

# Intermodulation Distortion Measurements on Modern Spectrum Analyzers

## Application Note

### Products:

- | R&S®FSV
- | R&S®FSW

This Application Note describes the differences in intermodulation distortion measurements between traditional spectrum analyzers with analog narrow-band IF signal path and modern spectrum analyzers using a wide-band IF signal path and digital RBW filters.

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# 1 Introduction

Intermodulation distortion (IMD) plays an important role for many spectrum analyzer based measurements. Mostly spectrum analyzers are used to measure IMD of a Device Under Test (DUT). However, a spectrum analyzer may also contribute to IMD due to the non-linear behavior of its different active components used in its signal path. The most critical intermodulation distortion product is the Third Order Intermodulation (TOI) product. It is generated when two signals are applied to a nonlinear device. The TOI products are usually within or close to the use-band, i.e. it can hit a neighboring channel in a communication system. This application note gives an overview which elements contribute. It also includes recommendations for the optimum level setting to minimize spectrum analyzer generated IMD.

Chapter 2 explains the theoretical background of intermodulation distortion in more detail.

Chapter 3 compares the differences in architecture between the analog narrow-band spectrum analyzer approach and a modern wide-band signal- and spectrum-analyzer.

Chapter 4 focuses on TOI measurements of a DUT in practice. A typical measurement setup is shown and the influence of signal generators and the spectrum analyzer on the measurement is explained. This chapter also gives recommendations how to avoid measurement errors resulting from intermodulation products of the signal generators or spectrum analyzers.

Readers who are interested in a guideline on how to obtain accurate IMD measurement results are welcome to proceed directly to chapter 4.

## 2 Intermodulation Distortion

Intermodulation distortion originates from non-linearities in electronic circuits. Non-linear elements generate harmonic signals, or in more detail: non-linear elements create signals at multiples of the input signal. Chapter 2.1 describes the mathematical background on harmonic signals caused by non-linear elements, whereas chapter 2.2 introduces intermodulation as a result of harmonics from non CW signals.

### 2.1 Harmonic signals

This section will show the basic equations for harmonics created in a single tone scenario.

Given the case that a single CW tone is applied to a non-linear element, additional signals, the so-called harmonics, will be generated at  $n$  times of the original frequency, with  $n$  being the order of the harmonic.

Any non-linear element can be described by a Taylor-series

$$P(s) = a_0 + a_1 \cdot s + a_2 \cdot s^2 + a_3 \cdot s^3 + \dots \quad (1)$$

with  $P(s)$  being its transfer function and  $s$  being the input signal. We will not look in detail on the factors  $a_n$ , but focus on the powers of  $s$ .

Assuming a CW input signal, the general formula for a signal  $s$  as a function of time  $t$  is

$$s(t) = B \cdot \cos(2\pi \cdot f \cdot t + \varphi). \quad (2)$$

Using the addition theorem for the cosine function, it is straight forward to figure out that the square term in Eq. (1) creates a signal with twice the original frequency (the second harmonic), the cube term the third harmonic and so on.

For a more detailed explanation of these formulas, please refer to the Rohde & Schwarz Application Note 1EF78, or the internet.

## 2.2 Intermodulation as a result of harmonic signals

Clearly, harmonics of a single tone are outside the usable band of an application, since they are at multiples of the original frequency. Once a second tone joins the input signal at a small frequency offset – the resulting output signal looks different. In contrast to the single tone scenario above, the signal  $s$  is now

$$s(t) = B_1 \cdot \cos(2\pi \cdot f_1 \cdot t + \varphi_1) + B_2 \cdot \cos(2\pi \cdot f_2 \cdot t + \varphi_2). \quad (3)$$

Since the dominating intermodulation products are the third order products, the following equations focus only on those. Calculating the third power terms (responsible for the third order intermodulation and third order harmonics) of the Taylor series (Eq. 1) with the two tone input signal from Eq. 3 yields the following result:

$$\begin{aligned} s^3(t) = & B_1^3 \cdot \cos^3(2\pi \cdot f_1 \cdot t + \varphi_1) + \\ & B_2^3 \cdot \cos^3(2\pi \cdot f_2 \cdot t + \varphi_2) + \\ & 3 \cdot B_1^2 \cdot B_2 \cdot \cos^2(2\pi \cdot f_1 \cdot t + \varphi_1) \cdot \cos(2\pi \cdot f_2 \cdot t + \varphi_2) + \\ & 3 \cdot B_1 \cdot B_2^2 \cdot \cos(2\pi \cdot f_1 \cdot t + \varphi_1) \cdot \cos^2(2\pi \cdot f_2 \cdot t + \varphi_2) \end{aligned} \quad (4)$$

The first two lines describes the third order harmonics for each of the input tones ( $\cos^3$ -terms), whereas lines 3 and 4 represent the third order intermodulation terms (mixed terms). From the above equations, the TOI frequencies can be derived using the addition theorem (for trigonometric functions) as

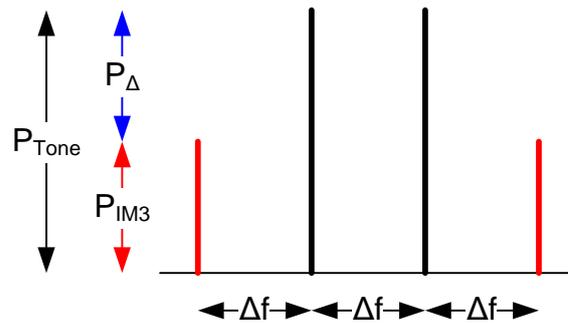
$$\begin{aligned} f_{\text{TOI1}} &= 2 \cdot f_1 - f_2 \\ f_{\text{TOI2}} &= 2 \cdot f_2 - f_1 \end{aligned} \quad (5)$$

While the 3<sup>rd</sup> order harmonics ( $3 \cdot f_1$  and  $3 \cdot f_2$ ) of the individual input tones can be easily suppressed by a low-pass filter, the third order intermodulation terms are often more critical for the application. The resulting frequencies are often in-band for a given application and therefore interfere with the wanted signal. Additionally, under the assumption  $B_1 = B_2$ , i.e. both tones have the same level, the intermodulation terms exceed the harmonic terms by a factor of 3 in amplitude (Eq. 4). A factor of 3 in amplitude converts to a 9.54 dB difference between the third order harmonics of the individual tones and the third order intermodulation products.

