

# Measuring with Modern Spectrum Analyzers

## Educational Note

This educational note examines both the theory and practice of state-of-the-art spectrum analysis by discussing specific measurements taken on an integrated RF amplifier.

The theoretical portion of this document first describes how modern spectrum analyzers are designed and how they work. That is followed by a brief characterization of today's signal generators, which are needed as a stimulus when performing amplifier measurements. The effects on the spectrum that are caused by the non-linearity of real devices under test are derived mathematically.

The practical portion of this educational note contains detailed test and measurement tasks that are suitable for use as lab exercises performed in small groups. The purpose of these tasks is to illustrate the explanations provided in the theoretical section and to provide a deeper understanding of the principles explained there.

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# 1 Overview

This educational note examines both the theory and practice of state-of-the-art spectrum analysis by discussing specific measurements taken on an integrated RF amplifier.

In chapter 2, the theoretical portion of this paper describes how modern spectrum analyzers are designed and how they work. That discussion is followed in chapter 3 by a brief characterization of today's signal generators, which are required as a stimulus when performing amplifier measurements. The theoretical portion of this document comes to its conclusion in chapter 4 by showing how the effects on the spectrum that are caused by the nonlinearity of real devices under test are derived mathematically.

In chapters 9 through 12, the practical portion of this educational note provides detailed test and measurement tasks that are suitable for use as lab exercises performed in small groups. The purpose of these tasks is to illustrate the explanations provided in the theoretical section and to provide a deeper understanding of the principles explained there.

The following topics will be covered:

- Measuring an amplifier's 3 dB frequency response
- Determining the 1 dB compression point
- Harmonic interference, IP2 / IP3 measurements
- Intermodulation measurements
- Measuring the crest factor and the ACLR
- Phase-noise measurements

An RF generator and a spectrum analyzer are used for these experiments. Most labs are equipped with these stand-alone devices.

If the analyzer has a built-in tracking generator, many of the experiments described here can be performed with that one device alone. In the same way, a network analyzer can handle many of these tasks alone.

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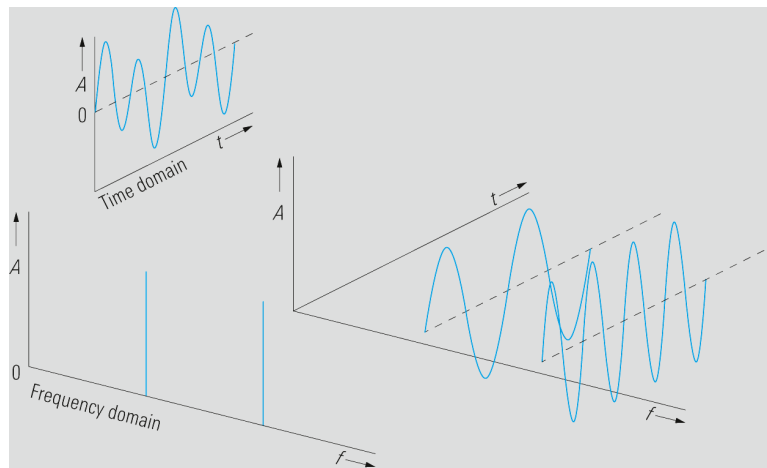
[TM-Applications@rohde-schwarz.com](mailto:TM-Applications@rohde-schwarz.com)

## 2 The Basics of Spectral Analysis

For the most part, the structure and content of this chapter have been taken from the book "Fundamentals of Spectrum Analysis" by Christoph Rauscher [1]. That work contains further references and links for a more in-depth look at this subject.

### 2.1 Correlation between the Time and Frequency Domains

Electrical signals can be represented and observed in both the time domain and frequency domain.



**Fig. 1: Representation of signals in the time and frequency domains.**

These two methods of representation are linked together via the Fourier transform, which means that there is a characteristic frequency spectrum for every signal that can be represented in the time domain. The following applies:

$$\underline{X}_f(f) = F\{x(t)\} = \int_{-\infty}^{\infty} x(t) \cdot e^{-j2\pi ft} \cdot dt \quad (1)$$

Or:

$$x(t) = F^{-1}\{\underline{X}_f(f)\} = \int_{-\infty}^{\infty} \underline{X}_f(f) \cdot e^{j2\pi ft} \cdot df \quad (2)$$

Where

$\underline{X}_f(f)$	Complex signal in the frequency domain
$x(t)$	Signal in the time domain
$F\{x(t)\}$	Fourier transform of $x(t)$
$F^{-1}\{\underline{X}_f(f)\}$	Inverse Fourier transform of $\underline{X}_f(f)$

### Periodic signals:

Any periodic signal in the time domain can be represented by the sum of sine and cosine signals of different frequencies and amplitudes. The resulting sum is called a Fourier series.

$$x(t) = \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cdot \sin(n \cdot 2\pi f_0 \cdot t) + \sum_{n=1}^{\infty} B_n \cdot \cos(n \cdot 2\pi f_0 \cdot t) \quad (3)$$

Where

$$A_0 = \frac{2}{T_0} \int_0^{T_0} x(t) \cdot dt$$

$$A_n = \frac{2}{T_0} \int_0^{T_0} x(t) \cdot \sin(n \cdot 2\pi f_0 \cdot t) \cdot dt$$

$$B_n = \frac{2}{T_0} \int_0^{T_0} x(t) \cdot \cos(n \cdot 2\pi f_0 \cdot t) \cdot dt$$

Since the frequency spectra of the sine and cosine signals having the frequency  $f_0$  can be represented by Dirac delta functions at the frequencies  $-f_0$  and  $f_0$ , the signal spectrum of a periodic signal can only consist of discrete spectral lines of defined amplitudes.

$$F\{\sin(2\pi f_0 \cdot t)\} = \frac{1}{2j} (\delta(f - f_0) - \delta(f + f_0)) \quad (4)$$

$$F\{\cos(2\pi f_0 \cdot t)\} = \frac{1}{2j} (\delta(f - f_0) + \delta(f + f_0)) \quad (5)$$

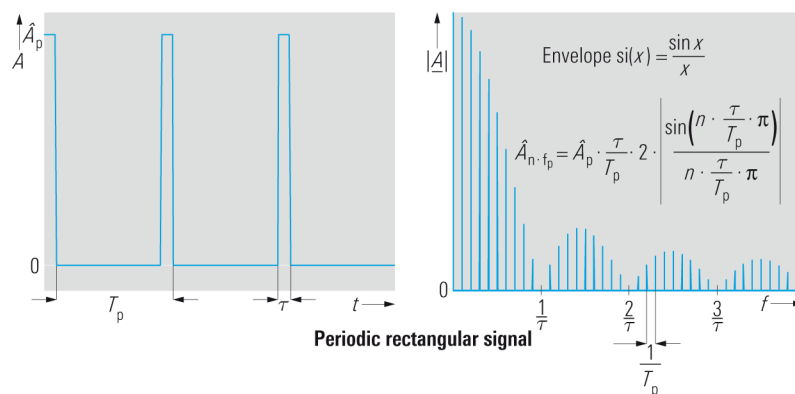
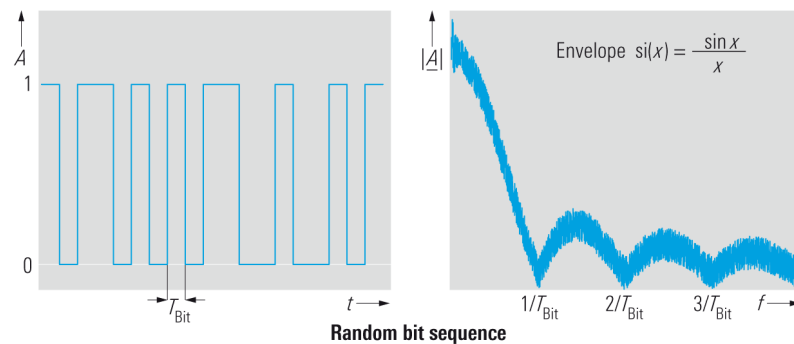


Fig. 2: Periodic rectangular signal in the time and frequency domains.

### Nonperiodic signals:

Signals with a nonperiodic time characteristic cannot be represented by a Fourier series. Time signals of this kind have no discrete spectral components, but rather a continuous frequency spectrum. Here, as for sinusoidal signals, a closed-form solution can be found for many signals (using Fourier-transform tables).

Nevertheless, there is seldom a closed-form solution for signals with a random time characteristic, such as noise or random bit sequences. In such cases, it is easier to determine the spectrum using a numeric solution for Eq. (1).



**Fig. 3: A nonperiodic sequence of random bits in the frequency and time domains.**

Signals can be described in both the time domain and in the frequency domain. Consequently, a signal can also be acquired in both the time domain and the frequency domain; provided the general conditions remain within certain limits, it is possible to convert the respective results back and forth.

For this reason, the next section will first cover fast Fourier transform (FFT) analyzers, which capture signals in the time domain. Then the discussion will move on to traditional spectrum analyzers, which tune the frequency range directly. Finally, we will use a state-of-the-art signal and spectrum analyzer as an example to show that it is possible to combine the advantages of both approaches.



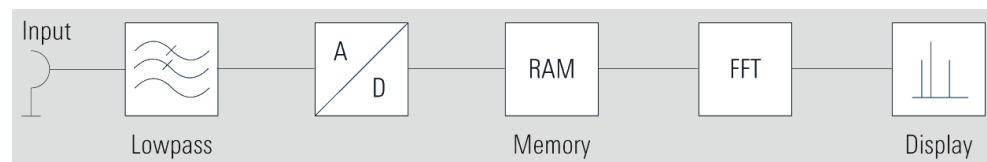
## 2.2 FFT Analyzers

A fast Fourier transform (FFT) analyzer calculates the frequency spectrum from a signal that was captured in the time domain. Nevertheless, performing an exact calculation would require an observation period of infinite length. Furthermore, achieving an exact result for Eq. (1) would require knowledge of the signal amplitude at every point in time. The result of that calculation would be a continuous spectrum, which means that the frequency resolution would be unlimited.

Obviously, it is not possible to perform such calculations in practice. Nonetheless, under certain circumstances, it is possible to determine the signal spectrum with sufficient accuracy.

### 2.2.1 Architecture

Fig. 4 shows a block diagram outlining the primary elements that make up an FFT analyzer.



**Fig. 4: Block diagram of an FFT analyzer.**

In practice, the Fourier transform is performed with the aid of digital signal processing (discrete Fourier transform), which means that the signal to be analyzed must first be sampled by an A/D converter, and its amplitude has to be quantized. In order to enforce conformity with the sampling theorem, an analog lowpass filter ( $f_g = f_{e,\max}$ ) is employed to limit the input signal's bandwidth before the signal arrives at the A/D converter. Once the time-domain signal has been digitized, the discrete time values of a specific amplitude are stored temporarily in RAM, and those values are used to calculate the spectral components by applying the fast Fourier transform. Then the spectrum is displayed.

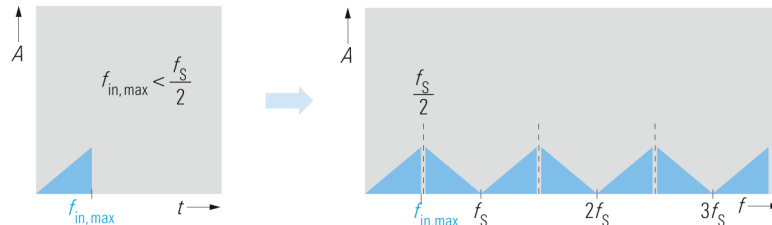
### 2.2.2 The Basics of How an FFT Analyzer Works

#### Sampling system

In order to ensure that aliasing effects do not cause ambiguity during signal sampling, it is necessary to limit the bandwidth of the input time-signal. According to Shannon's sampling theorem, the sampling frequency  $f_s = \frac{1}{T_s}$  of a lowpass-filtered signal must be at least twice as high as the maximum signal frequency  $f_{in,\max}$ . The following applies:

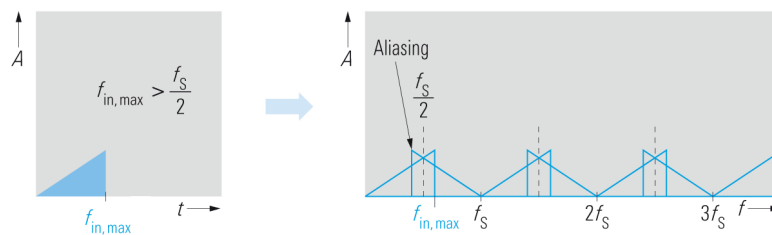
$$f_s \geq 2 \cdot f_{in,max} \tag{6}$$

Since the edge slope of the lowpass filter that is used to limit the bandwidth is not infinite, the sampling frequencies used in practice are significantly higher than  $2 \cdot f_{e,max}$ .



**Fig. 5: Sampling a lowpass signal without aliasing:**

$$f_{in,max} < \frac{f_s}{2}$$



**Fig. 6: Sampling a lowpass signal with aliasing:**

$$f_{in,max} > \frac{f_s}{2}$$

### Windowing

Only a portion of the signal is considered for the Fourier transform. Consequently, only a limited number  $N$  of samples is used to calculate the spectrum. During this windowing process, the input signal that has been discretized after sampling is then multiplied with a certain window function.

Sampling turns Eq. (1) into the discrete Fourier transform (DFT):

$$\underline{X}(k) = \sum_{n=0}^{N-1} x(n \cdot T_s) \cdot e^{-j2\pi kn / N} \tag{7}$$

Where

- $\underline{X}(k)$  Complex discrete frequency spectrum
- $x(n \cdot T_s)$  Sample at the time  $n \cdot T_s$
- $k$  Index of the discrete frequency bins,  $k = 0, 1, 2, \dots$
- $n$  Index of the samples,  $n = 0, 1, 2, \dots$
- $N$  Length of the DFT

This results in a discrete frequency spectrum that has individual components in what are known as frequency bins  $f(k) = k \cdot \frac{f_s}{N} = k \cdot \frac{1}{N \cdot T_s}$ . Here it is possible to recognize that the spectral resolution – i.e. the minimum spacing that two of the input signal's spectral components must exhibit in order to display two different frequency bins  $f(k)$  and  $f(k + 1)$  – depends on the observation period  $N \cdot T_s$ .

The following prerequisites must be met in order to enable precise calculation of the discrete signal spectrum:

The signal must be periodic (length of the period  $T_0$ ).

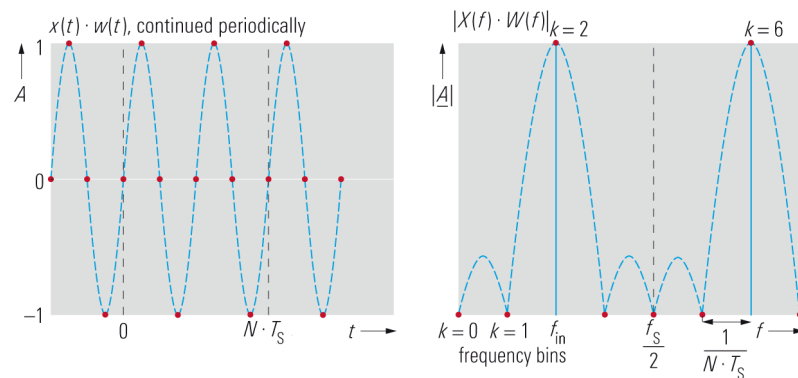
The observation period  $N \cdot T_s$  must be an integer multiple of  $T_0$ .

Here is the reason why:

In the frequency domain, multiplying the time signal with a square window corresponds to convolution of the spectrum with a si function.

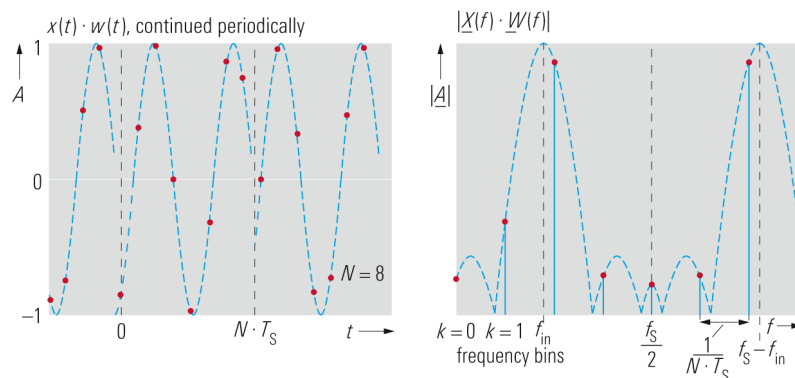
$$|W(f)| = N \cdot T_s \cdot \text{si}(2\pi f \cdot N \cdot T_s / 2) = \frac{\sin(2\pi f \cdot N \cdot T_s / 2)}{2\pi f \cdot N \cdot T_s / 2} \tag{8}$$

If the input signal is periodic and the observation period is an integer multiple of the period's length, no error arises, because – with the exception of the signal frequencies – the samples coincide with the si function's zero-crossings.



**Fig. 7: Error-free DFT for a periodic input signal at integer multiples of the period's length.**

If the above mentioned conditions are not met, convolution with the window function according to Eq. (8) will smear the resulting signal spectrum, thus widening it significantly. This effect is referred to as leakage. Simultaneously, amplitude errors arise.



**Fig. 8: DFT for a periodic input signal when the multiples of the period length are not integers (leakage).**

While extending the observation period can reduce leakage by boosting the resolution, this does not decrease the amplitude error. It is possible to reduce both effects, however, by employing an optimized window function instead of the rectangle window. Such window functions (such as a Hann window, for instance) exhibit lower secondary maximums in the frequency domain, which reduces leakage. Window functions can reduce the amplitude errors with a flat main lobe (flattop window: max. level error of 0.05 dB). Nevertheless, this results in the disadvantage of a relatively wide main lobe, which leads to a lower frequency resolution.

When a nonrectangular window is used, a systematic error always arises, even when the observation period is an integer multiple of the signal period (because the zero-crossings are no longer the same).

