

Optimizing IP3 and ACPR Measurements

Tips for Getting the Best Performance from Vector Signal Analyzers



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1. Overview

As wireless communication becomes widespread, the increased complexity of the devices and wireless protocols create an unrelenting demand for linear RF components and systems. These ubiquitous wireless devices require high performance test systems to characterize linearity. Linearity performance, particularly third order intercept point (IP3 or TOI), has become a defining characteristic as it affects power efficiency, channel density, signal coverage, and adjacent channel power ratio (ACPR).

IP3 performance is a figure of merit used to describe the linearity of components including power amplifiers, frequency mixers, switches, ADCs, DACs, and others. This measurement involves the use of a two-tone stimulus and requires the comparison of the power of two fundamental tones with the power of third order distortion products.

ACPR, also known as Adjacent Channel Leakage Ratio (ACLR), is an important metric for components designed for use in wireless standards such as UMTS or 3GPP LTE. This measurement describes the ratio of power in a modulated signal versus power emitted into an adjacent channel. In order to measure ACPR, one must provide the device under test (DUT) with a modulated stimulus – or in the case of a DAC – internally generate a modulated signal on the DUT itself.

2. Theory of Intermodulation Distortion

Intermodulation distortion describes the linearity of a device when presented with a multi-tone stimulus. An ideal component, such as a power amplifier, (PA) should have low noise figure and remain linear - so it does not distort the incoming signals. A weakly non-linear system has a voltage-in to voltage-out relationship that can be described by the power series shown in (1). When operating in the linear range of the DUT, the output voltage closely follows this power series.

$$v_{out} = a_0 + a_1 v_{in} + a_2 v_{in}^2 + a_3 v_{in}^3 + a_4 v_{in}^4 + \dots \quad (1)$$

Consider the behavior of the two-tone third-order intermodulation. If two continuous wave (CW) signals are placed at the input of the DUT, the voltage is represented by (2). The two tones have different radial frequencies denoted as ω_1 and ω_2 .

$$v_{in} = A \cos(\omega_1 t) + B \cos(\omega_2 t) \quad (2)$$

By replacing the expression for the input voltage into (1), the relationship between the input and output signal is more apparent. From the power series, one can easily see that a CW signal at the input generates output distortion signals at $2\omega_1$, $3\omega_1$, and so on. Each harmonic has amplitudes proportional to v_{in}^2 , v_{in}^3 , and so forth. The output voltage is represented by (3).

$$v_{out} = a_0 + a_1 \{ \dots \} + a_2 \{ \dots \} + a_3 \{ A^2 B \cos[(2\omega_1 - \omega_2)t] + AB^2 \cos[(2\omega_2 - \omega_1)t] + \dots \} \quad (3)$$

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The DUT generates many tones due to intermodulation (one tone mixing with the other tone) as shown in **Figure 1**. The intermodulation tones of most concern are the third-order tones at $2\omega_1 - \omega_2$ and $2\omega_2 - \omega_1$ as shown in **Figure 2**.

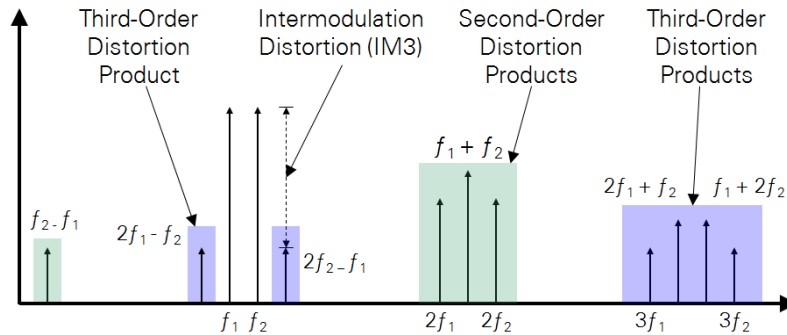


Figure 1: Intermodulation Tones Caused by Non-linear DUT

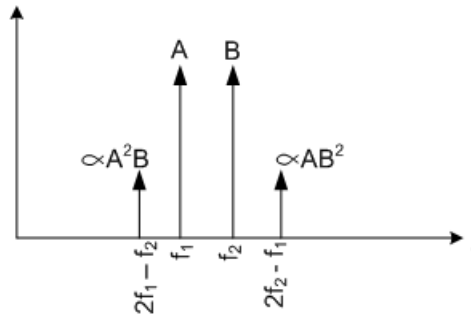


Figure 2: Two-Tone Third-Order Intermodulation Distortion

These distortion products are spaced too close in frequency to be effectively filtered. When considering IP3 performance, the fundamental tones powers are set to the same power level. From **Figure 2**, one can derive that a 1 dB change in the fundamental tone results in a 3 dB change in the third-order distortion products. When viewed on a log-power out versus log-power in graph, the fundamental tones powers increase with a slope of 1. The third-order distortion products increase with a slope of 3 as shown in **Figure 3**.