## 2.3 Characterizing IMD

There is a number of ways to visualize intermodulation distortion. Fortunately, the measurement method is identical and the results can be converted.

The measurement method used to characterize the IMD behavior of a DUT is the so-called two tone scenario. Two continuous wave (CW) tones with equal tone power ( $P_{\text{InTone}}$ ) and spaced by a given frequency ( $\Delta f$ ) are applied to the DUT input (see Figure 1). On the output side, the power level of the original tones may have changed to  $P_{\text{Tone}}$ . The intermodulation products can be measured with their absolute power or their relative power related to  $P_{\text{Tone}}$ , referred to as  $P_{\Delta}$ . In practice  $P_{\Delta}$  is also called intermodulation free dynamic range. Clearly, the 3<sup>rd</sup> order intermodulation tones have the same spacing to the upper and lower tone as the two original tones have ( $\Delta f$ ).

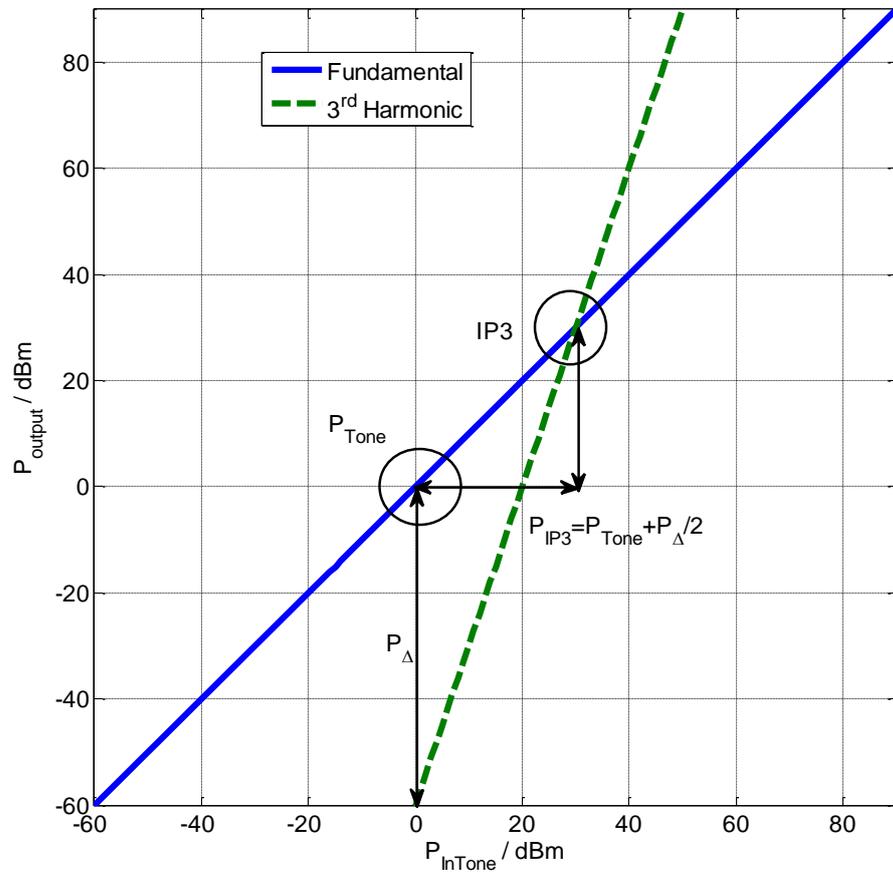


**Figure 1: Two tone scenario used for IMD testing**

Additionally, the so-called third order intercept point (IP3) can be calculated. It is a theoretical point, where the intermodulation products at the DUT's output grow as large as the original tones at the DUT output side. The IP3 can be derived on a logarithmic scale (i.e. all values in dBm or dB) as

$$IP3 = P_{\text{Tone}} + P_{\Delta} / 2. \quad (6)$$

Figure 2 shows graphically the relation of Eq. 6. It shows the theoretical lines of the fundamental and 3<sup>rd</sup> harmonic at the output of a 0 dB gain DUT.

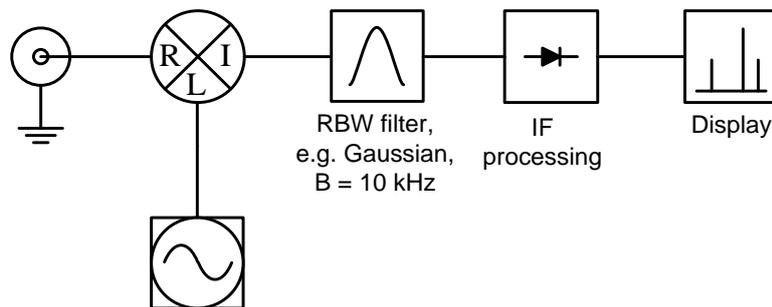


**Figure 2:** Graphical representation of Eq. 6, distance of the intermodulation products  $P_{\Delta}$  from the signal  $P_{\text{Tone}}$ , depending on the third order intercept point IP3

## 3 Comparison of Architectures

Chapter 2 introduced intermodulation as a result of non-linear elements. Clearly, when frequency band limitations such as band-pass filters come into the game, only those spectral components are to be considered for intermodulation, which are located within the filter bandwidth. This fact will turn out to be the main difference between analog swept spectrum analyzers and modern wide-band signal- and spectrum analyzers.