### 2.2.3 The Difference between FFT Analyzers and Oscilloscopes

Fig. 4 shows the similarity between FFT analyzers and oscilloscopes, which also sample the signal in the time domain and offer the option of spectral display. During the design process, however, different criteria are applied when selecting the A/D converter (ADC). The distinguishing feature of a spectrum analyzer is its high dynamic range. Engineers developing oscilloscopes, on the other hand, tend to select converters with a high sampling rate in order to be able to properly represent the steep edges of square waves and pulsed signals in the time domain. Nevertheless, the ADC's quantization depth depends on its maximum sampling rate. A/D converters follow this general principle: The higher the sampling rate, the lower the available quantization depth. The quantization depth is determined by the number of bits used to represent a sample. The quantization noise and the maximum dynamic range can be derived from the quantization depth. Therefore, to enable spectrum analyzers to achieve the ideal dynamic range, developers select ADCs with a greater quantization depth and a correspondingly lower sampling rate.

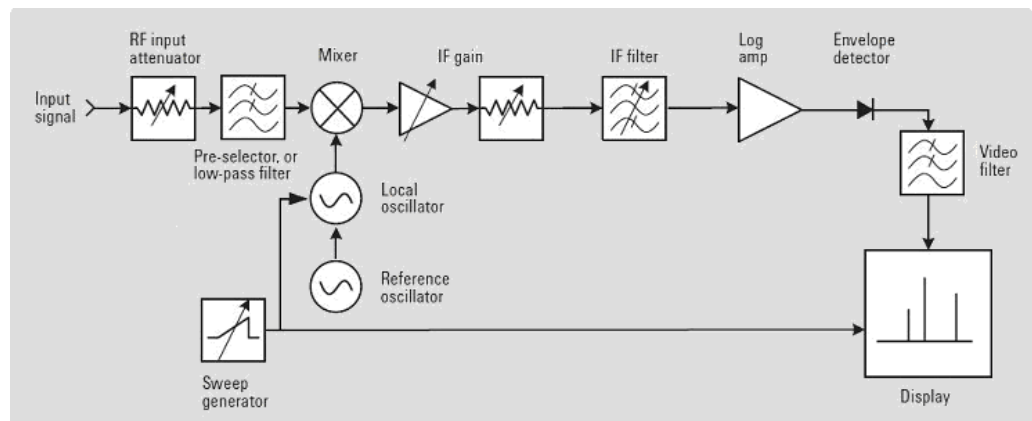
## 2.3 Analyzers that Use the Heterodyne Principle

In order to represent the spectra of radio-frequency signals all the way up into the microwave or millimeter-wave band, analyzers with frequency converters (heterodyne principle) are used. Here, the input signal's spectrum is not calculated from the time characteristic; instead, it is calculated by performing analysis directly in the frequency domain. The input spectrum can be broken down into its individual components using a bandpass filter that has been selected to match the analysis frequency, whereby the filter bandwidth represents the resolution bandwidth (RBW). From an engineering perspective, realizing such narrowband filters that can be tuned across the entire input frequency range is a difficult task. In addition, filters have a constant relative bandwidth with reference to the center frequency, which causes the absolute bandwidth to increase as the center frequency rises. For this reason, this concept is not suitable for spectrum analyzers.

As a rule, analyzers for higher input frequency ranges operate in the same way as a heterodyne receiver. Section 2.3.1 shows that the RBW remains constant in such cases.

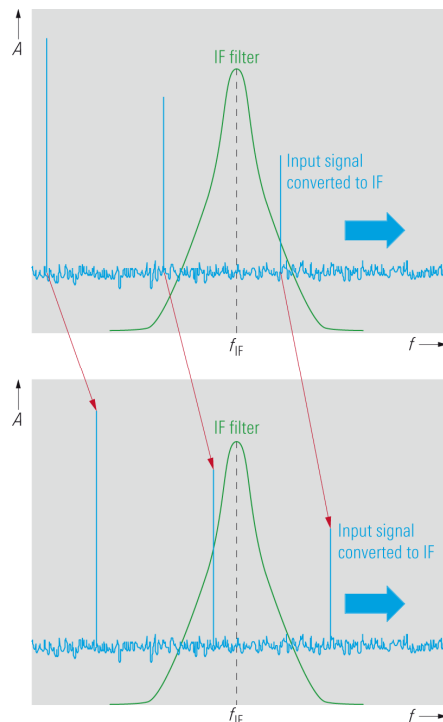
### 2.3.1 Architecture

Fig. 9 shows a block diagram outlining the design of a spectrum analyzer that employs the heterodyne principle.



**Fig. 9: Block diagram of a traditional spectrum analyzer that uses the heterodyne principle [2].**

In a heterodyne receiver, a mixer and a local oscillator (LO) are used to convert the (lowpass-filtered) input signal to an intermediate frequency (IF). The LO in Fig. 9 is tuned by a sweep generator in order to convert the entire input frequency range to a constant intermediate frequency. The IF signal is amplified and arrives at the IF filter with a definable bandwidth. The input signal is essentially "swept past" this filter with a fixed center frequency (Fig. 10).



**Fig. 10:** Inside the heterodyne receiver, the signal is "swept past" the resolution filter.

The IF filter in Fig. 9 is what determines the analyzer's resolution bandwidth (RBW).

In order to be able to display signals with a wide variety of levels on the screen simultaneously, the IF signal is compressed with the aid of a logarithmic amplifier. After that, the envelope detector and the video filter acquire the signal's envelope, and the noise is reduced with the aid of an averaging process, which smooths out the displayed video signal.

With earlier technology, the video signal was fed through a vertical cathode ray tube's vertical deflection. To display the frequency dependency, the tube's horizontal deflection was accomplished with the aid of the same saw-tooth sweep signal that was used to tune the LO. Since the intermediate frequency and the LO frequency are known, the relationship between the input signal and the display on the frequency axis is unambiguous.

Today's advanced analyzers use high-speed digital processing. This means that an ADC samples the IF signal and that the signal is then processed digitally.

With modern analyzers, the LO is locked to a reference frequency with the aid of a phase-locked loop (PLL) and coordinated in discrete steps by varying the scaling factors.

In addition, the video signal is prepared digitally, and a liquid-crystal display replaces the cathode ray tube.

### 2.3.2 Frequency Selection

The following mathematical context applies when converting the input signal in the mixer to the intermediate frequency (IF) with the aid of an LO signal:

$$|m \cdot f_{LO} \pm n \cdot f_{in}| = f_{IF} \quad (9)$$

Where

$m, n$	Natural numbers
$f_{in}$	Frequency of the input signal to be converted
$f_{LO}$	Frequency of the local oscillator
$f_{IF}$	Intermediate frequency

The following applies for the signal's fundamental (1st harmonic):

$$|f_{LO} \pm f_{in}| = f_{IF} \quad (10)$$

Or, when solving for  $f_{in}$

$$f_{in} = |f_{LO} \pm f_{IF}| \quad (11)$$

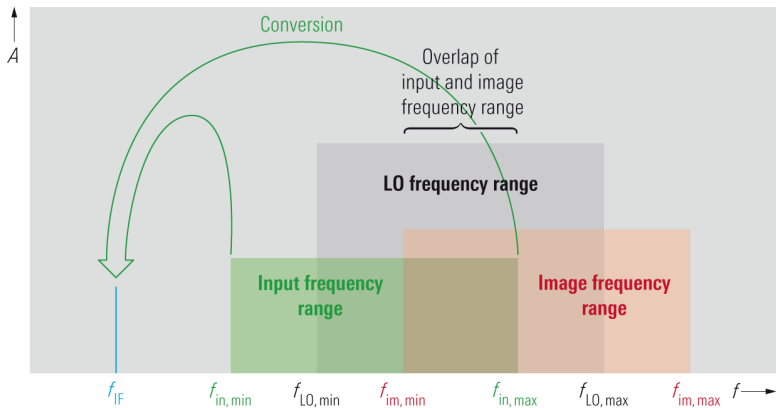
Taking a close look at Eq. (11) reveals that, for the oscillator frequencies and intermediate frequencies, two receiving frequencies always exist for which the criterion from Eq. (10) is fulfilled. This means that, besides the desired input frequency range, there is also another one, the "image frequency band."

In order to ensure unambiguous results, possible input signals for the image frequencies must be suppressed by employing corresponding filters ahead of the mixer's RF input.

#### Low IF:

A spectrum analyzer should be able to process the widest possible input frequency range; here, having a low IF leads to limitations:

If the input frequency range is greater than  $2 \cdot f_{IF}$ , the input frequency and image frequency ranges overlap, see Fig. 11.

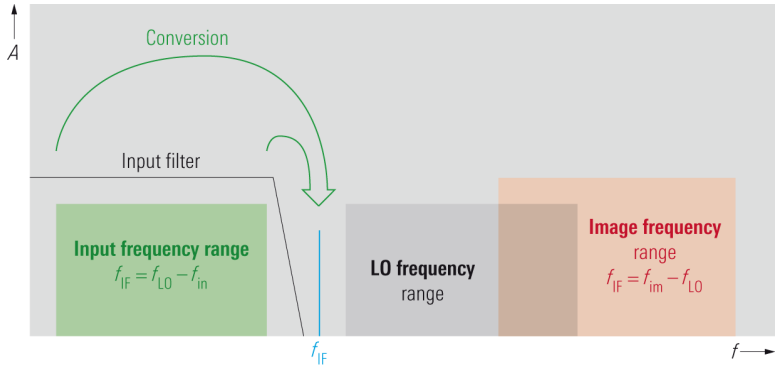


**Fig. 11: Low IF, large input-frequency range: The input and image frequency ranges overlap.**

With the lower LO frequencies, a signal is received from both the green input frequency range and from the red image frequency range. This means that, in order to achieve image frequency rejection without harming the desired input signal, the input filter must be implemented as a tunable bandpass filter. Doing that is highly complex from a technical standpoint.

**Principle of using a high first intermediate frequency:**

When a high first intermediate frequency is used, the IF lies above the input frequency range. Consequently, the image frequency range is then located above the input frequency range. Since the two ranges do not overlap, image rejection can be accomplished with a lowpass filter that has fixed tuning, see Fig. 12.



**Fig. 12: The high-IF principle.**

The following applies for conversion of the input signal:

$$f_{IF} = f_{LO} - f_{in} \tag{12}$$

or for the image reception areas:

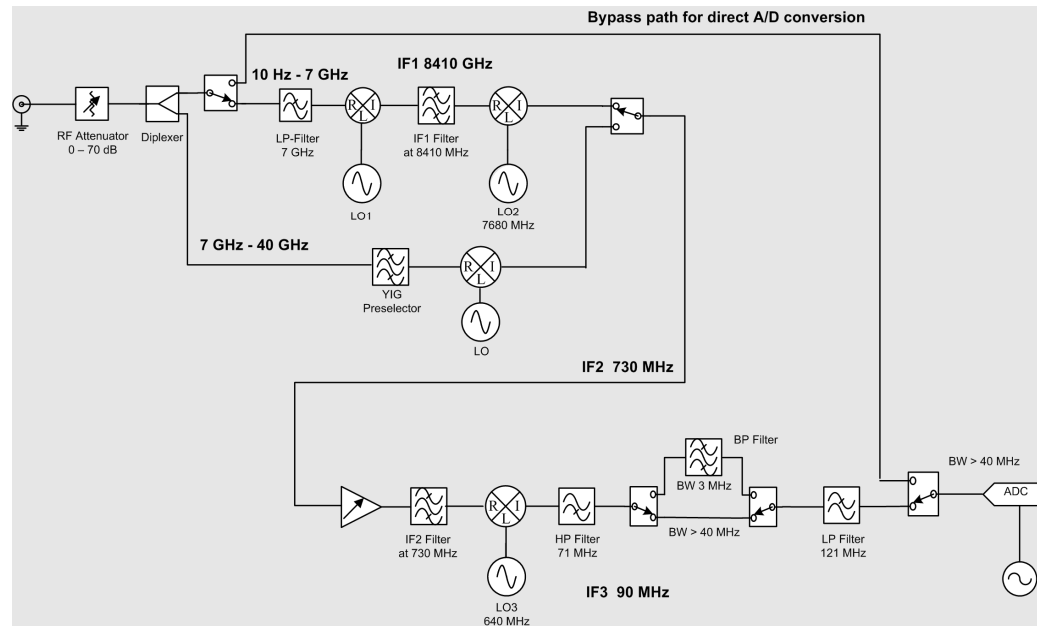
$$f_{IF} = f_{in} - f_{LO} \tag{13}$$



**Example:**

The following description applies for the R&S®FSV40 signal analyzer used in the lab exercises. To arrive at the shared intermediate frequency IF2, this analyzer uses different techniques for frequencies up to 7 GHz and for those higher than 7 GHz. In addition, a bypass for low frequencies allows direct sampling of the input signal. (These ranges and frequencies are similar with other analyzer models.)

Fig. 13 shows a block diagram of the analog stages.



**Fig. 13: Block diagram of the analog portion in the R&S®FSV40 signal analyzer.**

**Path for frequencies from 10 Hz to 7 GHz:**

The high-IF concept is employed here. The first intermediate frequency (IF1) is 8.41 GHz; thus, it lies above the highest receiver frequency. In order to be able to convert the entire input frequency range from 10 Hz to 7 GHz to 8.41 GHz, the LO1 signal must be tunable in the frequency range from 8.41 GHz to 15.41 GHz. The image range then lies in the frequency range from 16.82 GHz to 23.82 GHz. In this way, the low-pass filter (up to 7 GHz) that is employed ahead of the mixer has no trouble filtering out the input frequency range and sufficiently suppressing the image frequency. In order to add narrowband filtering on the 8.41 GHz signal and perform further processing on it, it must be reduced to a lower IF (in this case, approx. 90 MHz). Since the first IF is high, a very complex filter with a steep edge slope would be required for direct conversion to 90 MHz in order to suppress the nearby image frequencies. For this reason, the IF1 signal is first reduced to the middle intermediate frequency, IF2 (730 MHz); then it is amplified and filtered before being mixed down to IF3 at about 90 MHz.

### Path for Frequencies above 7 GHz

The principle of using a high first IF becomes increasingly difficult to implement as the upper input frequency rises. For this reason, this principle is not used here for input signals above 7 GHz; instead those signals are converted directly to a low intermediate frequency. Doing this requires a tracking bandpass filter for image frequency rejection. Converting this frequency range to a lower IF is possible, because

the frequency range from 7 GHz to 40 GHz covers less than a decade (10 Hz to 7 GHz, on the other hand, corresponds to 8.8 decades) and YIG technology makes it possible to build a narrowband bandpass filter that is tunable across this frequency range.

The frequencies above 7 GHz cannot be mixed down to the desired low intermediate frequency, IF3 (approx. 90 MHz) in a single step either. For this reason, these frequencies, too, are first converted to 730 MHz. After that, the signal is amplified and coupled into the IF signal path for the low-frequency input stage.

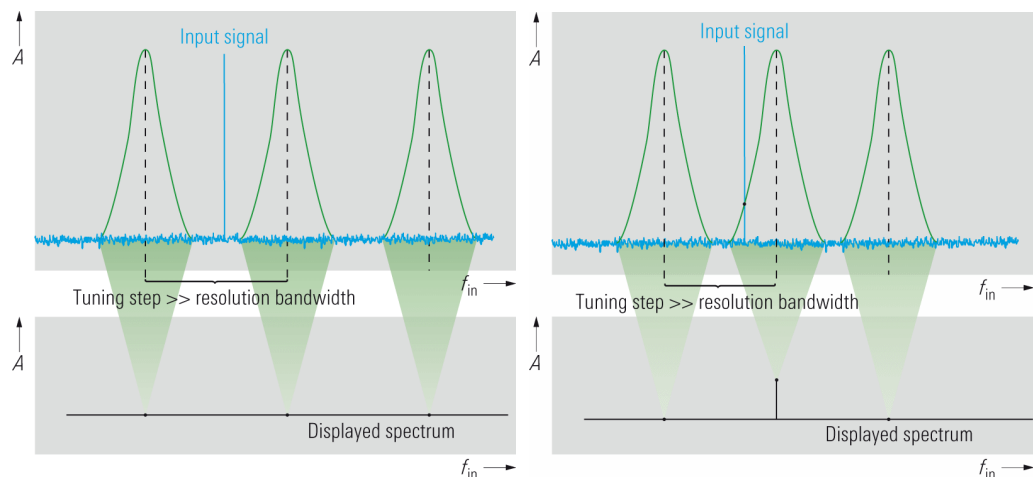
Further processing of the IF2 signal will be taken up again in section 2.3.4.

### 2.3.3 Step-by-Step Tuning of the Local Oscillator (LO)

Due to the broad tuning range and low phase-noise that this technology offers, a YIG oscillator is usually employed as the local oscillator. Some spectrum analyzers also use voltage controlled oscillators (VCO) for the LO.

In both cases, modern analyzers use a phase-locked loop to tie the oscillator to a reference signal. That is the only way to achieve good frequency accuracy and stability. Nevertheless, only discrete frequency steps are possible for this task. Consequently, such analyzers can only be tuned in discrete steps.

The step width required for this depends on the setting for the resolution bandwidth. A narrow resolution bandwidth requires small tuning steps, because larger steps would cause information from the input spectrum to become lost, or level errors could arise, see Fig. 14 on page 19. In order to prevent such errors, the analyzer automatically selects a step width that is significantly smaller than the resolution bandwidth – for example,  $0.1 \cdot \text{RBW}$  (see section 2.4.6 for more information on the resolution bandwidth).



**Fig. 14: Effects of using a tuning stepwidth that is too large: Left: Input signal is completely lost; Right: Level errors arise in the representation of the input signal**

### 2.3.4 IF Signal Processing

The amplified and filtered IF2 intermediate frequency signal with 730 MHz is converted to the low IF, IF3 (approx. 90 MHz), using the LO3 mixer.

In the low intermediate frequency, IF3, the signal is amplified again in certain stages and limited to the selected bandwidth by filters. Here, the amplification can be set in steps, which keeps the maximum signal level constant in a way that is independent of the attenuator setting, and thus independent of the mixer level. The settings for the IF gain can be chosen to allow the best possible exploitation of the ADC's dynamic range by setting the ADC's maximum input level so that it corresponds to the level of the largest signal within the IF bandwidth.

Depending on the concept used for the specific spectrum analyzer, with some analyzers, users can influence this gain. This is usually done by selecting the reference level. High reference levels result in a low IF gain; low reference levels result in high gain. With some spectrum analyzers, however, the reference level setting is decoupled from the IF gain, which means that the IF gain remains constant; in such cases, changing the reference level only influences the representation of the signal on the display through numeric scaling inside the computer.

With a traditional spectrum analyzer, the IF amplification is followed by what is referred to as resolution filtering, which is defined by the selected resolution bandwidth (RBW). The resolution filter shows the portion of the input signal that was converted to the IF range and is to be displayed at a specific point on the frequency axis.