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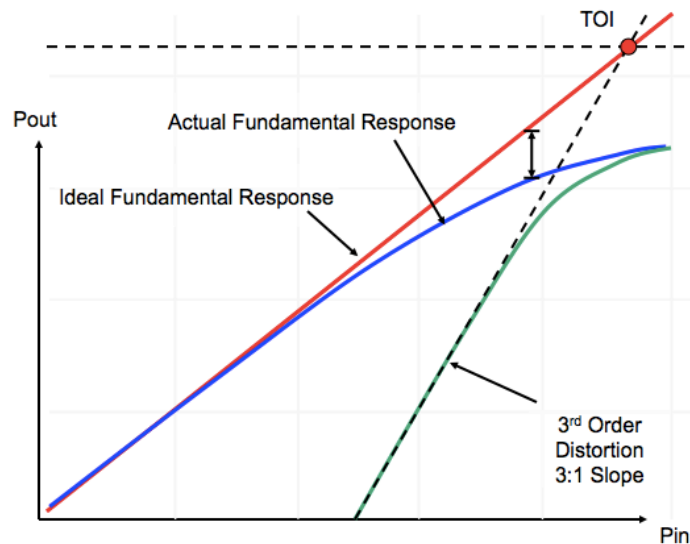


Figure 3: Power out versus Power in Plot for Fundamental versus Third-Order Intermodulation Distortion

At a theoretical power level, the fundamental and distortion level lines intersect. This intersection is referred to as the third-order intercept point (IP3 or TOI). However, the intersection is fictitious – it is realized only in mathematics and used as a figure of merit in industry. In practical use, a device goes deeply into gain compression well before the input power approaches the IP3 level. Moreover, the simple power series model that was expressed in (1) breaks down at higher power levels. In practice, the empirical formula to calculate IP3 is shown in (4), where P_{IMD3} is the third-order intermodulation distortion power.

$$IP3_{in} = \frac{3}{2} P_{in}(dBm) - \frac{1}{2} P_{IMD3}(dBm) \quad (4)$$

IP3 can be expressed as $IP3_{in}$ or $IP3_{out}$. For example, mixers use $IP3_{in}$ and amplifiers use $IP3_{out}$. The decision to specify either the input or output IP3 is typically driven by marketing and industry standards. The relationship between $IP3_{out}$ and $IP3_{in}$ is dictated by the gain (dB) of the DUT. For example, convention dictates that the linearity of a high-gain device is specified by its output IP3 and that the linearity of a passive device such as a mixer is specified by its input IP3.

3. Optimizing IP3 Measurements

Although IP3 measurements require complex instrumentation configurations, one can ensure the accuracy of these measurements using a few simple best practices. When measuring IP3, one must ensure that the distortion products of both the sources and the signal analyzer are significantly smaller than the distortion products introduced by the DUT. The lower the inherent IP3 introduced from the actual measurement setup, the more accurate the IP3 number. As a result, one must take care to ensure a low distortion signal source and optimize VSA settings such as reference

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level, resolution bandwidth, and attenuation. As shown in **Figure 4**, the IP3 measurement requires two RF signal generators and a power combiner to produce a high-quality two-tone stimulus.

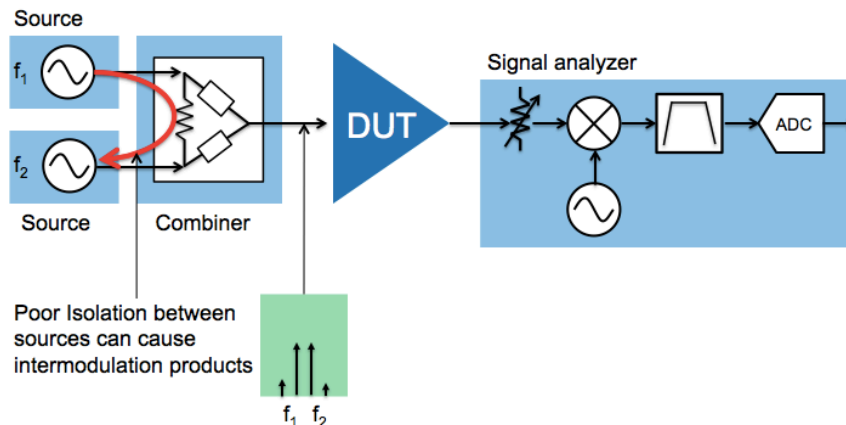


Figure 4: IP3 Measurement Setup Including Unwanted Transmission from Lack of Source Isolation

Source isolation is sometimes critical to making an accurate IP3 measurement. When using a power combiner, one must provide sufficient source isolation because without it, energy from one source can leak into the other source as shown in **Figure 5**. This leakage tone creates AM sidebands at the difference frequency, which falls at the same frequency as the measurement distortion products.

