. RF Path and Mixer tabs of the Modulation Distortion Converters Setup dialog box for the example measurement of a frequency converter in Figure 50.

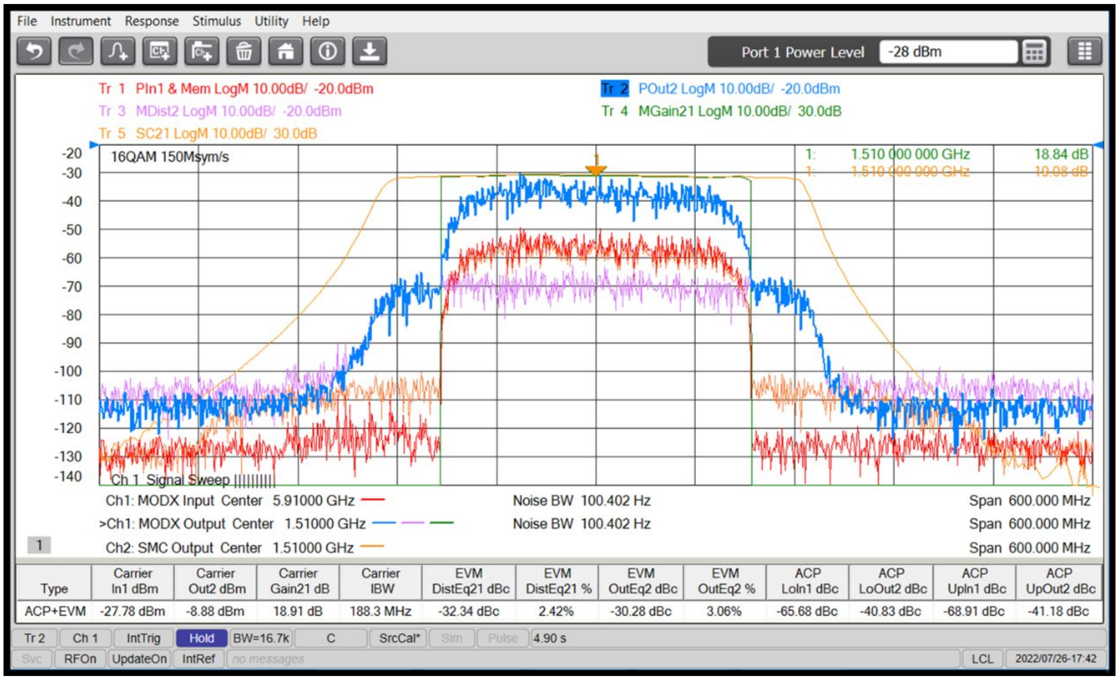


Figure 50. Example Modulation Distortion Converters measurement of ACPR and EVM with a 16QAM test signal consisting of 600 samples at 150 Msymbol/s.

The reconstructed time-domain I/Q data from the modulation-distortion channel can be streamed to Keysight's 89600 PathWave VSA software to show the output constellation diagram and the calculated EVM after symbol demodulation of the output waveform. The results of this are shown in Figure 51. The upper-left window shows the equalized constellation diagram of the 16QAM output signal, and the lower-left window shows the output spectrum which closely matches Tr 2 of Figure 50. The upper-right window shows the frequency response of the 13-symbol equalization filter, which results in an EVM of 2.78% shown in the table on the lower right. This value is a bit smaller than the single-ended EVM value calculated in the modulation-distortion channel, likely due to small differences in the respective equalization filters.

noise figure, spurious, phase noise, and modulated-carrier evaluation of ACPR, NPR, and EVM.

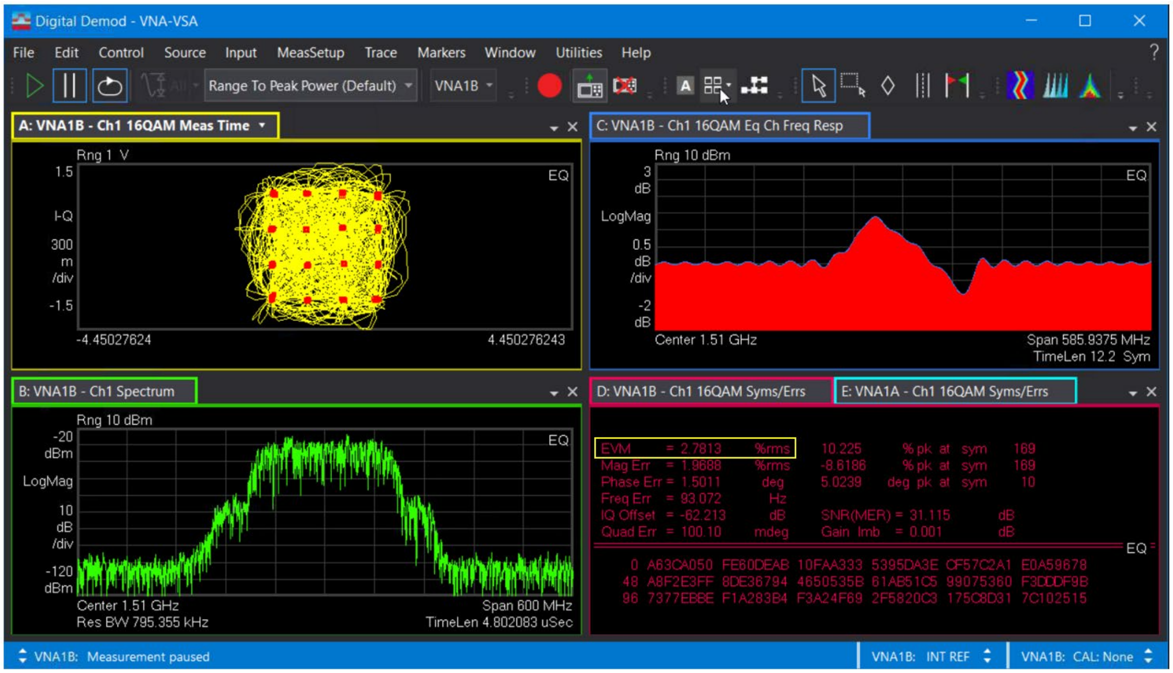


Figure 51. 89600 PathWave VSA results using the reconstructed time-domain I/Q data from the modulation-distortion channel showing a constellation diagram, frequency spectrum, equalization-filter response, and EVM calculation.

Conclusion

This application note has shown that the PNA-X's flexible hardware combined with many software measurement applications enable a broad range of frequency-converter measurements used to characterize linear and nonlinear behavior, all with a single set of connections the DUT. This is accomplished with frequency and power sweeps using combinations of CW, pulsed, and modulated carriers. Calibration methods have been developed for each measurement application to provide high accuracy. Linear measurements include conversion gain or loss, input and output match, group delay, deviation from linear phase, LO-path-phase variation, and isolation. Measurements to characterize nonlinear behavior include gain and phase compression, AM-to-PM conversion, phase transfer, IMD, noise figure, spurious, phase noise, and modulated-carrier evaluation of ACPR, NPR, and EVM.

Application note 1408-23 by David Ballo

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