Due to their steep edge slopes and the resulting spectral selection characteristics, rectangular filters would be very well suited to serve as resolution filters. Due to their long settling times, however, it would only be possible to tune the LO frequency very slowly, or level errors would arise. That translates to long sweep times and slow measurements.

Short measurement periods can be achieved by using Gaussian filters optimized for settling times. Since – unlike rectangular filters – the transition from passband to stopband is not abrupt, a definition must be found for the bandwidth. In general spectral analysis, the 3 dB bandwidth is usually specified. This is the frequency spacing between the two points for the transfer function that exhibit a magnitude reduction of 3 dB compared to the transfer function for the center frequency.

When measuring noise signals or noise-like signals, the levels must be referenced to the measurement bandwidth, i.e. to the resolution bandwidth. For this reason, the equivalent noise bandwidth  $B_{noise}$  for the IF filter must be known, and it is calculated as follows:

$$B_{noise} = \frac{1}{H_{V,0}^2} \cdot \int_{-\infty}^{\infty} H_V^2(f) \cdot df \quad (14)$$

Where

$B_{noise}$	Noise bandwidth
$H_V(f)$	Voltage transfer function
$H_{V,0}$	Value of the voltage transfer function at the center frequency $f_0$

To visualize this, you can think of the noise bandwidth as the width of a rectangle that has the same area as the area beneath the transfer function.

For measurements performed on correlated signals, such as the measurements normally used with radar technology, for instance, the pulse bandwidth  $B_I$  is also of interest. Unlike the noise bandwidth, this bandwidth results from the integration of the voltage transfer function. The following applies:

$$B_I = \frac{1}{H_{V,0}} \cdot \int_{-\infty}^{\infty} H_V(f) \cdot df \quad (15)$$

For Gaussian filters, or Gaussian-like filters, the pulse bandwidth approximately corresponds to the 6 dB bandwidth (which is customary in interference measurement equipment). For test and measurement tasks, the correlations between 3 dB, 6 dB, noise and pulse bandwidths are of particular interest.

According to the Fourier transform, when a sinusoidal input signal is acquired with a spectrum analyzer, this should result in an individual spectral line appearing on the screen at the signal frequency. In reality, however, the resolution filter's transfer function is what appears. The reason for this lies in the fact that the input signal that has been converted to the IF is "swept past" the resolution filter during the sweep period and is multiplied with that filter's transfer function (as in a convolution operation). It is also possible to think of this as the filter being "swept past" a fixed signal as shown in Fig. 15.

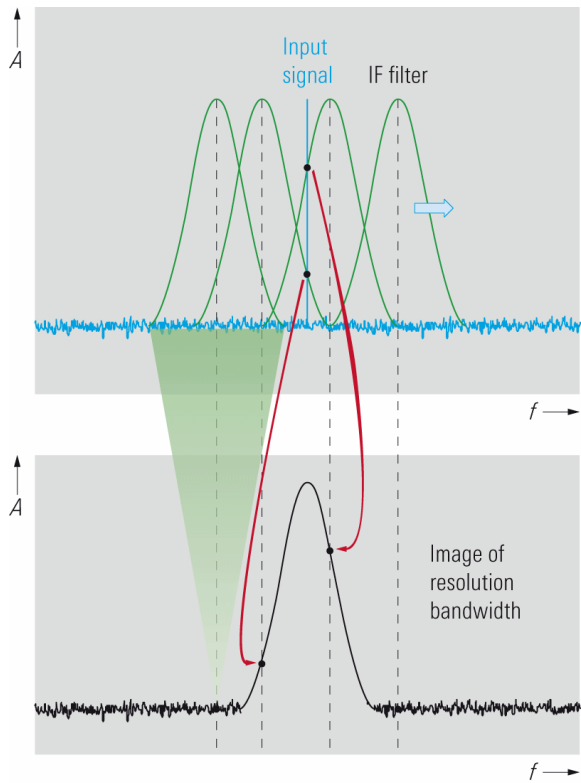


Fig. 15: Displaying the resolution by "sweeping it past" the input signal.

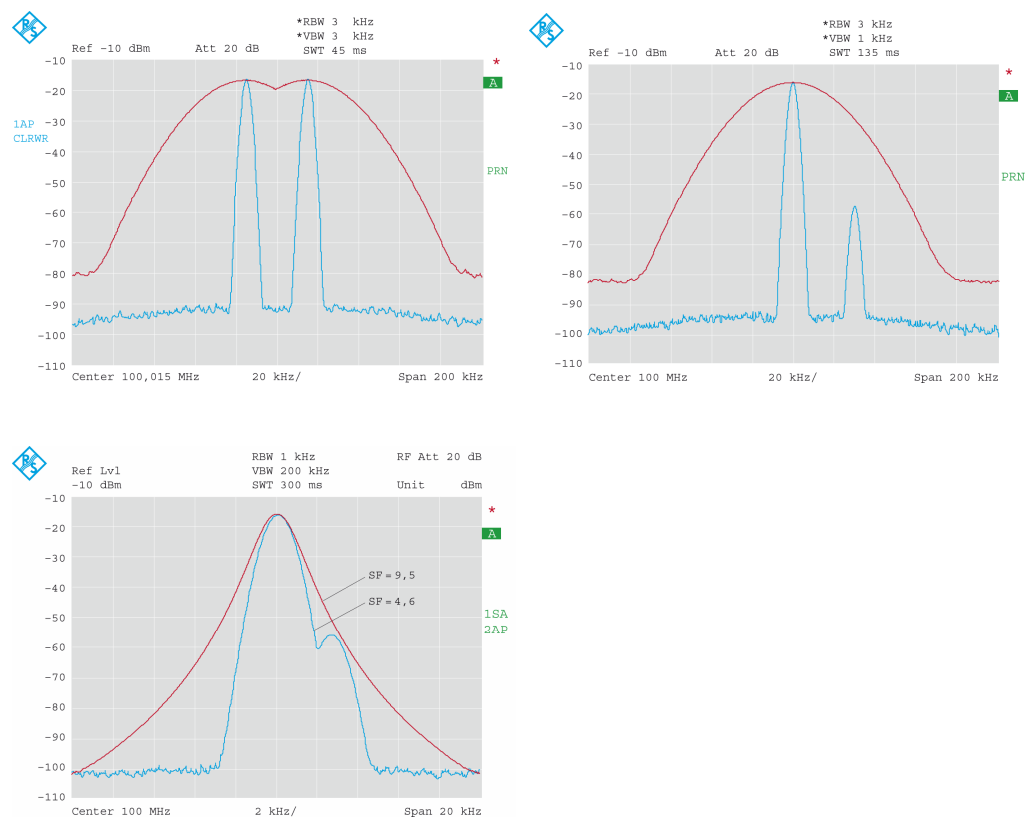
The analyzer's spectral resolution capabilities are, therefore, primarily determined by the resolution bandwidth (the bandwidth of the resolution filter in the IF signal processing stage), which is why these capabilities are also referred to as the resolution bandwidth (RBW). The IF resolution bandwidth (3 dB bandwidth) corresponds to the minimum required difference in frequency that two signals of the same level must exhibit so that they can be distinguished in the display by a dip of about 3 dB. If a significant difference in levels arises between adjacent signals, it is then no longer possible to display the weaker signal if the resolution bandwidth is too large. In order to improve the filter's selectivity, it is possible to reduce the resolution bandwidth, but it is also possible to steepen the solution filter's edge slope. The edge slope is determined by the shape factor, which is calculated as follows:

$$SF_{60/3} = \frac{B_{3dB}}{B_{60dB}} \tag{16}$$

where

- $B_{3dB}$  3 dB bandwidth
- $B_{60dB}$  6 dB bandwidth

Fig. 16 shows how the resolution bandwidth and edge slope affect the results.



**Fig. 16:** How resolution bandwidth and the edge slope affect the representation of two signals.

As the information above shows, the highest resolution is achieved with narrowband resolution filters. Since a narrowband resolution filter has longer settling times, the minimum sweep time must be raised accordingly. This means that it must always be possible to match the resolution capabilities with the measurement speed. For this reason, modern spectrum analyzers must be able to make it possible to set the resolution bandwidth across a wide range (10 Hz to 10 MHz).

State-of-the-art spectrum analyzers indicate when the sweep time has been set too low – i.e. to a point at which the filter no longer has enough time to settle to a steady state.

There are three basic types of filters commonly in use:

#### Analog filters:

Traditionally, spectrum analyzers have been equipped with analog filters for resolution filtering. Since analog filters provide a close approximation of Gaussian filters up to a bandwidth of 20 dB, their settling behavior is almost as good. The selectivity depends on the number of filter circuits. It is possible to achieve shape factors (SFs) of about 10 (this would be 4.6 for an ideal Gaussian filter).

Even advanced spectrum analyzers that already work with digital filters (see below) do not go completely without analog filters. With an analyzer built as depicted in Fig. 13, an analog prefilter with a bandwidth of approx. 3 MHz is switched into the IF3 signal path when small resolution bandwidths are used. This prefilter suppresses large signals that are located outside of the resolution bandwidth being observed. That makes it possible to employ a higher IF gain without overamplifying the ADC. The prefilter also lowers the noise bandwidth, and it suppresses undesired intermodulation products from the upstream mixing stages. These two aspects lead to a larger spurious-free dynamic range (SFDR).

#### **Digital filters:**

Digital signal processing nowadays makes it easy to achieve all required bandwidths, e.g. 1 Hz to over 50 MHz. In this way, it is possible to design ideal Gaussian filters (SF = 4.6) and thus achieve better selectivity than is possible when analog filters are employed (at a reasonable expense). Beyond that, digital filters do not have to be aligned; they remain stable across a range of temperatures, and they are not subject to aging – all of which enables them to achieve greater bandwidth accuracy.

The settling time of a digital filter is always given. Correction calculations make it possible to shorten the sweep time – without changing the resolution bandwidth – to a greater extent than is possible with analog filters.

Since digital filters do not have to be implemented in hardware, a multitude of different filter types can be made available on a spectrum analyzer. For instance, besides Gaussian filters, rectangular filters can also be provided for signal analysis (demodulation).

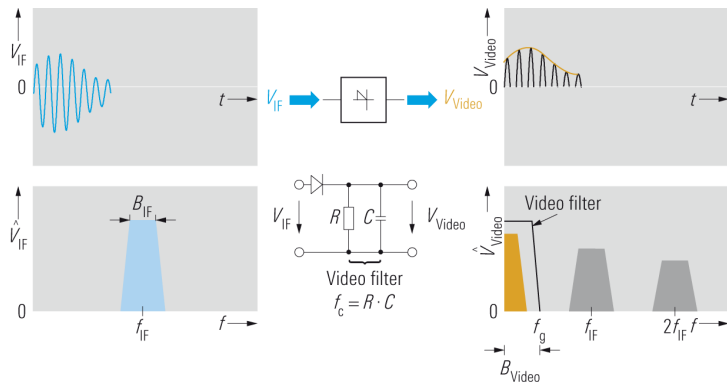
#### **Using FFT:**

This is not filtering in the traditional sense, but rather a combination of a tuned spectrum analyzer and an FFT analyzer. Here, small subranges of the spectrum to be represented are calculated using FFTs. For further details on this operating mode, see section 2.4 on page 31.

### **2.3.5 Envelope Detection and Video Filter**

As with amplitude-modulated signals, the information on the input signal's level is contained in the level of the IF signal, in its envelope. For this reason, after resolution filtering is performed on the last intermediate frequency, the system determines this IF signal's envelope.

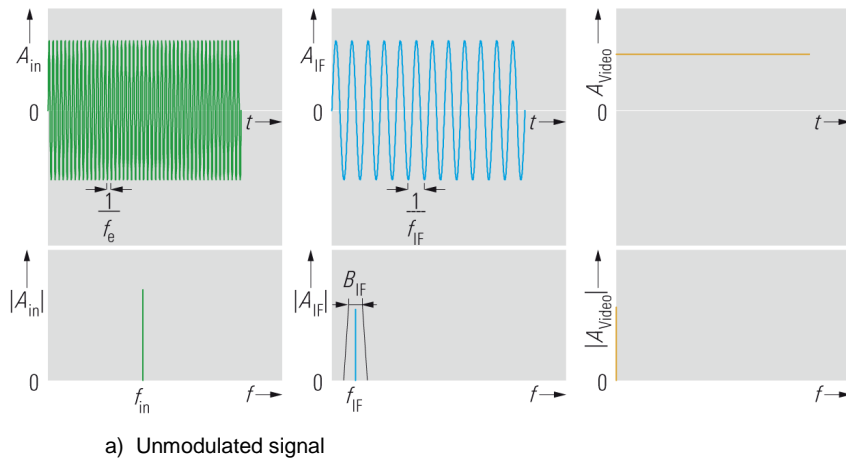
The procedure for doing this is comparable to demodulation of an AM signal, which means that it is possible, for instance, to employ an analog envelope demodulator. Here, the IF signal is rectified, and the RF signal components can be eliminated by a lowpass filter. The "video voltage" then arises at the output.



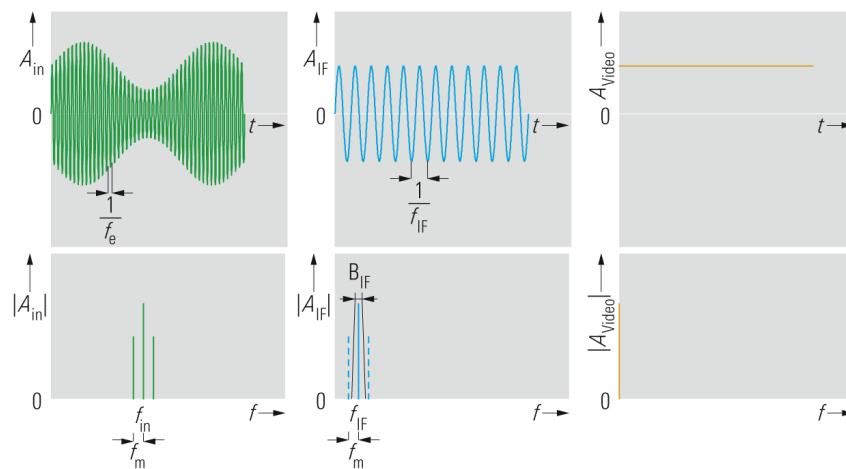
**Fig. 17: Envelope demodulation.**

If the IF signal processing is realized with the aid of a digital filter, the envelope is determined from the digital samples. The IF signal's envelope can be represented as the length of the complex rotating phasor that rotates with  $\omega_{IF}$  (which can only accept discrete values). As noted in section 2.2, the main difference from an FFT analyzer consists in the fact that the phase information is lost when the absolute value is determined.

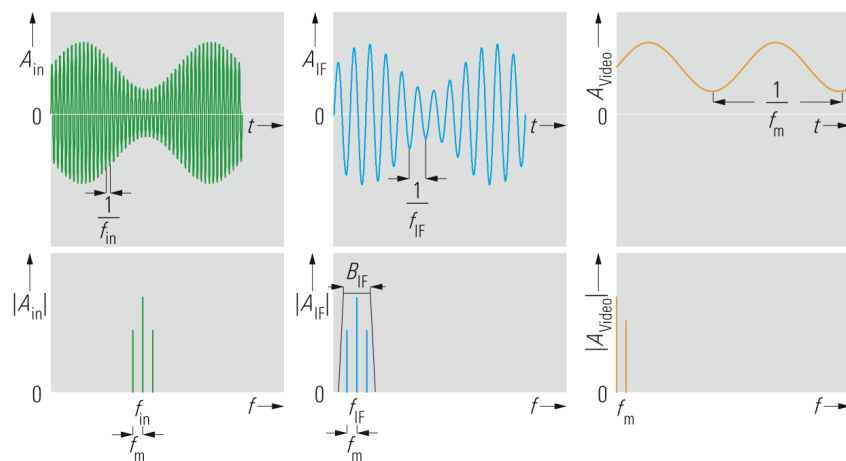
The spectrum analyzer's dynamic range (which is > 100 dB with modern spectrum analyzers) is largely determined by the envelope detector's dynamic range. Simultaneously displaying large differences in levels does not make sense with a linear scale. Consequently, a logarithmic calculation can be optionally be performed with the aid of a log amplifier positioned ahead of the envelope detector, which then increases the display's dynamic range. As Fig. 18 shows, the resulting video signal is dependent on the input signal and on the selected resolution bandwidth.







b) AM signal, resolution bandwidth less than the twice the modulation width



c) AM signal, resolution bandwidth greater than the twice the modulation width

**Fig. 18: For the input signal (green trace), the IF signal after a resolution filter with a specific RBW (blue trace) and the video signal (yellow trace).**

The video signal only contains all of the signal information when the resolution bandwidth large enough.

After the envelope detector follows the "video filter," which is used to establish the video bandwidth (VBW). This filter is a first-order lowpass filter, which frees the video signal of noise. That means that it smooths out the trace that is then displayed later. (With the signal analyzer used in the lab exercises, the video filter is also digital.)

If the video bandwidth is narrower than the resolution bandwidth (RBW), the former determines the maximum sweep speed.

Fig. 18 shows that the video bandwidth should be set to suit the current measurement application:

When measuring sinusoidal signals with a sufficient signal-to-noise ratio, the video bandwidth is selected to be the same as the resolution bandwidth. If the signal-to-noise ratio decreases, the noise can be averaged by reducing the video bandwidth, thus making it possible to achieve a significantly more stable display. (If the video bandwidth is selected to be lower than the bandwidth of the signal to be displayed, the system no longer displays the entire spectrum, which means that information is lost.)

### 2.3.6 Detectors

To display information, modern spectrum analyzers use liquid crystal displays (LCDs) with a discrete number of pixels. Since the LO's tuning step is approximately 1/10 of the resolution bandwidth (see section 2.4.2), and the span is larger than the RBW, multiple measurement results (samples) are available for each pixel. This affects the accuracy of the numerically displayed marker frequencies as well as the accuracy of the displayed measurement results.

The accuracy of the numerically displayed frequency at a marker position (for example with the *Marker to Peak* function) depends on the span and on the selected number of pixels (sweep points). Reducing the span or increasing the number of pixels boosts accuracy. For high precision, the analyzer used in the lab exercises has a *Signal Count* marker function. This function works independently of the evaluation of multiple samples, and it indicates the frequency with a high degree of accuracy.