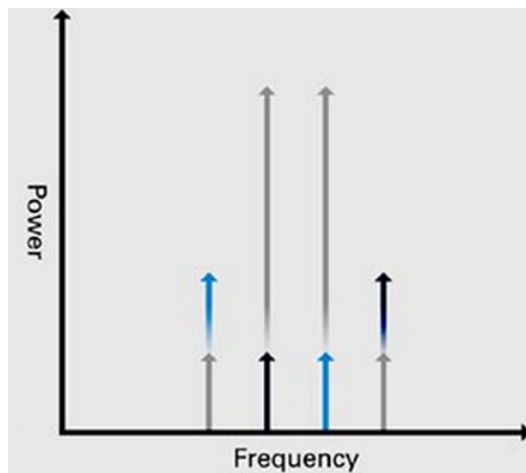


Figure 5: Distortion Caused by a Lack of Source Isolation

One way to improve the isolation is to choose a combiner with the best port-to-port isolation. Generally, purely resistive combiners feature only between 6 dB and 12 dB of isolation, depending on the resistor topology. Additional measures to increase the isolation using either directional couplers or isolators can provide up to 50 dB of isolation if used at both ports. However, couplers are often limited to a single octave and are thus not suitable for broadband applications.

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One solution to the isolation challenge is to place an attenuator at the output of each signal generator. The attenuator provides additional isolation for signals traveling in the reverse direction. A good rule of thumb is that roughly 40 dB of isolation is required to measure up to IP3 numbers above +25 dBm. Another option is to place an amplifier at the output of the sources. The amplifier not only provides the extra gain needed to produce high power signals, but also offers inherently good isolation. It is important to note that the amplifier must exhibit high linearity so excess distortion isn't introduced into the measurement.

Optimizing VSA Linearity

In addition to ensuring isolation between sources in the stimulus signal, one must also configure the VSA to use optimal linearity settings. As shown in Figure 6, the signal-to-noise ratio (SNR) and internally generated distortion products are related to the mixer level of the VSA. Typically, the mixer level is optimized for the best dynamic range automatically by adjusting the front-end attenuation/gain when the reference level is set. However, to manually adjust the mixer level, one can set the attenuation to achieve the desired response.

The ideal setting (point C on **Figure 6**) for the VSA occurs at the point where the mixer level is high enough to reduce the noise floor but also low enough that the internally generated distortion points do not interfere with the measurement. Point A on **Figure 6** shows a setting where the measurement is limited by the VSA noise floor. Essentially, more gain is required, which will reduce the noise floor enough to resolve the intermodulation tones. Point B on **Figure 6** demonstrates the case where the downconverter has too much gain and more attenuation is required to reduce the internally generated VSA distortion.

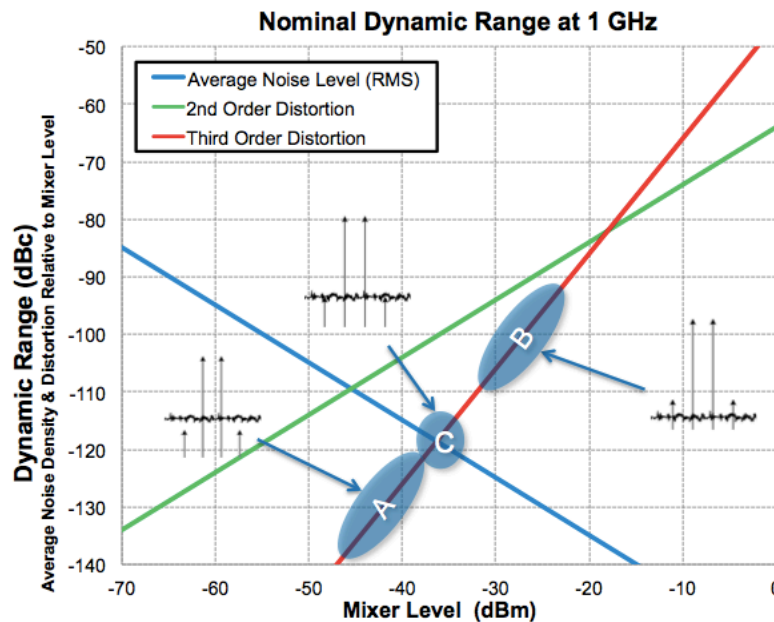


Figure 6: Dynamic Range Chart of the PXIe-5668R VSA

An easy way to determine if the VSA is contributing to the IP3 measurement is to increase the front-end RF attenuation and to observe whether the distortion decreases. If the distortion stays constant, then one can be certain that the distortion products are coming from the DUT. On the other hand, if the observed distortion decreases, one

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can infer that the signal analyzer distortion decreased because the signal level at the mixer is reduced. This technique, known as “rolling the step attenuator,” identifies the source of distortion. While it is important to set the reference level and attenuation appropriately, one can also reduce the effective noise floor by reducing the resolution bandwidth (RBW). **Figure 7** shows a TOI measurement being performed by the RFSA Soft Front Panel.