### 3.1 Traditional swept spectrum analyzers



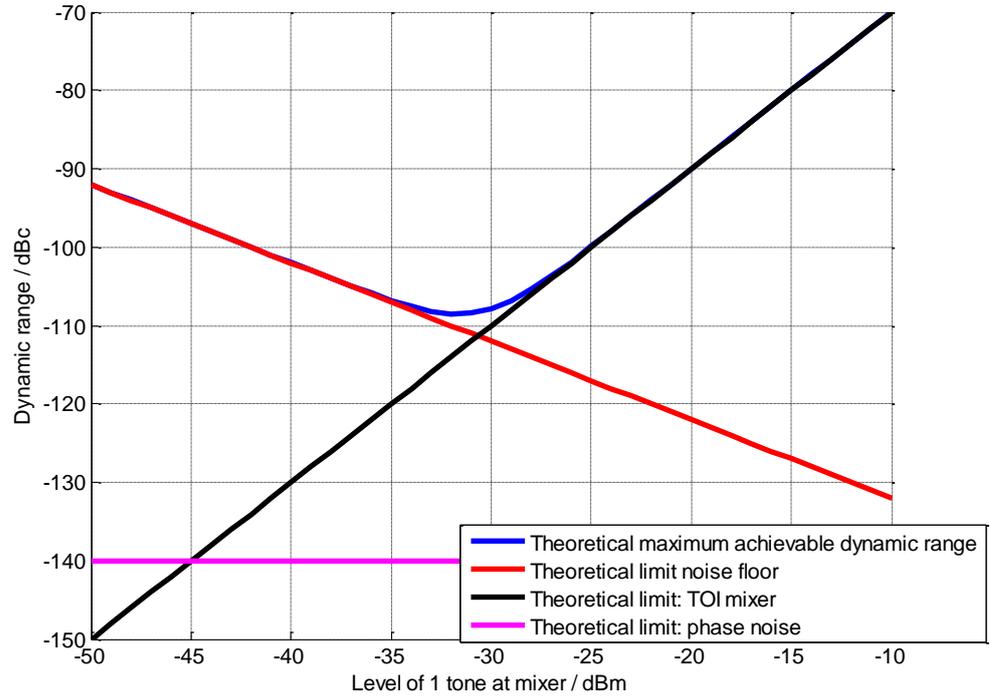
**Figure 3: Simplified block diagram of an analog swept spectrum analyzer. Diagram applies in principle also for spectrum analyzers with narrow-band digital back ends, as e.g. the R&S FSP and R&S FSU.**

The block diagram in Figure 3 shows the key elements that contribute to inherent intermodulation on a spectrum analyzer. The first component in the above block diagram is the input mixer. Since we assume no attenuation or amplification in front of the input mixer in this section, the mixer level is equal to the input level for now. For more details on the effect of attenuators and amplifiers in front of the input mixer, see section 3.4.

For increasing mixer levels (dependent on the mixer type, for the R&S FSU starting at approx. -30 dBm), the mixer intermodulation products start dominating the total spectrum analyzer intermodulation distortion. Starting from that mixer level, the spectrum analyzer will follow the TOI specifications of the mixer used in the design. The data sheet specification of the TOI of the spectrum analyzer is given for an input level where the mixer clearly dominates (e.g. -10 dBm for the R&S FSU).

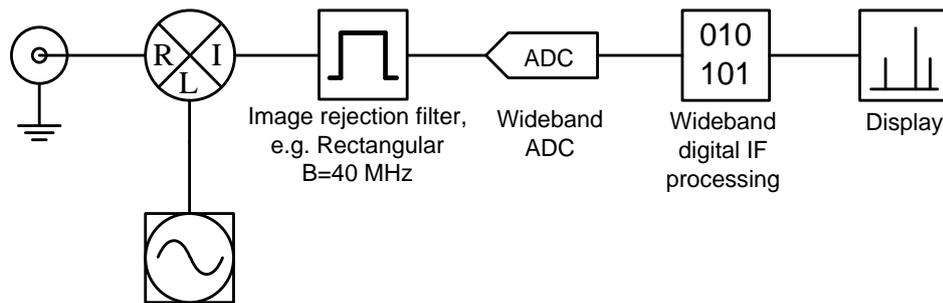
From the simplified block diagram (Figure 3) it can be concluded that the IF processing elements do not "see" the input tones simultaneously, since the analog RBW filter in use will in general have a lower bandwidth than the tone spacing. Therefore, the RBW filter prevents the IF signal chain from contributing to inherent intermodulation.

Figure 4 displays graphically how noise floor and TOI specification of a spectrum analyzer add up to limit the theoretically achievable dynamic range, depending on the input level. Up to approx. -35 dBm, the dynamic range increases by 10 dB with every 10 dB increase of signal level, simply because the noise floor stays constant. Above approx. -30 dBm of mixer level, the mixer TOI limit dominates, reducing the dynamic range by 20 dB for every 10 dB increase of signal level. Since the phase noise specification is well below the other limiting factors, it has no practical influence on this instrument (example: R&S FSU).



**Figure 4: Theoretical limits for the intermodulation free dynamic range of a spectrum analyzer. Typical specification of an R&S FSU is used with 10 Hz RBW**

### 3.2 Modern wide-band (signal- and) spectrum analyzers



**Figure 5: Simplified block diagram of a wide-band digital back end spectrum analyzer, as e.g. the R&S FSW.**

The important difference between Figure 3 and the simplified block diagram in Figure 5 is the IF filter used. Spectrum analyzers with narrow-band IF use RBW filters, which are implemented as analog filters or as a combination of analog and subsequent digital filters. For narrow RBWs the signal bandwidth in front of the A/D converter is therefore not more than a few kHz.

Modern signal and spectrum analyzers use only digital RBW filters. The wide-band approach does not only give these analyzers a huge speed advantage but also the possibility to demodulate wide-band signals. The wide IF analyzers still have analog filters in the IF chain, mainly designed for image rejection. Typically, wide-band signal- and spectrum analyzers have two to three of these analog filters with different bandwidths, e.g. 5 MHz, 17 MHz, and 80 MHz on the R&S FSW and 5 MHz and 40 MHz on the R&S FSV.