Which level sample is displayed depends on which detector has been selected; the detector assesses all of the samples responsible for a given pixel. Fig. 19 on page 27 illustrates the different kinds of results that arise in this way.

Most spectrum analyzers have min. peak, max. peak, auto peak, and sample detectors. With the analyzer used in the lab exercise, the detectors have been implemented digitally (the video signal is sampled before it reaches the video filter). Consequently, besides the detectors already mentioned, it was also possible to realize an average detector and an RMS detector as well as a quasi-peak detector (for interference measurements).

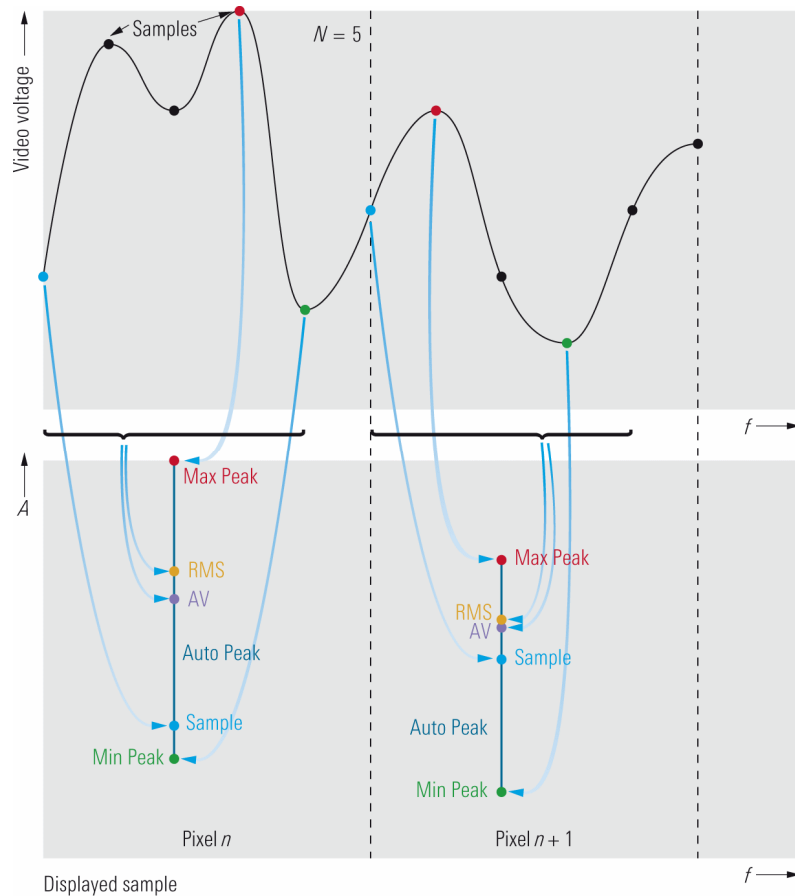
#### **Description of the detector functions:**

Max. peak detector:

The max. peak detector displays the maximum value. Out of all of the samples that are available for a specific pixel, the sample with the largest signal level is selected and displayed. Even when large frequency ranges are displayed with a resolution bandwidth that has been set to be very narrow, no input signals are lost (which is important for EMC measurements).

Min. peak detector:

Out of all of the samples that are allocated to a specific pixel, the min. peak detector displays the one with the lowest level, i.e. it indicates the minimum value.



**Fig. 19: Selection of the samples to be displayed based on the detector that has been chosen.**

Auto peak detector:

When the auto peak detector is used, the maximum value and minimum value are displayed simultaneously. Both values are measured and then indicated with a vertical line that connects the two values.

Sample detector:

At a constant time interval, the sample detector takes one of the samples assigned to a particular pixel, which means that it only samples the IF envelope once for each pixel of the trace to be displayed. When the frequency range to be displayed is much larger than the resolution bandwidth, the input signals are no longer being acquired reliably.

RMS detector (root mean square, average value):

The RMS detector uses the corresponding samples to calculate the power for every pixel of the displayed trace. The envelope samples are required in a linear level scale in order to perform this calculation. The following applies:

$$V_{\text{RMS}} = \sqrt{\frac{1}{N} \cdot \sum_{i=1}^N v_i^2} \quad (17)$$

Where

$V_{\text{RMS}}$  = RMS value for the voltage

$N$  = Number of samples that are allocated to a pixel

$v_i$  = Samples for the envelope

The power at the reference impedance is:

$$P = \frac{V_{\text{RMS}}^2}{R} \quad (18)$$

AV detector (average):

The AV detector calculates the linear average for each pixel of the displayed trace from the corresponding samples. The envelope samples are required in a linear level scale in order to perform this calculation. The following applies:

$$V_{\text{AV}} = \frac{1}{N} \cdot \sum_{i=1}^N v_i \quad (19)$$

$V_{\text{AV}}$  Average voltage

Quasi-peak detector:

The quasi-peak detector detects the peaks for interference measurements with a defined charge and discharge time. It is used to measure electromagnetic interference.

### How detectors affect representation of different input signals:

Depending on the input signal, the different detectors lead to different measurement results. For sinusoidal input signals with a sufficient signal-to-noise ratio, the video voltage is constant. The level of the displayed signal is, therefore, independent of the selected detector, because all samples have the same level, and the values (RMS, AV) that are calculated from them correspond to the level of each individual sample. In the case of random signals, such as noise or noise-like signals, the instantaneous power varies over time, which means that the maximum and minimum instantaneous value as well as the average and RMS value for the envelope differ in this case. The power for signals with a limited observation period  $T$  is calculated as follows:

$$P = \frac{1}{R} \cdot \frac{1}{T} \cdot \int_{t-T/2}^{t+T/2} v^2(t) \cdot dt \quad (20)$$

Furthermore, during the observation period  $T$  it is also possible to find the peak for the instantaneous power and then use that value to calculate the crest factor.

$$CF = 10 \cdot \lg\left(\frac{P_{\text{peak}}}{P_{\text{RMS}}}\right) \text{dB} \quad (21)$$

Where

- $CF$  Crest factor
- $P_{\text{peak}}$  Peak power during the observation period  $T$ , in W
- $P_{\text{RMS}}$  RMS power

With a pure noise signal, it is theoretically possible for all voltages to occur (the crest factor can be of any size). Nevertheless, the probability that very high or very low voltage values will arise is very low. Consequently, in practice, values that can be displayed are achievable when the observation periods are long enough (for example: crest factor = 12 dB for Gaussian noise).

### How the selected detector and the sweep time influence the display of stochastic signals:

Max. peak detector:

The system responds too strongly to stochastic signals; they result in the highest level display. If the sweep time increases, the dwell time in a frequency range that is assigned to a pixel rises. This increases the probability of higher instantaneous values arising, and the levels of the displayed pixels rise.

Using short sweep times delivers the same display as with the sample detector, because only one sample is recorded per pixel.

Min. peak detector:

This can be thought of in the same way as the max. peak detector, but for low signal levels.

Auto peak detector:

The results of the max. peak and min. peak detectors are connected with a line and displayed simultaneously. If the sweep time increases, this causes the displayed noise bandwidth to become significantly larger.

#### Sample detector:

Since it is always the case that only one sample is used at a defined time, the displayed trace varies due to the distribution of the instantaneous value around the average value for the envelope of the IF signal that results from the noise. In the case of Gaussian noise, this average value is 1.05 dB below the RMS value (also: using a narrow video bandwidth in the logarithmic scale results in display values that are lower by an additional 1.45 dB). Thus, the displayed noise is a total of 2.5 dB below the RMS value.

With this detector, changing the sweep time does not affect the display, because the number of evaluated samples is constant.

#### RMS detector:

An input signal's actual power can be measured independently of its time variation. When the signal power is determined using data from the sample detector or max. peak detector, the parameters for signal statistics must be known when working with stochastic signals. This information does not have to be known when the RMS detector is used.

Increasing the sweep time also causes an increase in the number of samples that enter the calculation, which smooths out the displayed trace. Since averaging through a narrow video bandwidth leads to falsification of the RMS display, the video bandwidth must be at least three times as large as the resolution bandwidth.

#### AV detector

In the linear level scale, an input signal's actual average value can be measured independently of its time variation. When averaging logarithmic values, the display level determined in this way would be too low, because higher signal values are subjected to greater compression. Raising the sweep time makes more samples available per pixel for the calculation, which smooths out the displayed trace.

Using linear levels at the video filter's input and reducing the video bandwidth will smoothen the signal.

As mentioned above, reducing the video bandwidth makes it possible to smooth out the trace display through averaging. The prerequisite for this, however, is that the signal levels ahead of the envelope detector or video filter must be in the linear scale. The resulting display then represents the actual average value. If, on the other hand, the IF signal is logarithmized, the displayed average value is lower than the actual average value.

#### Marker functions

The *Marker to Peak* and *Signal Count* marker functions make it possible to work around the fact that the screen resolution is limited. This makes it possible to read out measurement results at a substantially higher resolution.

## 2.4 Combining Both Implementation Approaches

As explained in sections 2.2 and 2.3, both FFT analyzers and spectrum analyzers that use the heterodyne principle offer specific advantages. The benefits of using an FFT analyzer are:

High measurement speeds at low resolution bandwidths.

Recording of the signal in the time domain with all of the phase information. This makes it possible to also analyze complex modulations (signal analysis).

Spectrum analyzers that employ the heterodyne scheme offer the advantage that the input frequency range is independent of the A/D converter rate.

When preselection is used, it is possible to achieve excellent suppression of harmonics and of other undesired spectral components.

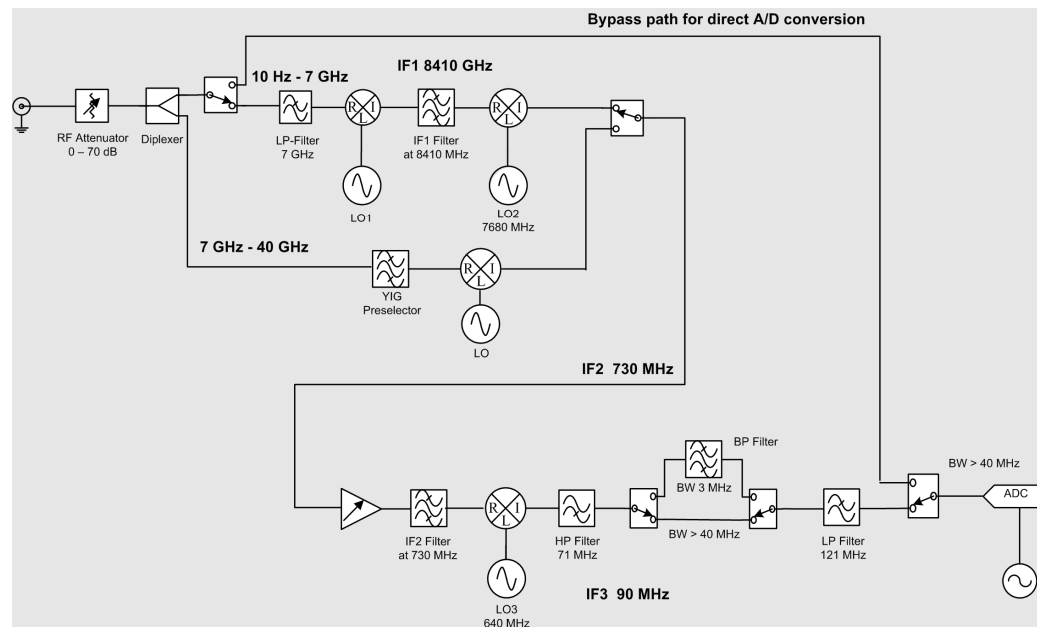
It is possible to secure all these advantages together by skillfully combining an FFT analyzer with a traditional spectrum analyzer. One of the key features of modern analyzers is that many of the processing steps performed by traditional analog spectrum analyzers have now been digitalized, meaning that they are implemented in software or digital hardware (such as an FPGA or ASIC). In order to provide sufficient dynamic range, ADCs that allow a high quantization depth are employed for this.

Fig. 20 on page 32 shows the analog portion of a modern analyzer. Its functions correspond to those of a heterodyne spectrum analyzer – up until the last IF stage. After that, further processing is accomplished digitally.

As with FFT analyzers, a sampled time-domain signal is made available after A/D conversion. That opens up the possibility for signal analysis – i.e. for demodulation of the signal. (The IF3 signal's bandwidth amounts to over 40 MHz. As a result, it is possible to acquire data in the formats employed for all commonly used communications standards and then demodulate and analyze that information using the corresponding software options. For this reason, modern spectrum analyzers are often referred to as "signal and spectrum analyzers." Nevertheless, the focus here will remain on the topic of "spectral analysis.")

The ADC in Fig. 20 on page 32 does not sample a baseband signal, but rather an IF signal. The system performs bandpass sampling, which means that it samples a signal associated with bandwidth  $B$ . The sampling rate used here can even be lower than the level of twice the largest frequency that arises ( $IF3+B/2$ ). Nonetheless, the sampling rate must at least meet the Nyquist criterion for the signal bandwidth (i.e. it must be larger than  $2*B$ ). For the analyzer outlined in Fig. 20, the bandpass filtering is realized prior to sampling through use of the 71 MHz highpass filter and the 121 MHz lowpass filter.

The result of the bandpass sampling is a time-discrete and value-discrete IF signal. In another processing step, "digital down conversion" is used to generate a complex baseband signal from this digital IF signal.



**Fig. 20: Block diagram of the analog portion of a modern signal and spectrum analyzer.**

The complex baseband signal contains a relative phase. Here, having a relative phase means that it is not possible to draw conclusions about the absolute phase value, but the phase relationships within the signals remain constant.

There are two possibilities for preparing a frequency-domain display from the time-domain signal:

1. Using digital filters with the RBW bandwidth: The magnitude of the signal filtered in this way now corresponds to the power within the RBW; in other words, this is the exact value that is to be displayed for the current input frequency. That corresponds to the way that an analog spectrum analyzer works, whereby the filtering and formation of the absolute value are accomplished digitally. Digital filters with low delay distortion can be designed in such a way that they settle to a steady state at a speed that is approximately 100 times faster than the corresponding analog filters achieve. Nevertheless, the sweep speed's dependency  $\text{span}/\text{RBW}^2$  remains.
2. Calculation of an FFT: Here, the calculation and recording parameters are set so that the FFT's resolution corresponds exactly to the RBW setting. Since no narrowband filters are used for this, the long settling time that narrow filters have does not dominate the sweep speed. Here, the maximum FFT width is limited by the analyzer's IF bandwidth (B); thus, in the case under consideration here, it is approximately 40 MHz. The observation period, i.e. the length of the recording, determines the RBW that can be achieved for an FFT.

In both cases it is necessary to "run through" the frequency range that has been set on the analyzer. With the first variant, this is done in very small steps:  $f_{\text{Step}} \ll \text{RBW}$ . That corresponds to the method that has been described for traditional spectrum analyzers.



With the second variant, it is possible to select the step width up to a size as large as the RBW. This means that the number of span/RBW FFT calculations can cover a specific span. Consequently, the sweep time is now no longer proportional to  $\text{span}/\text{RBW}^2$ . For this reason, the variant for settings with a small RBW is particularly well suited for this task.

Modern spectrum analyzers take advantage of the fact that it is also possible to switch the input signal directly to the ADC. The bypass in Fig. 20 is meant to accomplish this. Such direct sampling offers the advantage that neither mixers nor local oscillators influence the signal to be measured. This concept offers special advantages for noise and phase noise measurements: By using the direct path, it is possible to use a mid-range spectrum analyzer to measure signals that have a phase noise less than  $-130$  dBc/Hz at a 10 kHz offset.

Direct sampling is, however, restricted to frequencies lower than half the sampling rate for the ADC being used. Depending on the settings, the spectrum analyzer used for the lab exercises employs this possibility up to frequencies that are just higher than 20 MHz.

## 2.5 Important Terms and Settings

Frequency range to be displayed:

The frequency range to be displayed can be set using either the start and stop frequencies or using the center frequency and the span.

Level range to be displayed:

This range is set by establishing the maximum level to be displayed, which is referred to as the reference level, and by establishing the span. The levels can be displayed in the linear or logarithmic scale. The damping of the attenuator on the input end also depends on these settings.

Attenuator:

In order to be able to display high signal levels, the spectrum analyzer's input has been equipped with an attenuator that is set in steps. This attenuator can be used to set the signal level at the input for the first mixer – the mixer level.

Frequency resolution:

With analyzers that employ the heterodyne principle, the frequency resolution is set via the resolution filter's bandwidth (in the IF signal processing section). It is referred to as the resolution bandwidth (RBW).

Sweep time (only for analyzers that employ the heterodyne principle):

This is the time required to record the entire relevant frequency spectrum. The shortest sweep time is set automatically by the analyzer software, based on the selected bandwidths (RBW and VBW).

**Dependencies for the sweep time, span or resolution bandwidth and video bandwidth:**

The shortest permissible sweep time is derived from the settling time of the resolution filter and the video filter. The video filter only influences the sweep time if the video bandwidth is smaller than the resolution bandwidth. This is expressed by the following equation:

$$T_{sweep} = k \cdot \frac{\Delta f}{B^2} \quad (22)$$

Where

$T_{sweep}$  Minimum required sweep time (for the given span and RBW), in s

$B$  RBW if  $RBW \leq VBW$   
VBW if  $VBW \leq RBW$

$\Delta f$  Frequency range to be displayed (span)

$k$  Factor, depending on the type of filter and on the required level accuracy.

## 3 Generators and Their Use

RF generators are classified into two main types:

- Analog signal generators
- Vector signal generators

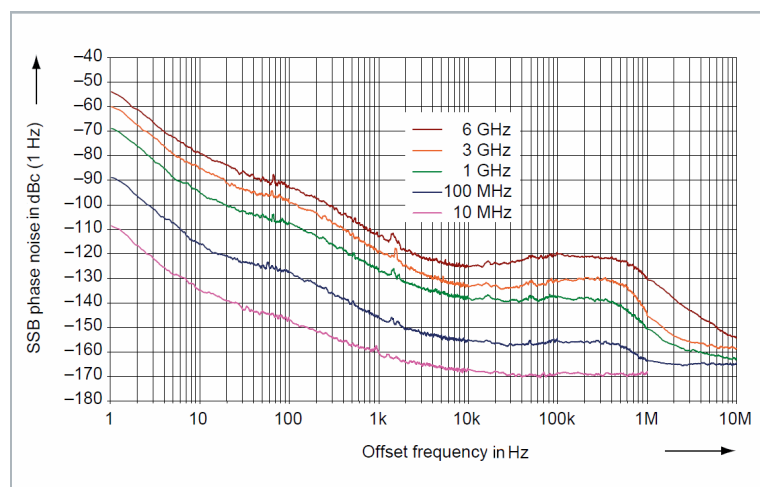
Besides those two, there are many other possible classifications based on various characteristics – e.g. according to frequency range or output power, according to the form factor, capabilities for remote control, power supply, etc. This document will not examine those classifications.