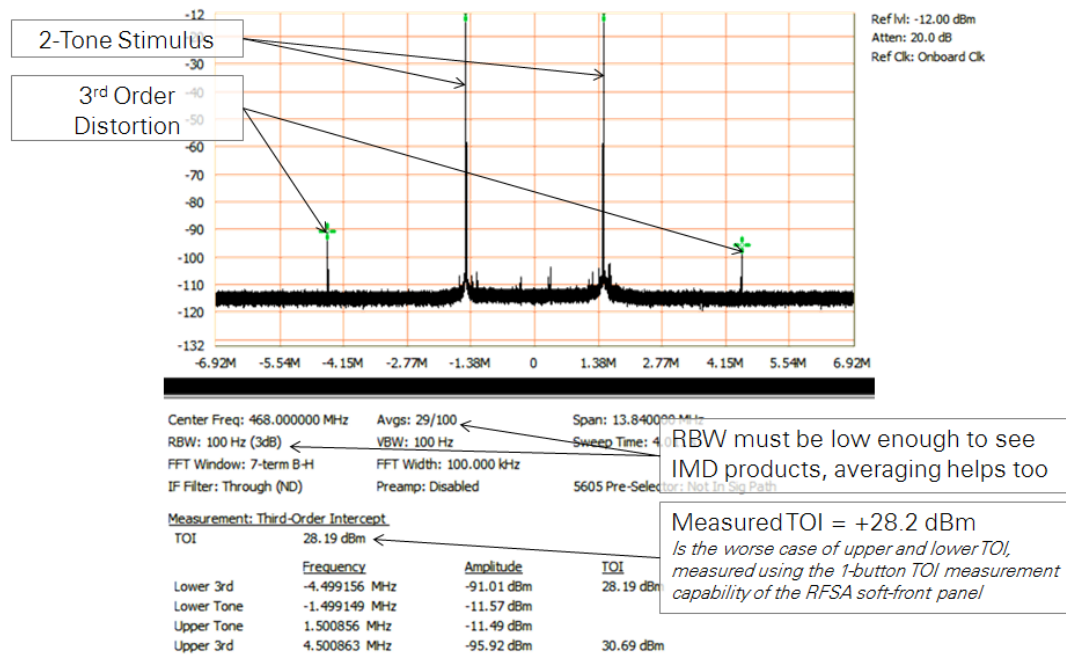


Figure 7: Anatomy of a TOI Measurement using the RFSA Soft Front Panel

Equation 5 shows how the noise floor relates to the noise figure and the RBW of the VSA. However, by decreasing the RBW, the measurement time increases. VSAs with inherently low noise figures help to minimize the measurement accuracy to speed tradeoff.

$$\text{Noise Floor} = -174 \frac{\text{dBm}}{\text{Hz}} + NF_{\text{VSA}} + 10 * \log_{10}(\text{RBW}) \quad (5)$$

VSA IF Conditioning

Another consideration when setting the VSA is the IF signal conditioning. Typically, one configures the gain of the analyzer to use an IF power that is slightly less than full scale of the analog-to-digital converter (ADC) to prevent clipping. One can improve the noise floor of the instrument by setting a narrow IF bandwidth and increasing the IF power level.

To accomplish this, one must space the distortion tones so they exceed the bandwidth of the VSA’s IF filter. By filtering out the two-tone stimulus, one also reduces the distortion products internally generated by the ADC, which allows for a more accurate IP3 measurement. **Figure 8** depicts a dynamic range chart including the ADC distortion versus VSA mixer level and **Figure 9** shows a graphical representation of this optimization technique.

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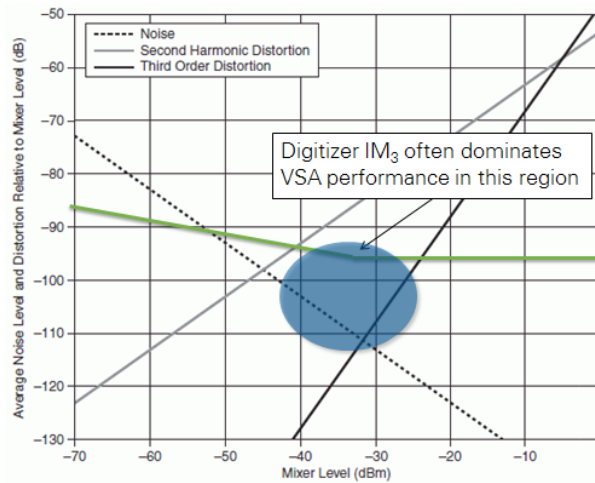