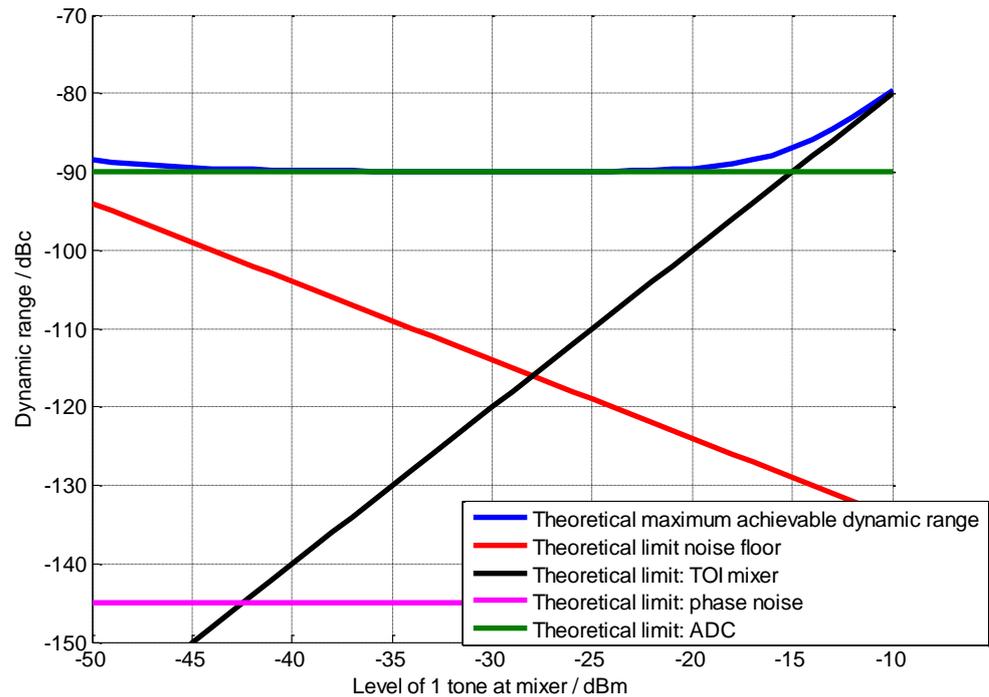
**Note:** All block diagrams in this application note show the wide-band digital spectrum analyzer with an analog IF filter with an ideal rectangular shape. In reality this filter shape is of course not rectangular, but it was chosen here to visualize the difference to the Gaussian shaped RBW filters. The key figure in terms of influence on inherent intermodulation is not the filter shape, but the filter bandwidth.

As long as the tone spacing of a signal is wider than the analog filter bandwidth in use, there is no difference in behavior compared to traditional spectrum analyzers. But with a most narrow analog filter bandwidth of 5 MHz, the IF chain of the analyzer will see both tones of the test scenario at the same time, if the tone spacing is less than 5 MHz. Figure 6 visualizes the difference between a narrow-band and a modern wide-band spectrum analyzer related to its IF filter bandwidth.



**Figure 6:** Two tone scenario, drawn with analog IF filters. Left hand side: narrow-band RBW filter, right hand side: wide-band image rejection filter.

As a consequence, non-linear elements in the IF chain following the analog filter will "see" both tones at the same time and therefore may contribute significantly to the inherent intermodulation distortion of the spectrum analyzer. Especially analog to digital converters (ADCs) contribute to intermodulation significantly at low power levels. The theoretical maximum achievable intermodulation free dynamic range is in this case of course influenced by the specification of the ADC. Figure 7 shows the influence of an ADC on the maximum achievable dynamic range, which is now influenced by the ADC over a wide input level range. At about -20 dBm, the mixer's intermodulation products start dominating over the ADC influence, whereas the input signal has to be below -50dBm, before any influence of the noise floor becomes visible. Details on how the ADCs influence the intermodulation behavior can be found in subsection 4.3.4.



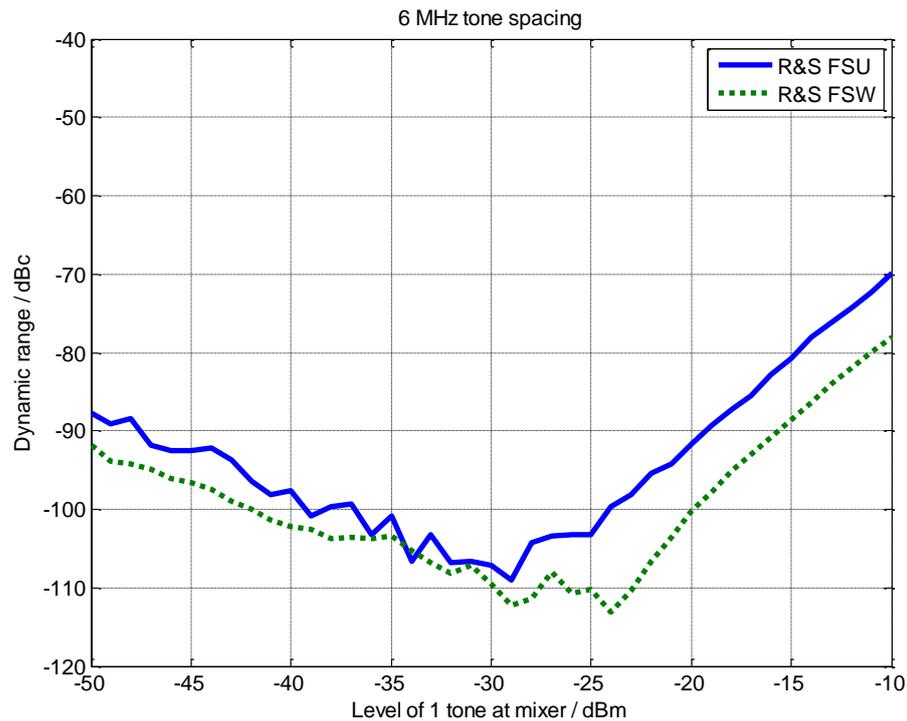
**Figure 7: Theoretical limits for the intermodulation free dynamic range of a wide-band spectrum analyzer with a test signal having a narrow tone spacing. Typical specification of an R&S FSW is used, with 10 Hz RBW**

### 3.3 Comparison

Figure 8 and Figure 9 compare the intermodulation free dynamic range of a narrow-band spectrum analyzer with a wide-band spectrum analyzer. Although the R&S FSU and R&S FSW were used in these measurements, the results are representative for any spectrum analyzer following the narrow- or wide-band signal path concepts.