Analog and vector signal generators generate their output signals in completely different ways. This results in different modulation types and different applications.

### 3.1 Analog Signal Generators

With analog signal generators, the focus is on generating a high-quality RF signal. These devices support the analog modulation types: AM / FM and  $\phi$ M. Some devices can also be used to generate precise pulsed signals. Analog generators are available for frequencies extending up to the microwave range. Their distinguishing features are:

- Very high spectral purity (nonharmonics), e.g.  $-100$  dBc
- Very low inherent broadband noise, e.g.  $-160$  dBc
- Very low SSB phase noise, e.g.  $-139$  dBc (1 Hz)  
(at these settings: 20 kHz carrier offset,  $f = 1$  GHz, 1 Hz measurement bandwidth)

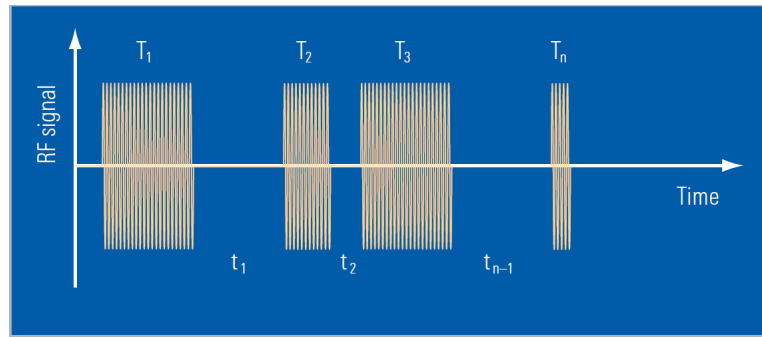


**Fig. 21: An analog signal generator's very low SSB phase noise.**

Analog signal generators are used:

- As stable reference signals (local oscillator, source for measuring phase noise, or as a calibration reference)
- As a universal instrument for measuring gain, linearity, bandwidth etc.
- In the development and testing of RF chips and other semiconductor chips, such as those used for A/D converters
- For receiver tests (two-tone tests, generation of interferer and blocking signals)
- For EMC tests
- For automated test equipment (ATE) and production
- For avionics applications (such as VOR, ILS)
- For military applications
- For radar tests

Fig. 22 shows an example of a special impulse sequence for radar applications:



**Fig. 22: Combination of impulses with different widths and interpulse periods for radar applications.**

Analog signal generators are available with different specifications in all price classes. As with vector signal generators, additional criteria can be crucial for making the right selection. These include, for instance, requirements for a high output power or for fast settling of frequency and level, a specific degree of accuracy for the signal level and frequency, a low VSWR, and possibly also the instrument's form factor and weight.

## 3.2 Vector Signal Generators

Vector signal generators are distinguished by the fact that they generate and process the modulation signal computationally in the baseband as a complex IQ data stream. This also includes computational filtering, and (if necessary) limitation of the amplitude (clipping); it can also include other capabilities, such as generating asymmetric characteristics. Some generators can calculate Gaussian noise into the signal. Moreover, some generators are able to numerically simulate multipath propagation (fading, MIMO) that will later occur for the RF signal.

In general, the complete generation of the baseband signal is accomplished through realtime computation. ARB generators are an exception to this (see section 3.3).

The baseband IQ data is ultimately converted to an RF operating frequency. (There are also vector generators that operate exclusively in the baseband without generating RF signals.)

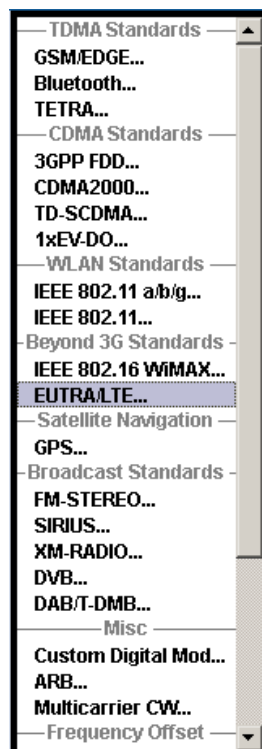
Often, vector signal generators are also equipped with analog or digital IQ inputs to make it possible to feed external baseband signals into the instrument.

Using IQ technology makes it possible to realize any modulation types – whether simple or complex, digital or analog – as well as single-carrier and multicarrier signals. The requirements that the vector signal generators must meet are primarily derived from the requirements established by wireless communications standards, but also from digital broadband cable transmission and from A&D applications (generation of modulated pulses).

The main areas of application for vector signal generators are:

- Generating standards-compliant signals for wireless communications, digital radio and TV, GPS, modulated radar, etc.
- Testing of digital receivers or modules in development and manufacturing
- Simulating signal impairments (noise, fading, clipping, insertion of bit errors)
- Generating signals for multi-antenna systems (multiple in / multiple out, MIMO), with and without phase coherence for beam forming
- Generating modulated sources of interference for blocking tests and for measuring suppression of adjacent channels

As an example, Fig. 23 shows a portion of the preprogrammed standards that a vector signal generator supports:



*Fig. 23: Preprogrammed standards for a vector signal generator.*

The individual communications standards generally specify test signals with a defined parameter configuration. In a vector signal generator, these signals can be preprogrammed. Fig. 24 shows a selection of these "test models" of the LTE standard (for the same generator).

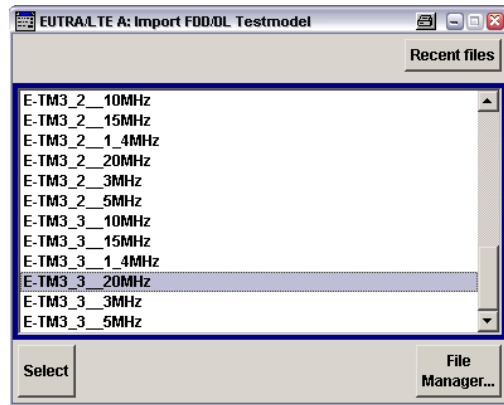


Fig. 24: Some of the preprogrammed test models for the LTE wireless communications standard.

Fig. 25 shows the spectrum for the E-TM3\_3\_\_20MHz test model that has been selected in Fig. 24.

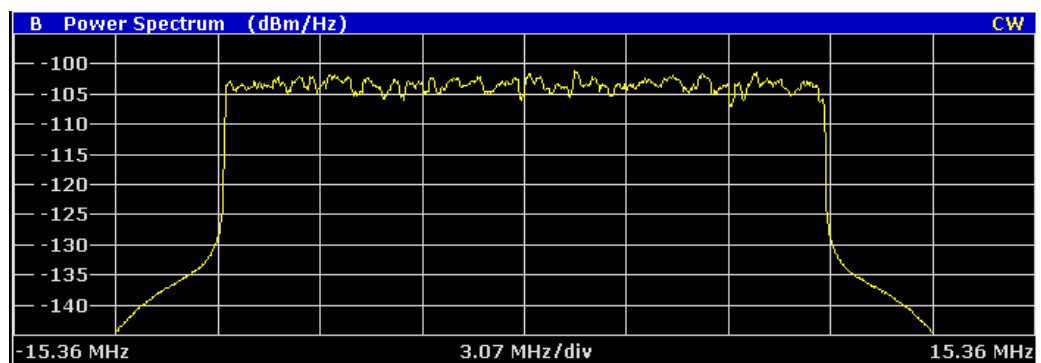


Fig. 25: Multicarrier spectrum for the E-TM3\_3\_\_20MHz test model from the LTE standard.

The spectrum is approx. 18 MHz wide. A closer examination reveals that it consists of 1201 OFDM single carriers which are each spaced apart by 15 kHz, but merge into each other in this display due to the screen-resolution setting.

Fig. 26 shows the test model's constellation diagram (IQ display).

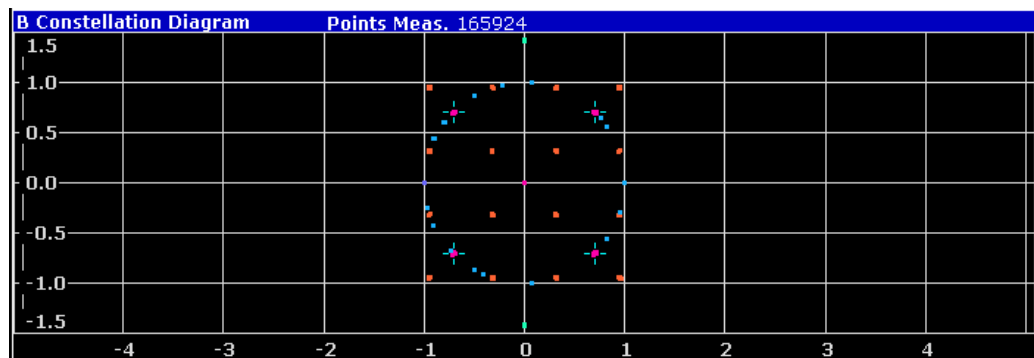


Fig. 26: Overall constellation for the E-TM3\_3\_\_20MHz LTE test model.

With the signal used in this case, the individual channels are modulated differently. Here, all of the modulation types that are used are summarized in one representation: BPSK (cyan), QPSK (red with blue crosses), 16 QAM (orange) and the constant amplitude zero autocorrelation (CAZAC, blue) bits that are typical for LTE on the unit circle.

Vector signal generators usually provide convenient triggering capabilities. This makes it possible, for example, to fit generator bursts precisely into a prescribed time grid (such as putting GSM bursts into the right time slots).

In parallel with the data stream, the generators generally also supply what are known as marker signals at the device's jacks. These signals can be programmed for activation at any position in the data stream (for example, at the beginning of a burst or frame), in order to control a DUT or measuring instruments.

Unlike analog signals, digitally modulated signals sometimes have very high crest factors. This means that the ratio between the average value and peak value can often be more than 10 dB. Even small nonlinearities in the generator's amplifiers, mixers and output stages can cause harmonics and intermodulation products more easily. In this respect, there are considerable differences in the quality of individual generators.

Important characteristics for vector signal generators are the modulation bandwidth and the achievable symbol rate, the modulation quality (error vector magnitude, EVM) and the adjacent channel power (ACP). State-of-the-art generators are prepared to meet future requirements, which means that they exceed the requirements of the wireless communications standards that are currently in common use by a significant margin.

General criteria for selecting the instrument are – as with analog generators – for example: the required output power, the settling time and accuracy for frequency and level, low VSWR, and sometimes the instrument's form factor and weight.

### 3.3 Arbitrary Waveform Generators (ARBs)

Arbitrary waveform generators (ARBs) are vector signal generators for which the modulation data is calculated in advance (rather than in realtime) and stored in the instrument's RAM. This RAM content is then read out at the realtime symbol rate. (Many vector signal generators are equipped with an ARB option, see the menu list in Fig. 23.)

With regard to their use and applications, ARB generators differ from realtime vector generators on the following points:

- There are no restrictions at all for configuring the content of an ARB's IQ-data stream
- It is only possible to use time-limited or periodically repeated signals (the RAM's depth is finite)

The IQ data sets' memory depth and word size are additional characteristics for ARBs.

As with realtime generators, there are different triggering options and the ability to output "marker signals" for controlling hardware and measuring instruments that are connected to the device.

For production tests, users can concatenate various sequences of different lengths. For example, this could be data streams with different bit rates that have to be checked during the manufacturing process.

Some ARB generators can computationally generate additional Gaussian noise; some are also able to simulate multipath propagation (fading) or multi-antenna systems (MIMO). In all cases, this is done in realtime in the baseband.

In many cases, vendors that offer ARB generators also offer software for creating standard modulation sequences (IQ datasets). As an example, Fig. 27 shows several windows from such a program.

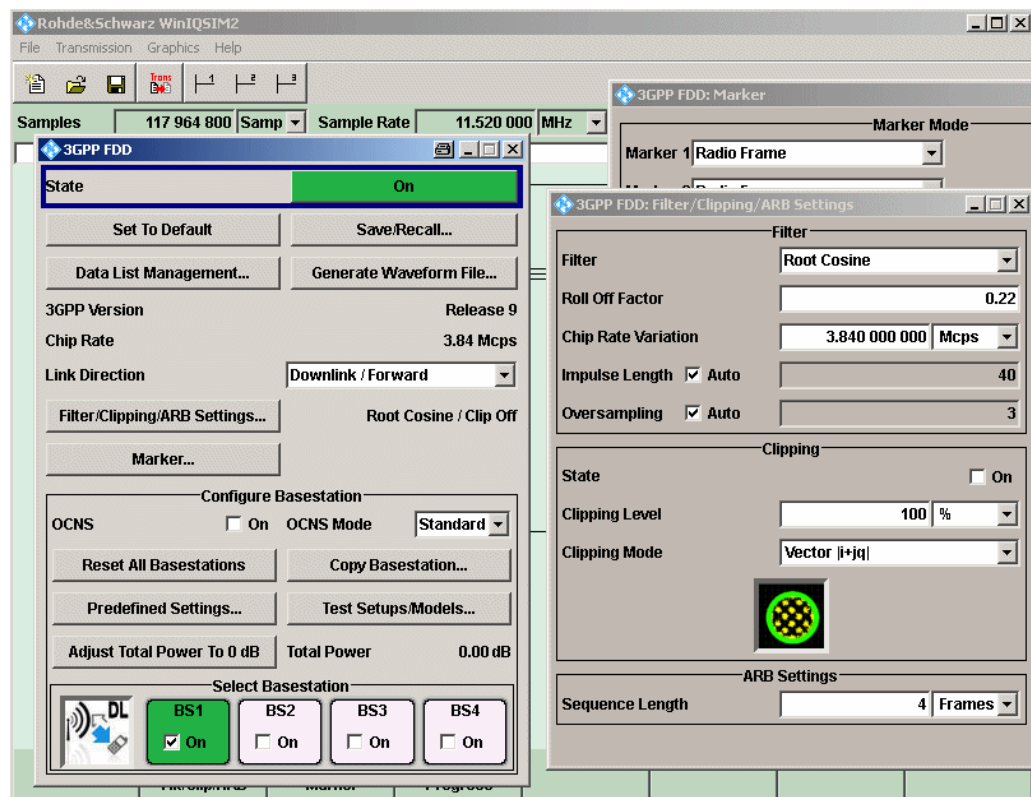


Fig. 27: PC program for calculating the IQ data for standard signals.

In the example shown here, the 3GPP FDD (UMTS) wireless communications standard has been selected. The system is creating a downlink, which is the signal from a base station (BS) to a mobile phone. The program can generate the signals from up to four base stations; in Fig. 27, only BS1 is active. The filtering complies with the UMTS standard. No *Clipping* is performed. Later, the *Marker1* device jack will deliver a signal with each new *Radio Frame*.

Once all the required entries have been made, the user initiates calculation of the IQ data by clicking the *Generate Waveform File* button. Once that step is completed, the data is transferred from the program to the ARB, and the output can be started immediately.



## 4 The DUT's Nonlinearities

For the most part, the structure and content of this chapter have been taken from the lecture notes on practical exercises entitled "Hochfrequenztechnik Labor, Spektrumanalysator" from the University of Graz, Austria [2]. Further information is available, for instance, in the white paper "Interaction of Intermodulation Products between DUT and Spectrum Analyzer" [4].

An ideal two-port device transfers signals from the input to the output without distorting them. The output signal follows any variation in the input signal in strict proportion. Only the same frequencies that are fed into the input arise at the output. The device under test (DUT), a real amplifier, is not an ideal two-port:

1. As the input power rises, the effective power gain decreases.
2. In addition, the higher-order harmonics and the intermodulation products from the input frequencies and their harmonics rise at the output.

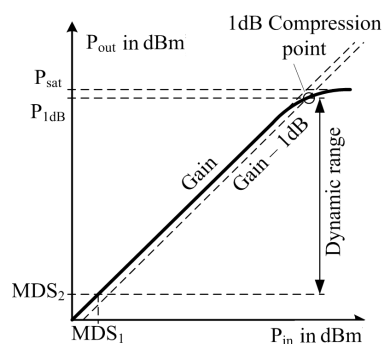
This first observation describes an amplifier's large-signal behavior. Above a certain input level, the amplifier reaches saturation. The maximum permissible input level is defined by the 1 dB compression point.

The second observation describes the small-signal distortions that always arise. Even when driven with small signals, an amplifier's characteristic curve is never purely linear. The fact that spectral components that were not present in the input signal arise in the output signal can be substantiated mathematically. Specifying intercept points makes it possible to compare amplifiers and properly dimension hardware setups.

(The fact that new spectral components arise at nonlinearities is, to some extent, employed intentionally in RF engineering – for example, to multiply frequencies on a diode characteristic curve or to convert frequencies in mixers.)

### 4.1 The 1 dB Compression Point

An amplifier's "1 dB compression point" is defined as the output power at which the nominal gain has dropped by 1 dB. Correspondingly, the output power has decreased by 1 dB compared to the nominal output power. In Fig. 28, the 1 dB compression point is located at  $P_{1dB}$ .



**Fig. 28:** Correlation between the input power ( $P_{in}$ ) and output power ( $P_{out}$ ) of an amplifier (logarithmic scale) with the 1 dB compression point.