Figure 8: Dynamic Range Chart Including ADC Distortion

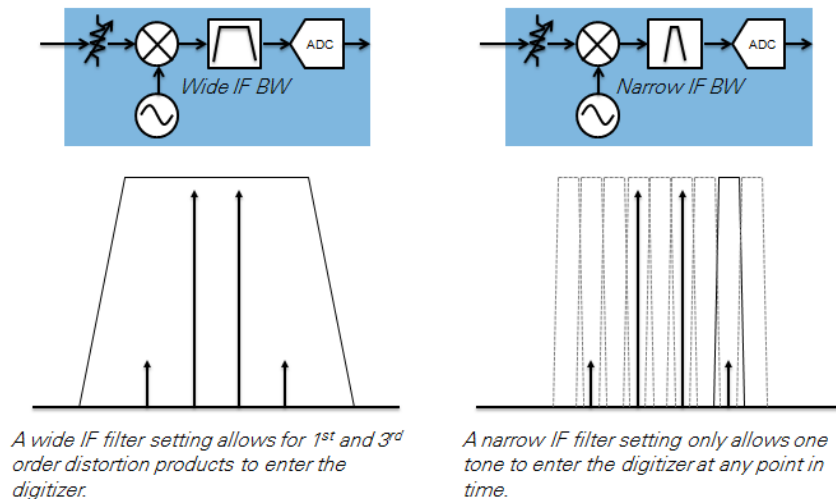


Figure 9: Graphical Representation of IF Filter Settings

Another valuable technique to improve the IP3 measurement is to apply IF gain re-ranging. The previous dynamic range charts in this paper describe a VSA setting with a constant reference level across the entire measurement span. With a narrow IF filter, one can apply more gain (lower reference level) while measuring the IMD3 tones to significantly improve the noise floor and decrease the inherent distortion. Since the two fundamental tones aren't present at the ADC, gain can be applied without overloading the backend. **Figure 10** shows this re-ranging technique where the appropriate reference level is still applied to the fundamental tones.

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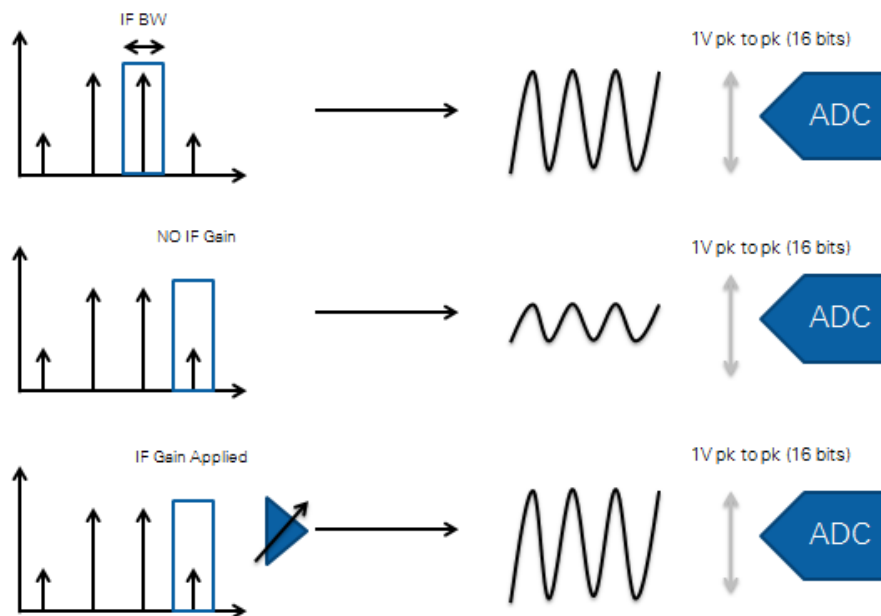


Figure 10: IF Gain Re-ranging to Improve the Noise Floor and Decrease the Inherent Distortion

IP3 is an important figure of merit used by industry to describe the linearity of a DUT, and it presents many challenges. While measuring IP3, careful consideration of the measurement setup including sufficient source isolation and optimized VSA settings is critical for accurate results. The ability to optimize an IP3 measurement is an appropriate transition to describing ACPR theory.

4. Theory of Adjacent Channel Power Ratio

The adjacent channel power ratio of a wireless communication signal describes the ratio between the integrated power in the carrier channel relative to the adjacent channel. ACPR is the wideband equivalent to the CW IP3 case. The adjacent channel power ratio of a wireless communication system is the integrated power in the carrier channel relative to the adjacent channel. ACPR is the wideband equivalent to the CW IP3 case. **Figure 11** depicts this specification graphically. When a modulated wideband signal is incident on a non-linear DUT, spectral re-growth will occur. This re-growth has the unintended consequence of leaking into the adjacent channel, which can cause over-the-air interference within a different carrier's channel bandwidth.

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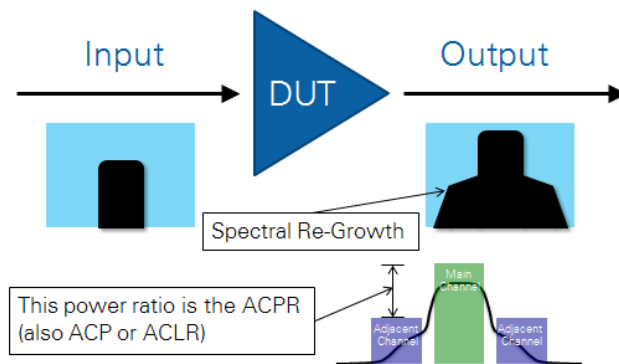


Figure 11: Graphical Depiction of ACPR

One technique to model spectral re-growth of a weakly non-linear system, such as the VSA when operating far below its gain compression limits, is to divide up the digitally modulated signal into equal frequency slices. Represent each slice as a CW whose average power is the average power of that slice. As described above, the tones will create intermodulation tones at predictable frequencies and amplitude levels.