The figures below show the following behavior:

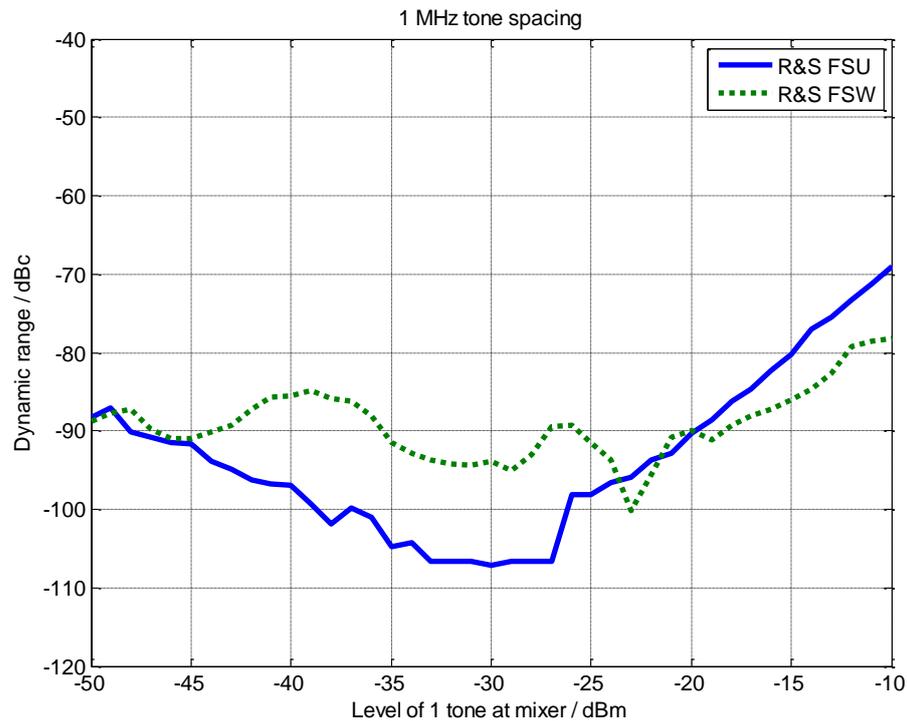
- For a tone spacing wider than the analog IF bandwidth (Figure 8), the R&S FSW performs better than the R&S FSU over the entire level range. The principal behavior, resulting in the typical “V” shape is identical for both instruments. The R&S FSW shows an intermodulation free dynamic range of approx. 110 dB from about -30 dBm to about -24 dBm mixer level.



**Figure 8: Comparison of intermodulation free dynamic range depending on the mixer level for the wide-band R&S FSW and the narrow-band R&S FSU, tone spacing 6 MHz, i.e. larger than the most narrow analog filter on the R&S FSW,  $f=1600$  MHz, 10 Hz RBW, no noise correction**

- For a tone spacing lower than the analog IF bandwidth (Figure 9), the dynamic range varies between 85 dB and 100 dB. The exact shape of the curve depends on the characteristics of the ADC in use and may therefore be different for instruments of different vendors or different spectrum analyzer families. Above -20 dBm, the influence of the input mixer dominates, so the total inherent IMD behavior is again comparable to that of traditional spectrum analyzers.

As a summary, the measured curve has the traditional "V" shape for wide-band spectrum analyzers, as long as the signal spacing is wider than the used analog IF filter bandwidth (Figure 8). As soon as the tone spacing is narrow, the curve will deviate from the theoretical "V" shape (Figure 9) and resemble the shape predicted in Figure 7.

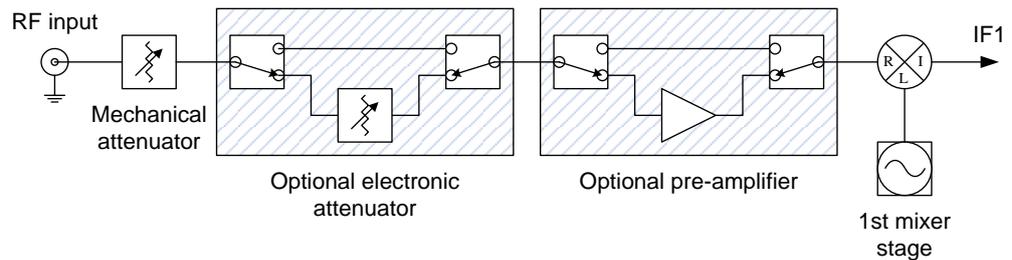


**Figure 9:** Comparison of intermodulation free dynamic range depending on the mixer level for the wide-band R&S FSW and the narrow-band R&S FSU, tone spacing 1 MHz, i.e. smaller than the most narrow analog filter on the R&S FSW,  $f=1600$  MHz, 10 Hz RBW, no noise correction

### 3.4 Impact of electronic attenuators and RF preamplifiers

A spectrum analyzer is equipped with one or more components that allow control of the mixer level, independent of the level applied to the RF input port. All analyzers are equipped with an input step attenuator. It allows attenuating the RF input level. The attenuators often have a range from 0 dB to 75 dB. As it is a linear passive component, a mechanical attenuator does not contribute to the inherent intermodulation of the analyzer. The issue with a mechanical attenuator is its lifetime due to wear and tear, which depends on the number of switching cycles applied.