The low end of the linear dynamic range is limited by the smallest signal power that can be differentiated from noise at the system's input. There must be a "minimum detectable signal"  $MDS_1$  present that is  $x$  dB larger than the unavoidable thermal noise floor ( $-174$  dBm per Hz at 290 kbit). For most applications, the minimum signal power should be twice as large as the noise power, thus:  $x = 3$  dB. When, however, we take the amplifier's actual bandwidth  $B$  and noise factor  $F$  into account, we arrive at the following:

$$MDS_1 = -174 \text{ dBm} + 10 \lg B \text{ dB} + 10 \lg F \text{ dB} + x \quad (23)$$

The following then holds true for the power that arises at the output in combination with gain  $G$ :

$$MDS_2 = -174 \text{ dBm} + 10 \lg B \text{ dB} + 10 \lg F \text{ dB} + x + 10 \lg G \text{ dB} \quad (24)$$

When the level is rising, the 1 dB compression point defines the end of the linear area. The output power  $P_{1\text{dB}}$  supplied at this point is normally indicated in product data sheets as the maximum power that the amplifier can deliver.

$$P_{1\text{dB}} = P_{\text{in } 1\text{dB}} + 10 \lg G \text{ dB} - 1 \text{ dB} \quad (25)$$

The amplifier's linear dynamic range  $D$  is now the decrease in amplitude between the 1 dB compression point and the power generated at the output  $MDS_2$  [2]:

$$D = P_{1\text{dB}} + 174 \text{ dBm} - 10 \lg B \text{ dB} - 10 \lg F \text{ dB} - x - 10 \lg G \text{ dB} \quad (26)$$

The more an amplifier is operated in the range that is no longer linear, the more the output power is spread into harmonics and intermodulation products. Power is then "shifted" to other spectral components than the ones that originally had the power. (This topic is covered in the introduction to chapter 4 Nonlinearities.) For this reason, the 1 dB compression point must be determined in a frequency-selective manner using a spectrum analyzer. In such cases, a broadband thermal power meter registers the entire spectrum and delivers incorrect results.

## 4.2 Mathematic Description of Small-Signal Nonlinearities

One simple method that can be used to describe a nonlinear two-port is to express the output voltage  $v_a(t)$  as a power series for the input voltage  $v_{in}(t)$ :

$$v_{out}(t) = \sum_{n=1}^{\infty} a_n \cdot v_{in}^n(t) = a_1 \cdot v_{in}(t) + a_2 \cdot v_{in}^2(t) + a_3 \cdot v_{in}^3(t) + \dots \quad (27)$$

Where

$v_{out}(t)$  Signal at the output port

$v_{in}(t)$  Signal at the input port

$a_n$  Coefficient

In general,  $v_{in}(t)$  is any band-limited signal in the time domain. If  $v_{in}(t)$  is periodic, it can be described as the sum of sinusoidal signals with different amplitudes and frequencies; see section 2.1 on page 6.

Consequently, as a simplification, for the following derivatives,  $v_{in}(t)$  is to first consist of two sinusoidal signals with the amplitudes  $V_1$  and  $V_2$  and the frequencies  $\omega_1$  and  $\omega_2$ :

$$v_{in}(t) = V_1 \cdot \sin(\omega_1 \cdot t) + V_2 \cdot \sin(\omega_2 \cdot t)$$

If this is plugged into the equation (27), the following results are reached when the angle theorems are applied:

$$\begin{aligned}
 v_{out}(t) = & \frac{1}{2} \cdot a_2 \cdot (V_1^2 + V_2^2) && \text{DC component} \\
 & + (a_1 \cdot V_1 + \frac{3}{4} \cdot a_3 \cdot V_1^3 + \frac{3}{2} \cdot a_3 \cdot V_1 V_2^2) \cdot \sin(\omega_1 \cdot t) && \text{Fundamental (first harmonic)} \\
 & + (a_1 \cdot V_2 + \frac{3}{4} \cdot a_3 \cdot V_2^3 + \frac{3}{2} \cdot a_3 \cdot V_1^2 V_2) \cdot \sin(\omega_2 \cdot t) \\
 & - \frac{1}{2} \cdot a_2 \cdot V_1^2 \cdot \cos(2 \cdot \omega_1 \cdot t) && \text{Second harmonic} \\
 & - \frac{1}{2} \cdot a_2 \cdot V_2^2 \cdot \cos(2 \cdot \omega_2 \cdot t) \\
 & + a_2 \cdot V_1 \cdot V_2 \cdot \cos((\omega_2 - \omega_1) \cdot t) && \text{Second-order} \\
 & - a_2 \cdot V_1 \cdot V_2 \cdot \cos((\omega_2 + \omega_1) \cdot t) && \text{intermodulation products} \\
 & - \frac{1}{4} \cdot a_3 \cdot V_1^3 \cdot \sin(3 \cdot \omega_1 \cdot t) && \text{Third harmonic} \\
 & - \frac{1}{4} \cdot a_3 \cdot V_2^3 \cdot \sin(3 \cdot \omega_2 \cdot t) \\
 & + \frac{3}{4} \cdot a_3 \cdot V_1^2 V_2 \cdot \sin((2 \cdot \omega_1 - \omega_2) \cdot t) && \text{Third-order} \\
 & - \frac{3}{4} \cdot a_3 \cdot V_1^2 V_2 \cdot \sin((2 \cdot \omega_1 + \omega_2) \cdot t) && \text{intermodulation products} \\
 & + \frac{3}{4} \cdot a_3 \cdot V_1 V_2^2 \cdot \sin((2 \cdot \omega_2 - \omega_1) \cdot t) \\
 & - \frac{3}{4} \cdot a_3 \cdot V_1 V_2^2 \cdot \sin((2 \cdot \omega_2 + \omega_1) \cdot t) \quad \dots && (28)
 \end{aligned}$$

The series is ended here after the third power.

This shows that other spectral components arise in addition to the fundamental: A DC component, harmonics with a multiple of the fundamental frequency, plus "intermodulation products" at the sums and differences of the fundamental frequencies and harmonics.

The order number for these spectral lines is defined by the number of terms required for the calculation. This calculation, for instance:

$$2\omega_1 - \omega_2 = \omega_1 + \omega_1 - \omega_2$$

requires three terms; consequently, it is a third-order frequency.

When the stimulus signal has only one sine wave ( $V_1$  or  $V_2$  is equal to zero), there are no intermodulation products.

The frequencies that arise follow the rules for forming Pascal's Triangle:

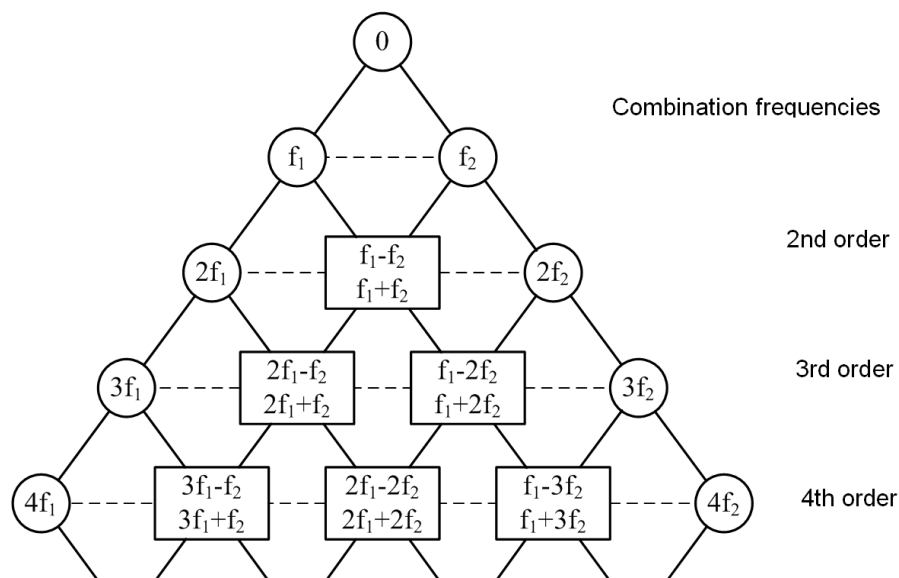


Fig. 29: A version of Pascal's triangle for forming the combination frequencies.

The polynomial's degree, *deg* (Eq. 27) determines the highest order of the harmonics that arise, or of the combination frequencies. The following applies:  $|n_1| + |n_2| \leq G$ .

Where

$n_1$ ..... Number of terms  $f_1$

$n_2$ ..... Number of terms  $f_2$

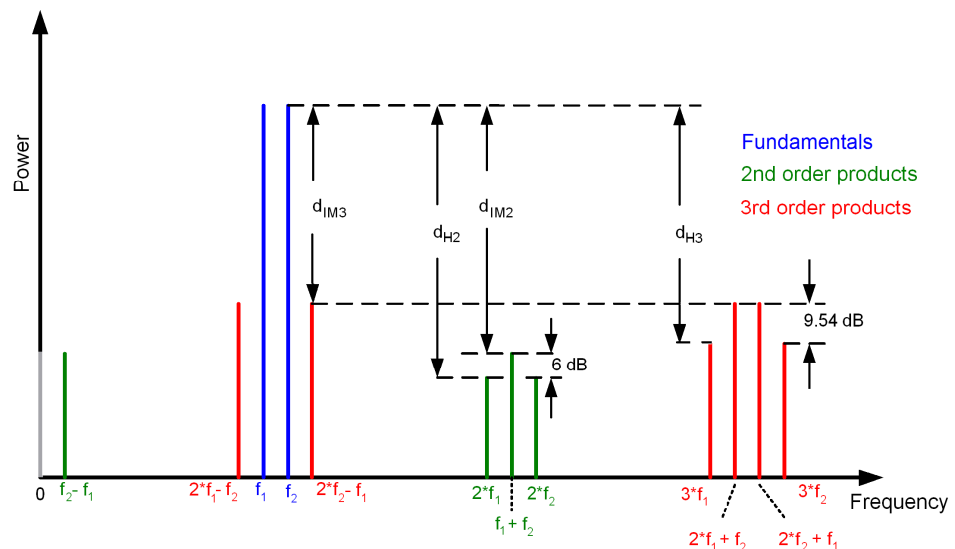
Eq. (28) was derived for voltages. One obtains the output power, when  $V$  is substituted by  $\frac{V^2}{R}$ . Eq. (28) on page 42 maintains its form.

When comparing the coefficients of the individual spectral components in Eq. (28), it becomes clear that the second-order intermodulation products are always 6 dB higher than the second harmonic, and that the third-order intermodulation products are always 9.54 dB higher than the third harmonic, because:

$$20 \cdot \lg\left(\frac{1}{2}\right) = -6 \text{ dB} \quad 20 \cdot \lg\left(\frac{1/4}{3/4}\right) = -9.54$$

The third-order intermodulation products are of particular importance. These components are very pronounced, and some of them arise close to the fundamental frequencies. In practice, it is difficult to filter out nearby interference signals. With wireless communications systems, for instance, it is possible that the adjacent channel might be affected. For this reason, the different wireless communications standards always include measurements that examine the extent of the TX stage's emissions in to the neighboring channel by measuring the adjacent channel leakage ratio (ACLR).

Fig. 30 shows the spectral components for a two-tone measurement (up to the third order) taken at an amplifier output.

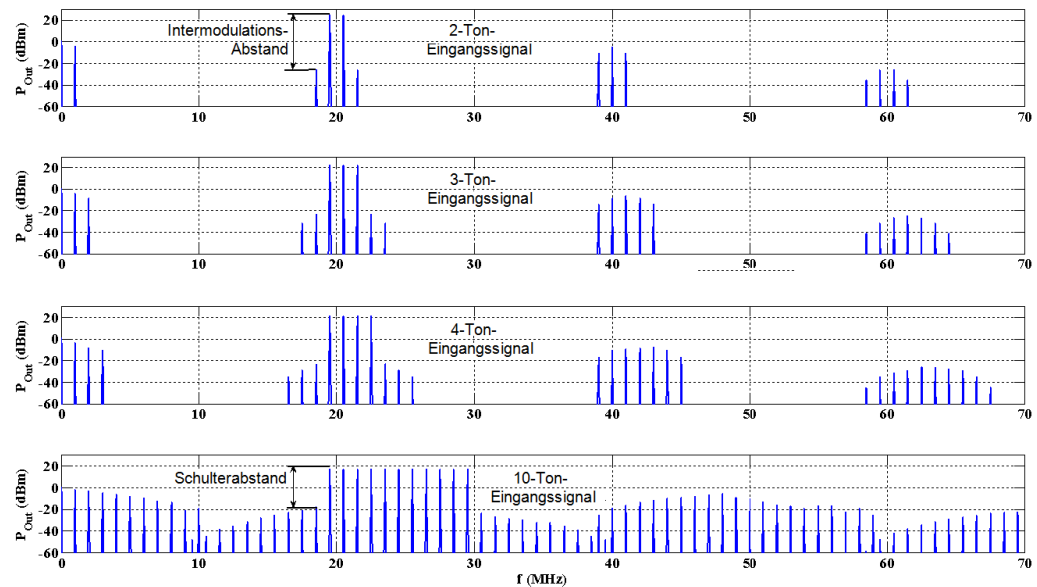


**Fig. 30: Result of a two-tone measurement with the level differences for harmonics and intermodulation products.**

The figure above makes it clear that two third-order intermodulation products arise near the operating frequencies. The difference between the level of the wanted signals and level of the intermodulation products is referred to as the intermodulation distance ( $d_{IMV}$ ). The difference in the level of the wanted signal and the level of the harmonics is called the harmonic distance  $d_{Hv}$  (and sometimes  $d_{kv}$ ).

The current relationship between the levels for the various spectral components (as shown in Fig. 30, for example) is only valid for the current input or output level.

Eq. (28) on page 42 was derived for two drive frequencies. The larger the number of frequencies applied, the more intermodulation products arise, see Fig. 31. Even when the nonlinearities are weak, a dense interference spectrum can arise. This is often observed, for example, in the case of cable TV systems that use many channels.



**Fig. 31: Output spectra for slight nonlinearity for 2, 3, 4 and 10 carriers [2].**

When the number of carriers rises (multicarrier systems), instead of intermodulation products, the "shoulder distance" is indicated in order to characterize the system's nonlinearity (Fig. 31).

With real multicarrier systems, the interference spectrum can take on noise-like characteristics.

### Discussion of the trace

The current relationship between the levels for the various spectral components (as shown in Fig. 30 for example) is only valid for the current input or output level. This section will discuss how the harmonics and the intermodulation products are a function of the drive level.

To simplify, we first set  $V_1 = V_2 = 1$ .

By ending after the third-order terms, Eq (28) ignores components; this begins as early as the 5th power on third-order frequencies. This is permissible as an approximation, because the absolute value of a spectral component decreases significantly as the order rises. For instance, with an average amplifier in its linear range, the third harmonic is usually already more than 30 dB below the fundamental, which means that coefficient  $a_3$  in Eq. (28) on page 42 is much smaller than  $a_1$ .

For this reason, for the fundamental, the term  $\frac{3}{4} \cdot a_3 \cdot U_1^3 + \frac{3}{2} \cdot a_3 \cdot U_1 U_2^2$  can be ignored compared to  $a_1 \cdot U_1$ . For the following general assessment of the power, the stimulus  $\sin(\omega_v \cdot t)$  is also left out. The goal here is not to determine the behavior in the time domain, but rather the relationships between the individual spectral components. Since the equation (28) is also valid for powers, one arrives at the following when both input signals have the same level:

$$P_{f0} = a_2 \cdot P^2 \quad \text{DC component} \quad (29a)$$

$$\begin{aligned} P_{f1} &= a_1 \cdot P && \text{Fundamental (first harmonic)} \\ P_{f2} &= a_1 \cdot P && \end{aligned} \quad (29b)$$

$$\begin{aligned} P_{2f1} &= \frac{1}{2} \cdot a_2 \cdot P^2 && \text{Second harmonic} \\ P_{2f2} &= \frac{1}{2} \cdot a_2 \cdot P^2 && \end{aligned} \quad (29c)$$

etc.

In a logarithmic representation, the following arises:

$$\lg P_{f0} = \text{const}_2 + 2 \cdot \lg P \quad \text{DC component} \quad (30a)$$

$$\begin{aligned} \lg P_{f1} &= \text{const}_1 + 1 \cdot \lg P && \text{Fundamental (first harmonic)} \\ \lg P_{f2} &= \text{const}_1 + 1 \cdot \lg P && \end{aligned} \quad (30b)$$

$$\begin{aligned} \lg P_{2f1} &= \text{const}_{21} + 2 \cdot \lg P && \text{Second harmonic} \\ \lg P_{2f2} &= \text{const}_{21} + 2 \cdot \lg P && \end{aligned} \quad (30c)$$

$$\begin{aligned} \lg P_{f2-f1} &= \text{const}_{22} + 2 \cdot \lg P && \text{Second-order} \\ \lg P_{f2+f1} &= \text{const}_{22} + 2 \cdot \lg P && \text{intermodulation products} \end{aligned} \quad (30d)$$

$$\begin{aligned} \lg P_{3f1} &= \text{const}_{31} + 3 \cdot \lg P && \text{Third harmonic} \\ \lg P_{3f2} &= \text{const}_{31} + 3 \cdot \lg P && \end{aligned} \quad (30e)$$

$$\begin{aligned} \lg P_{3f1-f2} &= \text{const}_{32} + 3 \cdot \lg P && \text{Third-order} \\ \lg P_{3f1+f2} &= \text{const}_{32} + 3 \cdot \lg P && \text{intermodulation products} \\ \lg P_{3f2-f1} &= \text{const}_{32} + 3 \cdot \lg P \\ \lg P_{3f2+f1} &= \text{const}_{32} + 3 \cdot \lg P && \end{aligned} \quad (30f)$$

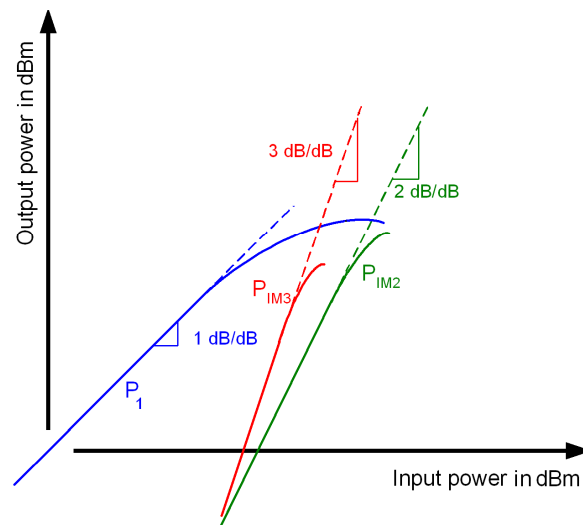
For operation **in the amplifier's linear range**, it follows from the discussion above that:

If the input is increased by 1 dB in each case, the fundamental's output power also increases by 1 dB.

If the input is increased by 1 dB in each case, the output power for the spectral components of the  $n$ -th order increases by  $n$  dB. This holds true for harmonics and for intermodulation products.

If, for example, an amplifier's input level is increased by 3 dB, the third-order intermodulation product grows by 9 dB.