Using the CW IP3 behavior, one can model the spectral re-growth by adding the amplitudes as voltages (assuming the phases of the individual contributors line up). To increase the model accuracy, one can continue to decrease the frequency slice bandwidths, which increases the number of CW tones. Essentially, by taking the limit as the BW of each slice approaches zero, the model accuracy improves. Primarily, the intermodulation is third-order, but fifth and even seventh-order products can affect ACPR. **Figure 12** shows the modeled spectral re-growth as a discrete spectrum of CW tones.

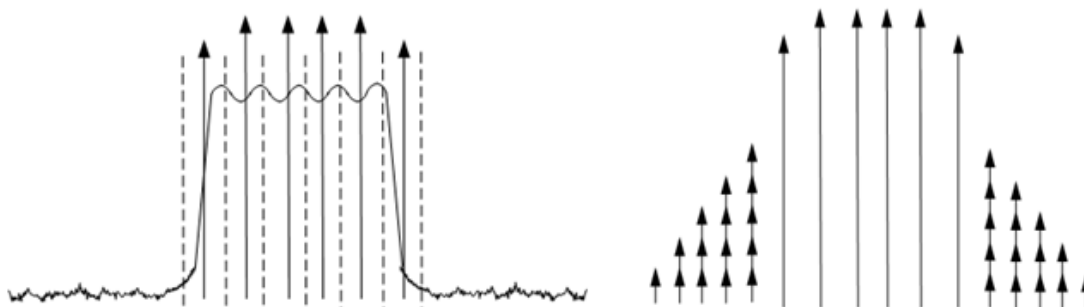


Figure 12: Spectral Re-growth Modeled as Intermodulation Tones

Through this exercise, it's apparent that the linearity of a device or system directly impacts the ACPR performance. As shown in **Figure 13**, other factors include broadband noise and phase noise.

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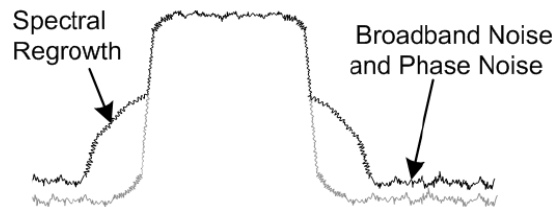


Figure 13: Broadband Noise and Phase Noise Contribution to ACPR

Although factors such as phase noise theoretically can limit the ACPR specifications of measurement devices such as a signal analyzer, modern high-performance signal generators and analyzers typically exhibit phase noise performance that well exceeds what is required for measuring ACPR on modern communications signals.

5. Optimizing ACPR Measurements

Similar to the IP3 measurement case, IF filtering plays a critical role in improving the ACPR measurement performance. Although the source is a digitally modulated wideband signal, by decreasing the IF filter and increasing the IF gain, one can achieve better ACPR numbers. The RFSA software will collect data in multi-span FFT acquisitions and stitch the spectrums together. Like before, the tradeoff for narrowing the IF is measurement speed. There is a proportionate increase in acquisitions that must happen to cover the wideband signal. **Figure 14** demonstrates this technique graphically.

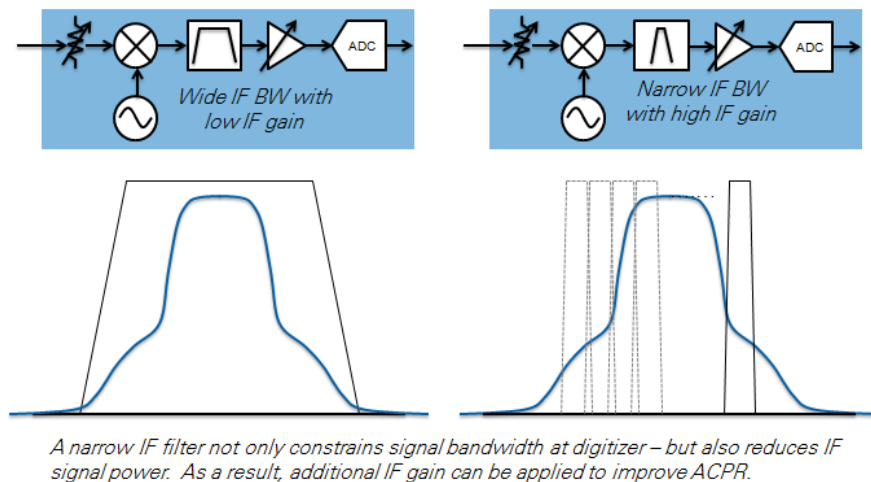


Figure 14: Graphical Depiction of ACPR Measurement with Narrow IF Filtering

Another technique that can be applied to increase the ACPR measurement performance of the VSA is noise correction. When configured to use noise correction, the VSA first measures its internal noise floor, and caches this value. It then subtracts this cached “noise power” from resulting spectrum measurements to more accurately reflect the noise power of the actual signal. As a result of noise correction, one can reduce the inherent ACPR of the VSA

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by more than 7 dB. Note that additional measurement time is required to perform the computation necessary to measure and subtract the VSA's inherent noise floor. **Figure 15** shows both the inherent VSA noise and actual noise at the VSA input.