For applications that require many switching cycles of the input attenuator, spectrum analyzers offer electronic attenuators in addition. Due to the absence of moving mechanical parts, the number of switching cycles is irrelevant for these components. When using electronic attenuators for intermodulation measurements, the TOI specification of the electronic attenuator should be kept in mind, in order to avoid another source of inherent IMD. The electronic attenuator should be bypassed, if its IMD specification comes close or exceeds the expected IMD of the DUT. The optional electronic attenuators of the R&S FSW and the R&S FSU both have specifications that meet or exceed the base unit's specification (R&S FSW-B25: 40 dBm for  $f > 500$  MHz).



**Figure 10: Additional elements allow the control of the signal level applied to the 1<sup>st</sup> mixer**

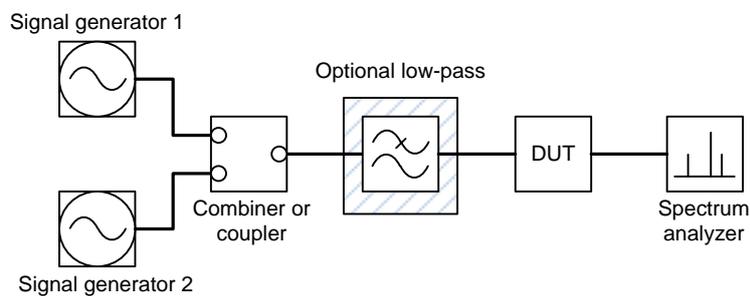
Many spectrum analyzers also offer optional pre-amplifiers to obtain more sensitivity. As the characteristic of an amplifier is usually not completely linear, an amplifier is always a good source for intermodulation products. Not only are intermodulation products generated in the amplifier itself, but it also increases the signal level at the input mixer, which in turn may cause more intermodulation from the mixer.

In general, it is therefore recommended not to switch on an existing pre-amplifier when measuring IMD. Only in scenarios with very low input levels the use of a pre-amplifier may make sense. The R&S FSW's optional preamplifier (R&S FSW-B24) is specified with a third order intercept point of -10 dBm ( $f < 1$  GHz), with two -50 dBm input tones applied.

## 4 Measurement of Intermodulation Distortion

This chapter highlights the critical points when measuring IMD. It starts with the generation of the two tone scenario and continues with the influence of the different stages of a spectrum analyzer. This chapter also gives recommendations on how to optimize the measurement setup for minimum intermodulation, in order to get a good measurement of the IMD of the DUT. It covers the signal generation side as well as the spectrum analyzer side.

### 4.1 Measurement setup



**Figure 11: Typical measurement setup for a two tone intermodulation measurement.**

Typically, a measurement setup like in Figure 11 is used for intermodulation measurements. Two signal generators, one for each tone, create the test signal. The two signals are combined by a power combiner or coupler, before being fed into the DUT input port. The DUT output port is connected to the spectrum analyzer.

As there are quite some other non-linear elements besides the DUT in this schematic, it is necessary to optimize the measurement setup in a way that the measured results describe the DUT and not the spectrum analyzer or the signal generators.

### 4.2 Generator influence

Even though each signal generator in the above setup creates a single tone only, precautions need to be taken in order to avoid influence from the generators. Modern signal generators are equipped with automatic level control (ALC) functionality. This feature is a feedback loop, which controls the generator output power. Since the power detection of the ALC is not frequency selective, it will also “see” signals that originate from the second signal generator. As the resulting sum power of two non-coherent CW signals depends on their phase relation, the ALC power reading will fluctuate, as long as the signal generators are not phase locked to each other.

There are two ways to avoid level fluctuations during the measurement due to ALC influence:

1. Isolate the signal generators from each other: Use a coupling device that has a high isolation between the input ports. Couplers with 20 or more dB are better suited than power combiners (resistive design) with only 6 dB of isolation.

2. Turn off the ALC: Most signal generators allow disabling the ALC. This results in an output level that may not correspond exactly to what the generator displays, but it avoids changes of the output level due to phase changes between the two tones. Since the spectrum analyzer measures the tone power, it is not necessary to know the tone power precisely in advance.

Generators may also generate harmonics. The second harmonic (signal generator caused) of one tone will mix with the fundamental of the other tone to the same frequency as the TOI product of both fundamentals. To avoid the influence of generator created harmonics:

3. Use a low-pass filter: Depending on the harmonic suppression of the signal generators in use, it may be necessary to utilize external harmonic suppression filters to minimize the generator influence on the intermodulation measurement. For high dynamic range measurements (> 90 dB), it is generally recommended to use low-pass filters to minimize the influence of harmonics created by the signal generators (see Figure 11).

Although vector signal generator are capable of generating signals with two or more carriers, it is not recommended to use a single source for both tones. Internal intermodulation in the generator and limited dynamic range of the so called arbitrary waveform generators limit the maximum achievable dynamic range.

## 4.3 Spectrum analyzer influence

This section focuses on the spectrum analyzer influence on the measurement. Other than chapter 3, this section does not focus on the design background, but shows the impact of each element on the measurement and how it can be avoided.

### 4.3.1 RF preamplifier and electronic attenuator

Both RF preamplifiers and electronic attenuators are not ideally linear elements which, when used, will contribute to the IMD measurement results. The recommendation is therefore to switch the RF preamplifier off. If this is not possible, e.g. because the DUT output power is very low and requires an RF preamplifier to lower the noise floor, the datasheet specifications covering the residual IMD of these components should be carefully taken into account (R&S FSW-B24: -10 dBm with two -50 dBm tones,  $f < 1\text{GHz}$ ).