Fig. 32 illustrates this relationship using the second and third-order intermodulation products as an example:



**Fig. 32: Characteristic curves for the fundamental (blue) and for the second-order (green) and third-order (red) intermodulation products.**

The curve for the second harmonic lies 6 dB below the green curve, and the curve for the third harmonic lies 9.54 dB below the red curve in Fig. 32.

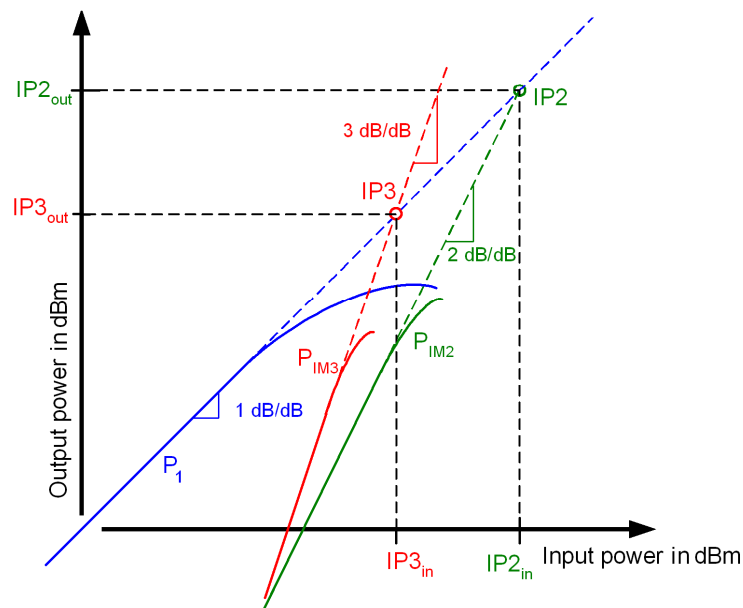
### 4.3 The Intercept Points IP2 and IP3

The harmonics and intermodulation products arising for a nonlinear two-port depend on the input level. For example, in order to compare an amplifier independently of the excitation, and in order to estimate the interference that is to be expected from a specific drive level, intercept points were introduced.



In the logarithmic representation of output power vs. input power, interfering spectral components (in the two-port's linear range) take the form of straight lines. The characteristic curves for components of the  $n$ -th order exhibit a slope of  $n$  dB per 1 dB of change in the input power.

If the straight characteristic curves – for example, for the second and third intermodulation products in the diagram – are extrapolated far beyond the possible operating range, these straight lines intersect with the extrapolated lines for the fundamental frequency, see Fig.



**Fig. 33: Determining the (fictitious) intercept points.**

The points at which these lines intersect are defined as the intercept points. Depending on which spectral components are being observed, there are different intercept points:

The intercept points for the second and third-order harmonics are referred to as the second-order harmonic intercept point (SHI) and third-order harmonic intercept point (THI).

The intercept points for the second and third-order intermodulation products are expressed simply as the second-order intercept point (SOI) and third-order intercept point (TOI).

In practice, it is above all the intercept point for the third-order intermodulation products (the TOI) that is of interest. This value is generally specified in product data sheets, for example.

These intermodulation products are particularly pronounced, and some of them are very close to the wanted frequencies. That makes them difficult to suppress with filters. Higher-order harmonics, on the other hand, generally have very low levels; they can usually be ignored.

For applications that use pure, unmodulated signals, the third-order harmonic intercept point (THI) is of interest. The THI is located 9.54 dB above the TOI.

Knowing the (fictitious) output power at the intercept points makes it possible to predict the levels that can be expected for the harmonics or for intermodulation products in the selected operating range:

For an output signal that is  $x$  dB below the  $n$ -th-order intermodulation products (IPs), the power  $P_n$  for the  $n$ -th component is:

$$P_n = \text{fictitious output power for IP}_n - n \cdot x \quad (31)$$

For example, an output signal that is 40 dB below a TOI of 35 dBm comes with a third-order intermodulation product of the power  $P_3$ :

$$P_3 = 35 \text{ dBm} - 3 \cdot 40 \text{ dB} = -85 \text{ dBm}$$

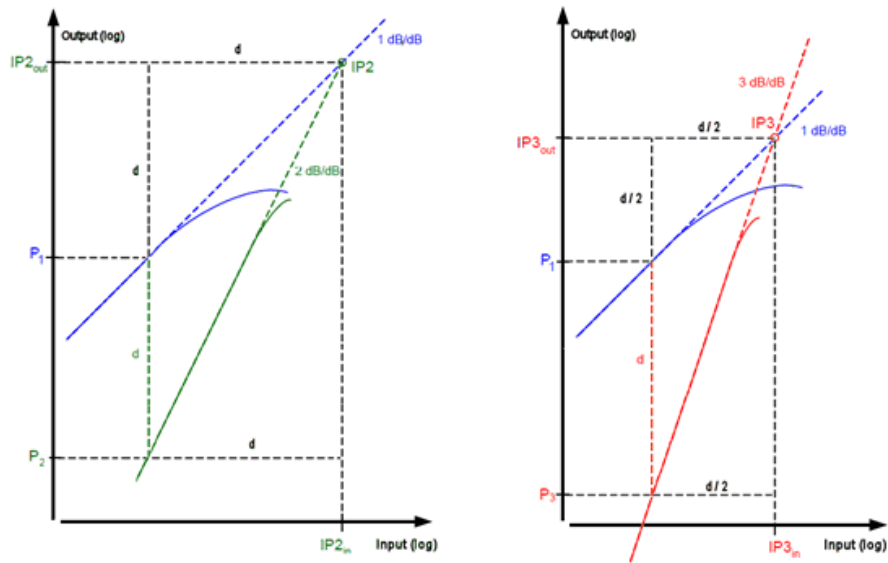
From a test and measurement perspective, there are two ways to determine the intercept points:

Measure the spectral components with a sinusoidal input signal, or  
Use the two-tone method (see section 4.2)

Measuring harmonics with a pure sinusoidal input signal requires a high dynamic range. This is done by performing multiple series of measurements to determine the characteristics and then plotting them on a graph and extrapolating the curves by extending them with straight lines that have the corresponding slopes. The power at the intercept points of the fundamental line and the corresponding interference lines can then be read from the graph, or calculated using the formula below. Performing test series to measure the data only makes sense when the intention is to work with pure sinusoidal signals.

The dynamic-range requirements are less stringent for the two-tone measurement method, because the intermodulation products are higher than the harmonics. That makes the measurement more reliable. Since they are close together, it is possible to capture the important IM3 products and the fundamentals together in one span. The intercept point can then be determined with a single two-tone measurement and a simple calculation.

The calculation exploits the fact that the  $n$ -th order characteristic curve rises by  $n$  dB per 1 dB. In such a case, it is possible to imagine the extrapolation of the straight lines as a diagonal inside a rectangle with an aspect ratio of 2:1 for second-order components or 3:1 for third-order components:



**Fig. 34: Geometric considerations for calculating the intercept points.**

The second-order intercept point can be obtained by taking the difference  $d$  from the measured second-order line and adding it to the current amplifier output power,  $P_1$ . The third-order intercept point can be determined by taking half the difference from the measured third-order line and adding it to the current amplifier output power,  $P_1$ , and so forth.

$$IP_n = P_1 + \frac{d}{n-1} \quad (32)$$

Advanced spectrum analyzers support the two-tone method for the third-order intercept point: They analyze the spectrum and supply numeric values for the TOI.

## 5 Crest Factor and CCDF

This chapter covers RF signals that have a high crest factor

$$C_F = \frac{V_{\text{Peak}}}{V_{\text{RMS}}}$$

I.e. it covers signals with peaks high above the RMS value. Signals of that kind primarily arise when advanced digital modulation schemes such as nPSK, QAM, CDMA or OFDM<sup>1</sup> are employed. Signals with a high crest factor arise in cellular networks, in digital television, and in many broadband transmission systems. In the time domain, as in the allocated frequency range, these signals are similar to thermal noise at first glance.

An RF signal's crest factor can refer to the overall signal or only to the modulated envelope. The discussion below is based exclusively on the latter view (modulated envelope). Consequently, the crest factor is the same for generation in the baseband as it is for the operating frequency: The crest factor for an unmodulated RF signal is:  $C_F = 0$ . If the first perspective had been taken, its crest factor would have been:  $C_F = 3.01$  dB (sine-wave carrier).

It makes sense to indicate the crest factor in dB. That way, only one value is required to examine both the voltage and power levels.

The crest factor focuses the view on the signal peaks. This is important for configuring a system to have the proper amount of electrical strength.

In practice, however, the probability of signal peaks arising is low. The probability of what the level might be at a given point in time is determined using the complementary cumulative distribution function (CCDF). Advanced spectrum analyzers offer this measurement function. Fig. 35 (on page 53) shows a measurement of this type.

The yellow line shows the probability that certain levels will be exceeded. This line is typical: a relative flat beginning is followed by a rapidly increasing drop. This means that the larger the peaks are, the lower the probability is that they will arise. A theoretical maximum level can be calculated.

For reference purposes, the red line shows the CCDF for white noise. Unlike the CCDF for the 3GPP-FDD signal, in this case, there is no rapid decrease of that kind, nor is there a maximum level. Theoretically, over the course of an infinite acquisition time (AQT), an infinitely high level would arise at least once.

<sup>1</sup> n-Phase Shift Keying, Quadrature Modulation, Code Division Multiple Access, Orthogonal Frequency Division Multiplex

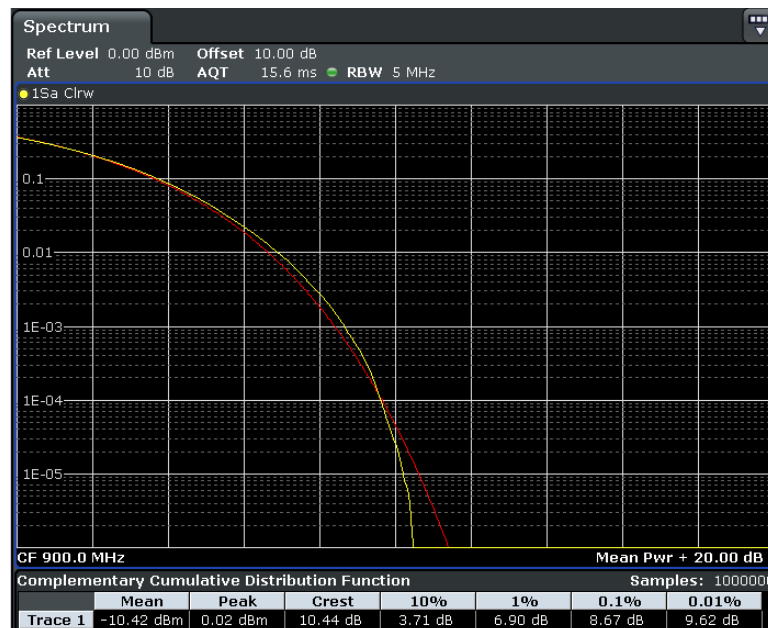


Fig. 35: CCDF for a UMTS signal, derived from 1000000 samples.

Significant, individual measurement results are indicated numerically on the bottom edge of the screen: For example, for 10 % of the observation period, the average value is exceeded by more than approximately 3.71 dB. This also means that for 90 % of the observation period, the signal remains below the level equal to approximately twice the RMS value (to be more precise: below the RMS value + 3.71 dB).

The CCDF derives frequency distribution for the level and for the RMS value from a large number of individual measurements. The longer the measurement period is, the more measurements are made and, as a result, more "rare" levels can be acquired. Consequently, the numerically indicated crest factor in the figure refers to the acquisition time of 15.6 ms selected here. With the R&S®FSV, this corresponds to a count of exactly  $10^6$  samples. From a statistical perspective, no events with a probability of  $< 10^{-6}$  will be acquired within this time frame. (To acquire components for which  $P < 10^{-6}$ ,  $10^{-7}$  to  $< 10^{-8}$  samples should be considered.)

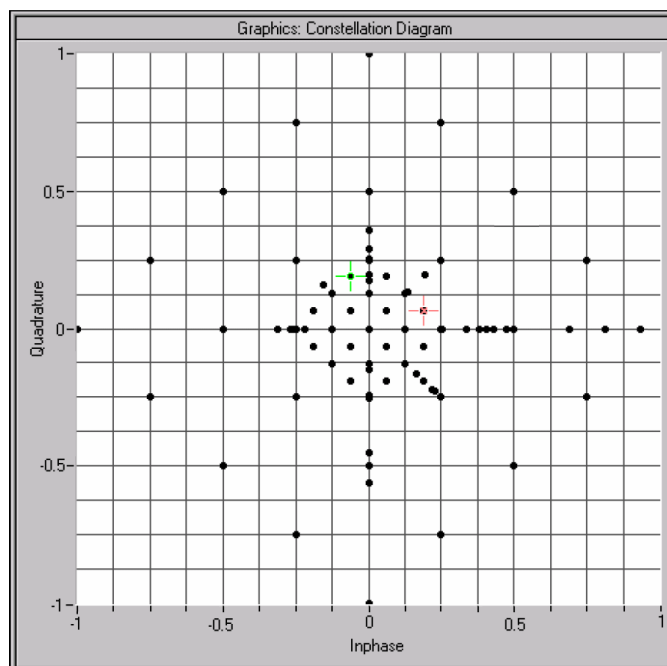
The signal peaks that can arise in a system directly influence the selection and dimensioning of the required components. Transmitter and receiver antennas, for instance, must be configured to achieve sufficient electrical strength. Voltage flashovers usually put electric components out of operation immediately. Passive elements, such as cables, are also susceptible to permanent damage from even just one overvoltage event.

When a drive signal's peaks extend into an amplifier's non-linear range, powerful, unwanted intermodulation products can arise both in the operating frequency range and at adjacent frequencies (spectral regrowth). For this reason, all wireless communications standards, for instance, require tests in order to keep these undesired products below the useful-channel power by a certain minimum amount, which is known as the minimal "adjacent channel leakage ratio" (ACLR).

In practice, in order to be able to use components that do not have to have the ability to withstand such high loads, and thus to lower costs, engineers reduce a signal's crest factor by cutting off the signal peaks, a process known as clipping. Theoretically, this does not influence the ACLR. Nevertheless, by its very nature, clipping lowers the signal quality and thus the transmission reliability. The error vector magnitude (EVM), a measure of modulation errors, rises, and the bit error rate increases. With digital transmissions systems, it is assumed, however, that rare bit or symbol errors that could arise in the clipped peaks can be tolerated and corrected by implementing effective methods for error detection. The percentage of the signal that can be clipped without exceeding a certain EVM must be examined carefully in each individual case.

When the peaks of a modulated signal are reduced to 50 % of the maximum value, for instance, the signal's crest factor does not decrease in equal measure by half (i.e. by 3 dB). This is due to the fact that, when the peaks are cut off, not only does the maximum value change; the average value has to be recalculated, too. The new ratio depends on the current signal statistics.

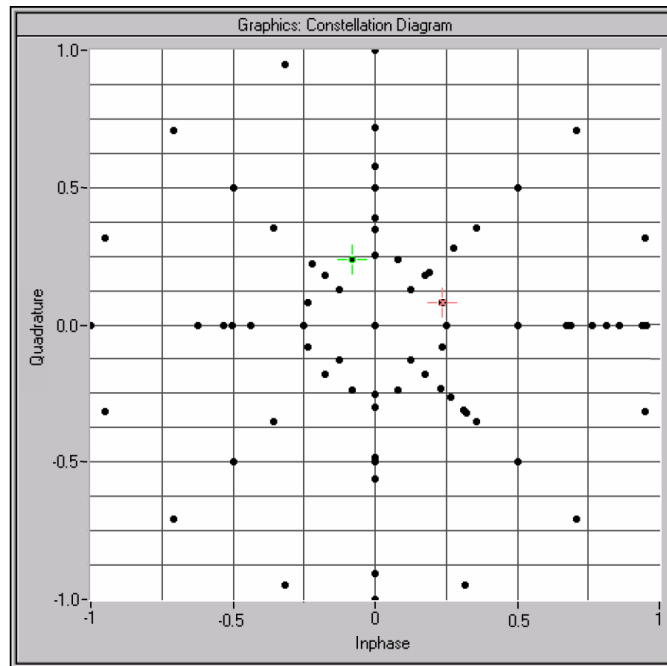
Modern digital modulation schemes always map a fixed number of bits into symbols and then transmit those symbols. A "constellation diagram" shows the modulation signal's phase and amplitude states for each symbol, see Fig. 36.



**Fig. 36: Sum constellation of a UMTS signal without clipping.**

In this example, the symbols of all the transmission channels in one UMTS signal are displayed. (Here, groups of channels are modulated in different ways.) The distance of each symbol point from the diagram's origin is proportional to the modulation signal's amplitude for this symbol; its angle determines the phase. The amplitudes are normalized to 1: All symbols are located within a circle or on the circle around the origin with radius of 1. Clipping to 50 % means that all points outside a new circle with the radius 0.5 in the constellation diagram are moved to this circle, preserving the angles.

As can be seen in Fig. 37, after clipping, it makes sense to normalize again in order to achieve the maximum word size, and thus return to the digital signal processing's highest dynamic range. All symbols are stretched "outward" to the unit circle. This increases the average value at the same peak amplitude as before, and the crest factor goes down. The automatic level control in a signal generator reduces amplification of the output stage in order to maintain the RMS value.



**Fig. 37:** Sum constellation for a UMTS signal after a 50 % clipping.

Theoretically, clipping does not influence the ACLR. Nonetheless, since the crest factor becomes lower, when the signal's RMS value remains the same, a lower peak modulation amplitude arises at an amplifier's input, for example. This can lower the ACLR at the output.

## 6 Phase Noise<sup>2</sup>

Phase noise describes a frequency's short-term stability, which makes it one of the signal source's important characteristics (along with other characteristics, such as the frequency range and long-term stability, power, spectral purity etc.).

An oscillator's phase noise is the reason why the oscillator not only appears as a line in the spectrum, but also appears continuously at frequencies below and above the target frequency – although the probability decreases sharply as the distance increases.

Phase noise, or rather the measurement of the phase noise of oscillators and synthesizers, is very important, particularly in wireless transmission systems: With receivers, the conversion oscillators' phase noise reduces sensitivity in adjacent channels when a strong input signal is present. In the case of transmitters, the transmitter oscillator's phase noise is, together with the modulator characteristics, responsible for the undesired power emitted in the adjacent channels.