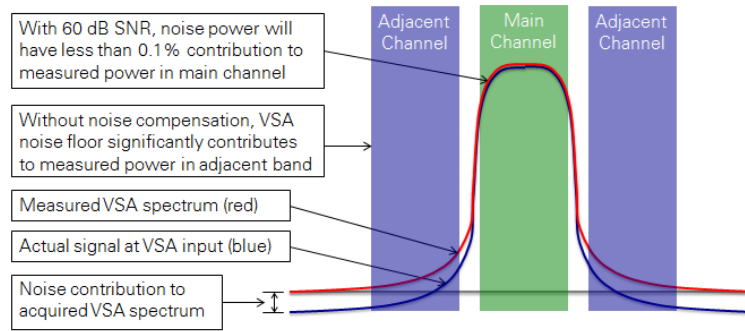


Figure 15: Graphical Depiction of Noise Correction

The modulation signal is noise-like so the power spectral density of both the signal and the noise will scale by the same amount as a function of RBW setting. For this reason, SNR is independent of RBW setting. By decreasing the RBW, the main channel power will decrease proportionally with the adjacent channel noise. Therefore, even with the optimization techniques mentioned above for IP3 and ACPR, the ACPR measurement capability of a VSA distills down to the dynamic range of the receiver. Best-in-class VSAs like the PXIe-5668R are capable of measuring ACPRs above 80 dB.

Figure 16 shows the PXIe-5668R with the RFSA Soft Front Panel measuring the ACPR of a WCDMA signal without narrow filtering and without noise correction. The main channel is centered at 468 MHz with a channel bandwidth of 3.84 MHz.

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Figure 16: PXIe-5668R Measuring WCDMA ACPR with 100 MHz Wide IF Filter and Noise Correction Disabled

By decreasing the reference level, the VSA’s ACPR measurement performance increases if the signal measurement is noise limited. However, at a certain point distortion created by the VSA results in spectral re-growth as shown in Figure 17.

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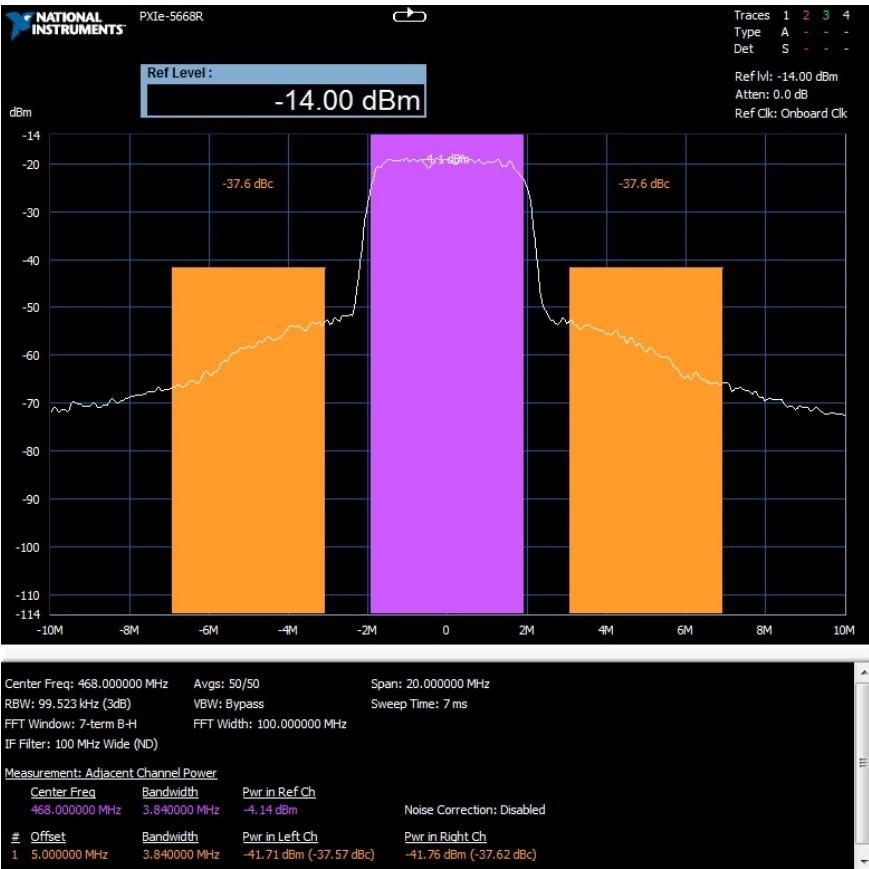


Figure 17: PXIe-5668R Measuring WCDMA ACPR showing ADC Distortion

Figure 18 shows the PXIe-5668R with the RFSA Soft Front Panel measuring the ACPR of a WCDMA signal with the 300 kHz narrow IF filter and without noise correction. By reducing the bandwidth of the analog IF filter, one can now increase the IF gain by lowering the reference level to improve the ACPR capability.