Electronic attenuators are however not as critical, since their TOI specification often exceeds the specification of the spectrum analyzer base unit (R&S FSW-B25: 40 dBm with two -15 dBm tones,  $500\text{ MHz} < f < 13.6\text{ GHz}$ ).

### 4.3.2 Mixer stage influence

The input mixer is usually the first non-linear element in the signal path of the spectrum analyzer (see Figure 3 and Figure 5), assuming that an available RF preamplifier is switched off and that an optional electronic attenuator is bypassed.

As shown in Figure 4 and Figure 7, the intermodulation distortion created by the first mixer stage dominates for high mixer levels. The counter measure against mixer based intermodulation distortion is a reduction of the signal level at the mixer input, ideally without creating additional intermodulation. As spectrum analyzers are equipped with mechanical attenuators (see Figure 10) which contain only passive components and therefore do not contribute to IMD, increasing the mechanical attenuation will improve the intermodulation behavior of the mixer stage.

Since third order intermodulation products created by a mixer decrease by 30 dB with every 10 dB decrease of the signal level, an additional 10 dB in mechanical attenuation results in roughly 20 dB more dynamic range. This holds true for the case that the measured intermodulation products originate from the mixer only.

### 4.3.3 Noise floor influence

The noise floor may be a limiting factor during intermodulation measurements, since it may limit the dynamic range at low signal levels, given that intermodulation products are much smaller than the displayed noise floor. Figure 4 and Figure 7 show in red the limitation of the dynamic range by the noise floor. As for all spectrum analyzer measurements, the noise floor can be lowered by three methods:

1. Reduce the resolution bandwidth: Measurements on CW signals are not dependent on a specific minimum RBW. Therefore, the RBW can simply be reduced to benefit from the decrease in noise floor.
2. Use noise cancellation: Modern spectrum analyzers provide a so-called noise cancellation function. Its name may be different for different spectrum analyzer vendors, but its mode of operation is always a subtraction of the known inherent noise power of the spectrum analyzer from the measured power. Noise cancellation may improve the S/N ratio by up to 10 dB. Although this method works in principle for all trace detectors, it provides the highest benefit when using the RMS detector. When using the RMS detector, the span/RBW ratio should be kept below a certain ratio (e.g. one third of the number of sweep points or less), to avoid averaging the tone with neighboring noise.
3. Use an internal pre-amplifier: If available, an internal pre-amplifier may be used. As a preamplifier is a non-linear element located in front of the input mixer, the preamplifier creates additional intermodulation signals (see section 3.4). So this method must be used with care, but may be useful for very low power levels.

### 4.3.4 ADC influence

ADCs are non-linear components whose intermodulation distortion follows rules which are different from other common RF components. Their intermodulation distortion is not specified using the TOI, but is included in the spurious free dynamic range (SFDR) specification. SFDR covers not only the intermodulation products, but all unwanted signals. Since SFDR specifications are given in "dB below full scale" (dBFS), it is essential to scale the signal correctly before applying it to the ADC. On a spectrum analyzer, the so-called IF gain, which is often coupled to the reference level, is used to optimize the signal level in front of the ADC.

The intermodulation products of an ADC remain more or less constant, independent of its input signal level. This in return means that the SFDR is dominated by the signal level, which means that a higher ADC input level will result in the same increase in SFDR.

Recommendations:

1. The best method to get around the ADC related IMD contribution is to avoid having two tones at the ADC input simultaneously by selecting an appropriate tone spacing (e.g. > 5 MHz for the R&S FSW and R&S FSV).
2. If the tone spacing is fixed and cannot be changed, the CW tones should be close to the full scale level of the ADC. The R&S FSW in default setting automatically takes care about the signal scaling for the ADC, using as much of the ADC's scale as possible and avoiding an ADC overload at the same time.

### 4.3.5 Finding the ideal mixer level

As a conclusion, this section will focus on the probably most important point: how should the spectrum analyzer be configured in order to get the highest possible intermodulation free dynamic range.

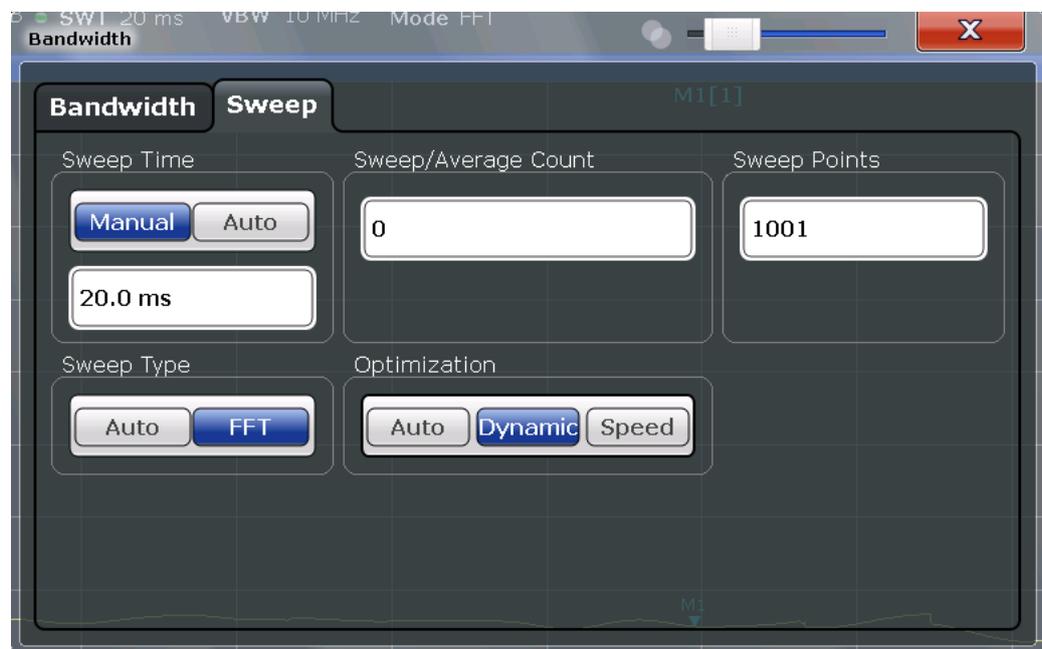
For a spectrum analyzer with analog RBW filters which follows the classical "V" shape with its intermodulation free dynamic range versus input level as shown in Figure 4, it is straightforward to determine the ideal mixer level for IMD measurements. It is the intersection of the noise floor line with the mixer TOI line, the so-called "sweet spot". Its characteristic is that the "sweet spot" is the only minimum on the inherent intermodulation curve and is therefore easy to find.