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Figure 18: PXIe-5668R Measuring WCDMA ACPR with 300 kHz Wide IF Filter and Noise Correction Disabled

Figure 19 shows the PXIe-5668R with the RFSA Soft Front Panel measuring the ACPR of a WCDMA signal with the 300 kHz narrow IF filter and with noise correction. Since the receiver is still noise limited, one can now enable noise correction to reveal the source’s noise. It is important to note that IF gain re-ranging can also be applied for these measurements. The IF gain can be increased within the adjacent channel bandwidth to improve the VSA dynamic range. Also, to increase measurement speed, a larger bandwidth IF filter can be used in the adjacent channel without performance degradation as long as the narrow IF filter is used within the main channel. The RFSA Soft Front Panel does not perform this re-ranging, so additional ACPR performance can be achieved with the appropriate implementation.

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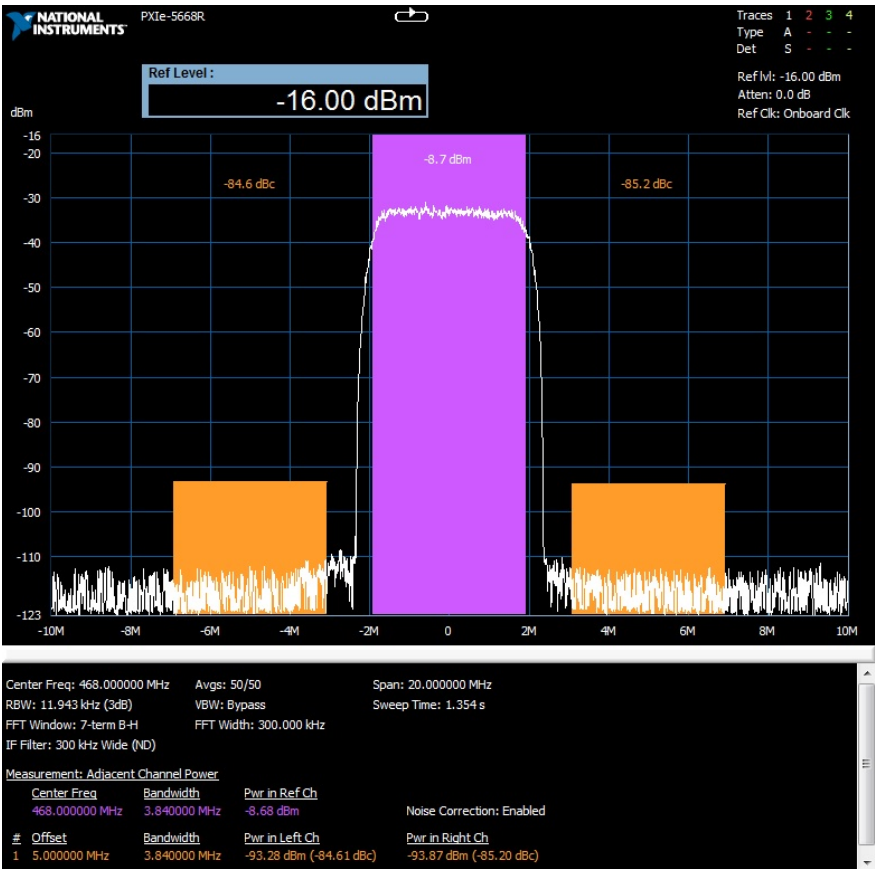


Figure 19: PXIe-5668R Measuring WCDMA ACPR with 300 kHz Wide IF Filter and Noise Correction Enabled

In accordance with theory, the measurement ability increased with a narrow IF filter and with noise correction enabled. The table below summarizes the ACPR WCDMA measurement capabilities of the PXIe-5668R using the RFSA Soft Front Panel at 468 MHz.

IF Filter BW	Noise Correction	ACPR (dBc)
100 MHz	Disabled	-75.0
300 kHz	Disabled	-77.4
300 kHz	Enabled	-84.6

Table 1: Summary of PXIe-5668R ACPR WCDMA Measurement Capabilities Using RFSA Soft Front Panel

As devices continue to advance, it will become more difficult to measure linearity performance and even more important to invest in best-in-class instruments to measure these sensitive but important specifications. The PXIe-5668R High-Performance Vector Signal Analyzer offers best-in-class performance up to 26.5 GHz for linearity measurements.