For signal- and spectrum analyzers with wide-band signal paths, the theoretical sweet spot can be used as a starting point for the measurement setup, but under certain conditions the ADC behavior must be considered as it may dominate the inherent intermodulation distortion. Its influence can be minimized by following these recommendations:

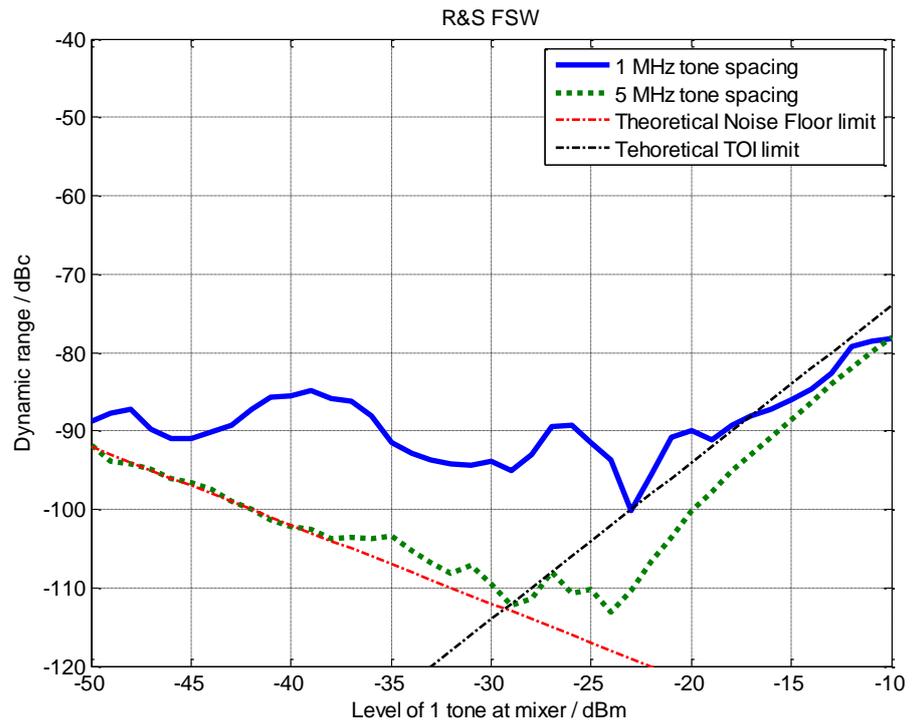
1. Increase the tone spacing: Once the tone spacing of the test signal is wider than the analog IF bandwidth in front of the ADC, the behavior of a wide-band analyzer is identical to a spectrum analyzer with analog RBW filters (see Figure 8).
2. Use the smallest available analog IF filter bandwidth: The R&S FSW for example offers a "dynamic" setting of the analog IF pre-filter in front of the ADC for sweep optimization (see Figure 12). This setting selects the analog IF filter with the smallest bandwidth for the current sweep settings. On the R&S FSW, this is a 5 MHz filter for RBW settings up to 300 kHz.

- Choose the optimum mixer level: If the tone spacing is narrow and cannot be varied due to a given specification, fine adjust the signal level at the input mixer to make use of minima in the IMD curve of the spectrum analyzer. As shown in Figure 9 and Figure 13, the dynamic range may have local minima. A dynamic range of roughly 90 dB can be obtained with the R&S FSW between -50 dBm and -20 dBm, with an optimum value of 100 dB and a worst case value of 85 dB. To obtain the optimum mixer level, it is necessary to vary the RF attenuation by at least  $\pm 5$  dB. The optimum mixer level is reached as soon as the intermodulation products do not decrease any more. As mentioned before, modern signal- and spectrum analyzers, such as the R&S FSW will take care of the ADC level automatically, if they are in either "Auto" or "Dynamic" mode (see Figure 12).

Figure 13 shows the intermodulation measurement on an R&S FSW with a tone spacing wider than the analog IF filter bandwidth (green) compared to a tone spacing smaller than the analog IF filter bandwidth (blue). Clearly, for the wide tone spacing, it is easy to obtain more than 100 dBc intermodulation free dynamic range, using the above recommendations.



**Figure 12: Sweep optimization for maximum dynamic range on the R&S FSW (Sweep Config dialog). This setting selects the smallest possible analog filter.**



**Figure 13: Measured inherent IMD of the R&S FSW, plotted against the theoretical limits,  $f=1600$  MHz, 10 Hz RBW, no noise correction**

## 5 Ordering Information

R&S FSW8	Signal- and Spectrum analyzer 2 Hz to 8 GHz	1312.8000.08
R&S FSW13	Signal- and Spectrum analyzer 2 Hz to 13.6 GHz	1312.8000.13
R&S FSW26	Signal- and Spectrum analyzer 2 Hz to 26.5 GHz	1312.8000.26
R&S FSV3	Signal analyzer 10 Hz to 3.6 GHz	1307.9002.03
R&S FSV7	Signal analyzer 10 Hz to 7 GHz	1307.9002.07
R&S FSV13	Signal analyzer 10 Hz to 13.6 GHz	1307.9002.13
R&S FSV30	Signal analyzer 10 Hz to 30 GHz	1307.9002.30
R&S FSV40	Signal analyzer 10 Hz to 40 GHz	1307.9002.40

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