Mathematically, the output signal  $u(t)$  of an ideal oscillator can be described as follows:

$$v(t) = V_0 \sin(2\pi f_0 t) \quad \text{Where}$$

$V_0$       Signal amplitude  
 $f_0$       Signal frequency and  
 $2\pi f_0 t$     Signal phase

With real signals, both the signal's amplitude and its phase are subject to variation:

$$v(t) = (V_0 + \mathcal{E}(t)) \sin(2\pi f_0 t + \Delta\varphi(t)) \quad \text{Where}$$

$\mathcal{E}(t)$     The signal's amplitude variation and  
 $\Delta\varphi(t)$     The signal's (phase variation or) phase noise

When working with the term  $\Delta\varphi(t)$ , it is necessary to differentiate between two types:

- Deterministic phase variation due, for instance, to AC humming or to insufficient suppression of other frequencies during signal processing. These fluctuations appear as discrete lines of interference.
- Random phase variation caused by thermal, shot or flicker noise in the active elements of oscillators.

<sup>2</sup> The information provided in this chapter has been taken largely from the application note "Phase Noise Measurement with Spectrum Analyzers of the FSE Family" by Josef Wolf [6].



One measure of phase noise is the noise power density with reference to 1 Hertz of bandwidth:

$$S_{\Delta\phi}(f) = \frac{\Delta\phi_{\text{rms}}^2 \text{ rad}^2}{1\text{Hz} \text{ Hz}}$$

In practice, single sideband (SSB) phase noise  $L$  is usually used to describe an oscillator's phase-noise characteristics.  $L$  is defined as the ratio of the noise power in one sideband (measured over a bandwidth of 1 Hz)  $P_{\text{SSB}}$  to the signal power  $P_{\text{Carrier}}$  at a frequency offset  $f_m$  from the carrier.

$$L(f_m) = \frac{P_{\text{SSB}}(1\text{Hz})}{P_{\text{Carrier}}}$$

If the modulation sidebands are very small due to noise, i.e. if phase deviation is much smaller than 1 rad, the SSB phase noise can be derived from the noise power density:

$$L(f) = \frac{1}{2} S_{\Delta\phi}(f)$$

The SSB phase noise is commonly specified on a logarithmic scale:

$$L_c(f_m) = 10 \lg(L(f_m)) \text{ dBc}$$

#### Measurement methods:

The simplest and fastest way to determine an oscillator's phase noise is to perform a direct measurement using a spectrum analyzer.

For this measurement, the following conditions must be met:

- The DUT's frequency drift must be small relative to the spectrum-analyzer sweep time. Otherwise, the signal-to-noise ratio will not be calculated correctly. The synthesizers commonly used in radiocommunications always fulfill this condition. The DUT can be locked to a reference of sufficient stability, or the DUT and analyzer can be synchronized directly.
- The spectrum analyzer's phase noise and AM noise must be low enough to ensure that the focus is on the DUT's phase noise and that it is not the spectrum analyzer's characteristics that are measured. The analyzer always provides the sum of the DUT's AM noise and phase noise and its own phase noise and AM noise.
- The DUT's amplitude noise must be significantly lower than the phase noise. This requirement is certainly met in the range close to the carrier frequency.

Another frequently used method employs a reference oscillator and a phase detector to perform the measurement [6]. Here, the fundamental frequency component within a certain bandwidth is suppressed. Thus, a very highly dynamic measurement can be made. Nevertheless, due to the significant amount of effort that this requires, every attempt will be made to determine the phase noise by performing a direct measurement with the aid of a spectrum analyzer. When necessary, calculations must compensate for the portion of the phase and AM noise that the spectrum analyzer itself contributes to the measurement results due to its inherent noise.

Specially designed spectrum analyzers have very low levels of phase and amplitude noise. They can determine a DUT's phase noise very precisely in a direct measurement. Furthermore, they often feature capabilities for performing measurements with a reference oscillator and a phase detector.

Many spectrum analyzers support direct measurement of the phase noise by providing a dedicated measurement option that simplifies the device settings and the evaluation of the measurement results. Fig. 38 shows a measurement of this kind taken with the aid of a feature known as a phase-noise marker.

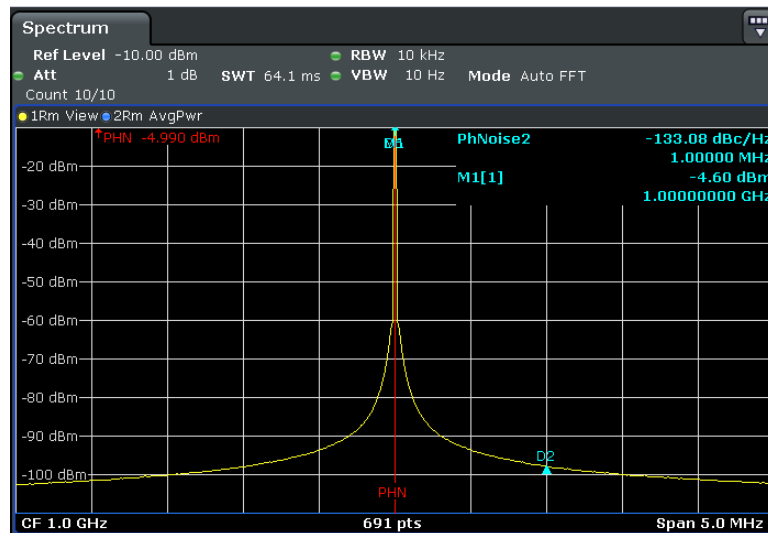


Fig. 38: Phase-noise measurement with the aid of a phase-noise marker.

The yellow line shows the smoothed measurement trace. It is important to keep in mind that this is always the sum of the DUT's phase-noise powers plus the component of the phase and AM noise that is due to the analyzer, even if that portion is small.

In this case, the SSB phase noise that is spaced 1 MHz from the carrier (Marker D2) is approximately  $-133$  dBc (1 Hz).

A measurement taken without an input signal (blue line in Fig. 38) supplies the value for the AM noise. If the analyzer's phase noise is also known, the DUT's phase noise can be determined with a high degree of accuracy from the overall sum that was measured.

## 7 Mixers<sup>3</sup>

Mixers are three-port components with two inputs and one output. An ideal mixer multiplies the two signals that are fed into its input ports. This makes it possible to convert signals to different frequencies. In the case of sinusoidal signals where

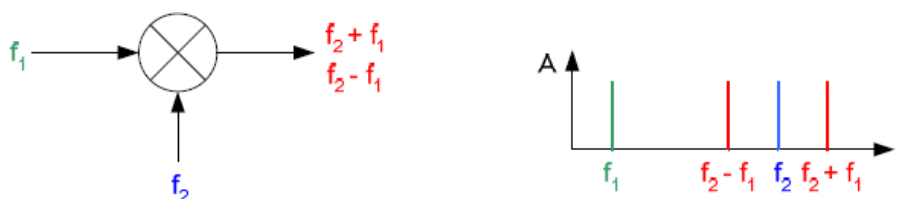
$$f_1 = \omega_1 / 2\pi \quad \text{and} \quad f_2 = \omega_2 / 2\pi ,$$

it is possible to transform the multiplication operation

$A(t) = A_1 \sin(\omega_1 t + \varphi_1) \cdot A_2 \sin(\omega_2 t + \varphi_2)$  as follows:

$$= \frac{A_1 A_2}{2} \{ \cos [(\omega_1 - \omega_2)t + (\varphi_1 - \varphi_2)] - \cos [(\omega_1 + \omega_2)t + (\varphi_1 + \varphi_2)] \}$$

This equation shows that the output signal from a mixer that multiplies in an ideal way consists of (exactly) two frequency components: one is the sum of the two input signals and the other is the difference between them.



**Fig. 39: Input and output signals of an ideal mixer.**

In standard nomenclature, the mixer ports are referred to as the radio frequency (RF) port, the local oscillator (LO) port and the intermediate frequency (IF) port. If the mixer operates as an up-converter (meaning that its output frequency is higher than its input frequency, as shown in Fig. 39, for example), the IF port serves as the input and the RF port the output. Conversely, with a down-converter (when the output frequency is lower than the input frequency), the RF port serves as the input and the IF port as the output. In both cases, a constant signal is applied at the LO at a fixed power level and at the suitable frequency.

A spectrum at the input (e.g. with a modulated signal) appears – once in the normal position and once in the inverted position – in the output signal spectrum as the upper and lower sideband. Generally, only one sideband is used, and the other is filtered out. The characteristics of a real-world mixer differ from the theoretical ideal. The primary characteristic variables for a real mixer are:

<sup>3</sup> The explanations provided in this chapter have been taken primarily from the "Up-converting Modulated Signals to Microwave with an External Mixer and the R&S@SMF100A" application note by C. Tröster, F. Thümmel and T. Röder [7].

- Conversion loss
- Isolation
- Harmonics and intermodulation products
- Linearity, 1 dB compression
- Impedance and VSWR

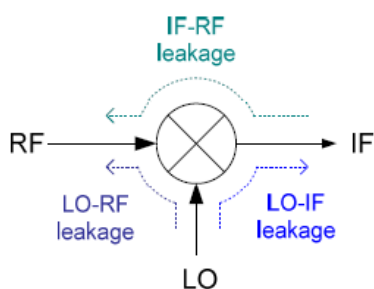
The relative importance of these individual characteristics varies depending on the application at hand. For example: In the case of an up-converter employed in a transmission system, the harmonics and the intermodulation products determine the quality of the overall system.

### Conversion Loss

Conversion loss is a measure of how efficiently a mixer transports the input signal's energy to the output. It is defined as the ratio between the input power and output power. Conversion loss depends on the frequencies that are used, on the signal level itself, and on the power level at the LO. In particular, the broadband signals used in advanced communications technology require a flat curve for the frequency response.

### Isolation

Isolation is a measure of the signal leakage or "crosstalk" between mixer ports.



**Fig. 40: Signal leakage in a mixer.**

Generally, the level at the LO is very high compared, for instance, to the level at the signal input. For this reason, even with good mixers, the LO frequency is usually clearly visible in the output signal spectrum. With up-converters in particular, LO-RF isolation is the most important parameter; the smaller the LO component in the output signal, the higher the quality.

### Harmonics and Mixing Products

An ideal mixer produces exactly two frequencies

$$f = |f_{LO} \pm f_{in}|.$$

A real mixer, on the other hand, produces mixing products and harmonics in accordance with the formula

$$f_{\mu,\nu} = |\mu \cdot f_{LO} + \nu \cdot F_{in}|$$

where  $\mu$  and  $\nu$  are integers (...,-2, -1, 0, 1, 2...). For example, the lower sideband arises with the values  $\mu = 1$  and  $\nu = -1$ , and the upper sideband arises with the values  $\mu = \nu = 1$ . The individual components differ inherently in amplitude. The lower and upper sidebands are the strongest mixer products; all other mixing products have lower amplitudes. Nevertheless, those other products can still result in a relatively large power level in the output spectrum. That is particularly the case for the LO harmonics.

The higher  $f_{LO}$  is, the higher the frequencies of the LO harmonics are. Since the system is band-limited at the mixer output, these harmonics do not appear in their full strength.

### Linearity

Just as is the case with amplifiers, with mixers, the power level at the output only remains proportional to the level at the input within a certain range. At a certain input power, the output signal begins to reach saturation. The 1 dB compression point is defined as the input power at which the output power sinks to 1 dB below the ideal linear characteristic curve.

The difference in amplitude between the noise floor and the 1 dB compressions point is referred to as the (linear) dynamic range.

### Intermodulation Products

Like amplifiers, mixers do not offer ideal linearity, even when they are driven within the range below the 1 dB compression point. For this reason, when two or more signals are applied simultaneously at the input, a whole series of intermodulation products – most of which are interference – arise at the output (in addition to the harmonics and mixing products with the LO signal).

When there are two signals with the frequencies  $f_1$  and  $f_2$  at the input of a component that is not ideally linear, the output contains spectral components at the frequencies

$$f_{n,m} = |n \cdot f_1 + m \cdot f_2|$$

where  $m$  and  $n$  are integers (...,-2, -1, 0, 1, 2...). The most problematic type are the third-order intermodulation products at the frequencies  $2 \cdot f_1 - f_2$  and  $2 \cdot f_2 - f_1$ , because they are close to the fundamental frequency.

The same principles and measurement procedures apply here as for amplifiers.

## 8 Introduction to the Measurements

This chapter begins the practical portion of this educational note. In the sections that follow, you will find common test and measurement tasks that are used every day for generating, processing and propagating communications signals.

For this educational note, the experiments were performed on Rohde & Schwarz® measuring instruments as examples. All of the user interfaces, measurement results and screen shots in the sections that follow are from these instruments:

[R&S® SMBV Vector Signal Generator](#) and  
[R&S® FSV Signal Analyzer](#)

The data sheets and manuals for these instruments (and others) can be downloaded free of charge from the Rohde & Schwarz® website: [www2.rohde-schwarz.com](http://www2.rohde-schwarz.com).

The [MPN2-00200200-27P](#) broadband amplifier from the company MITEQ® was used as the device under test (DUT). The [VLF-1000](#) filter from Mini-Circuits® was used as the lowpass filter. The attenuators are from the company Suhner®, and the combiner from the company Weinschel®.

Nevertheless, you do not have to use the above mentioned instruments and components to perform the measurement tasks described below. All of the experiments are described in such a way that you can also perform them using similar test and measurement hardware from other vendors.

For these and other reasons, the individual steps described in the experiment descriptions are not "written in stone." Instead, in some cases, you will first have to find the suitable settings (e.g. drive level, the measurement range, filter bandwidths, etc.). That is part of the exercise.

The purpose of these measurements is above all to enable you to become very familiar with the spectrum analyzer as an instrument. For this reason, while you are taking the measurements, make it a point to vary individual settings and observe the effects. This is the case, for instance, for selecting the detectors, the resolution bandwidth, the number of sweep points, etc.

## 9 Measuring the 3 dB Frequency Response

Determining or verifying the frequency response of a two-port device is one of the test and measurement tasks performed most frequently on AF and RF equipment.

The most commonly used technique for doing this is to feed a sinusoidal input signal into the device under test (DUT) and then measure the output signal. In the process, the test proceeds incrementally through the entire frequency range of interest.

This essentially "clamps" the DUT into a pair of tongs, so to speak. In the simplest case, this "tong tester" consists of two standalone instruments: a generator and an analyzer.

If the spectrum analyzer is equipped with a tracking generator, a single instrument is sufficient; in this case, the instrument's firmware usually supports this type of measurement. To an even greater extent, the same holds true for network analyzers.

When a generator and a spectrum analyzer are used, you measure the magnitude of the frequency response; with a network analyzer, it is possible to measure the magnitude and also the phase.

Amplifiers are generally two-port devices with a bandpass characteristic. They are optimized to achieve an almost constant frequency-independent gain in the passband. Outside the passband, the gain drops toward lower and higher frequencies.

**Definition:** An amplifier's bandwidth (operating bandwidth) is defined as the area between the two "cutoff frequencies," which are the lower frequency and upper frequency at which the gain falls by 3 dB.

Ideally, an amplifier's bandwidth should be independent of the instantaneous amplitude. In practice, this is not always the case. This is primarily due to the fact that the thermal situation changes along with the instantaneous amplitude.

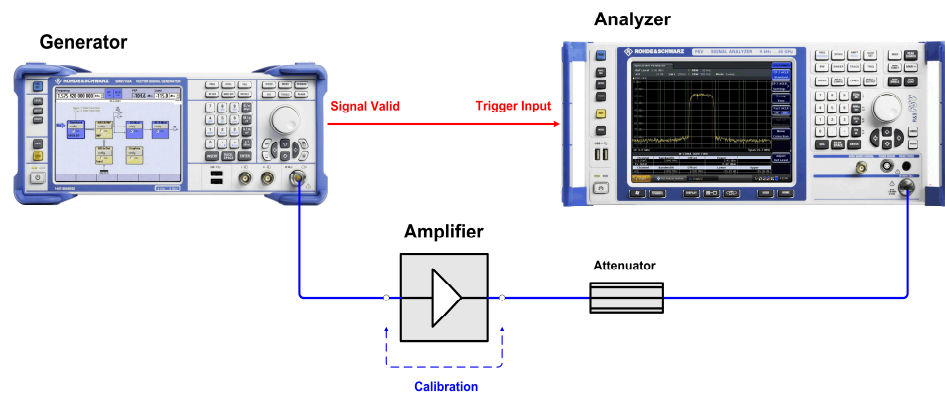
The following task is to measure an amplifier's frequency response at two different drive levels. This is first done with manual operation. After that, the measurements are repeated with the program controlling the measurement.

### 9.1 Test Setup

Fig. 41 shows the test setup for determining an RF amplifier's 3 dB bandwidth.

The test setup consists of a signal generator and a spectrum analyzer. The amplifier is the device under test.

For this experiment, the generator feeds the amplifier with an unmodulated RF drive signal with a constant level but with a frequency that changes in steps. Whenever the generator output settles, the instrument provides a trigger signal for the spectrum analyzer to measure and analyze the output of the DUT.



**Fig. 41: Test setup for determining an RF amplifier's 3 dB bandwidth**

To prevent the analyzer from being overloaded or even damaged, an attenuator is inserted into the measurement path.

To calibrate the test setup (described further below), replace the amplifier with a direct connection.

(The amplifier requires a power supply. That device is not shown in Fig. 41.)

- ▶ Never operate the amplifier with an open input or open output. The input is terminated when you connect the amplifier with the generator, even if the generator is not turned on. The components that are connected to the output must be able to cope with the amplifier's power.

#### Configuring the attenuator:

All analyzer families have their own specified maximum input power. This power must never be exceeded. For this reason, RF power should always be attenuated to a large extent outside the analyzer. In addition, high attenuation between the DUT and analyzer provides a certain level of protection against spikes, which can, for example, arise when an amplifier is switched on or off. Some of these peaks are significantly higher than the nominal power.

- ▶ Operate the analyzer with a nominal power of approx. 0 dBm at the input.

#### Example:

A measurement is to be taken at the amplifier's 1 dB compression point. The output power at the 1 dB compression point is 27 dBm for the DUT that is being used (as specified in its data sheet).

-> Attenuator: 30 dB for approx. -3 dBm at the input jack.

- ▶ Use an attenuator that can reliably cope with the amplifier's output power. Keep in mind that the output power beyond the 1 dB compression point can continue to rise: It just does not do this linearly!

#### Example:

Max. output power: 30 dBm -> Attenuator power rating > 1 W, preferably 2 W.



## Measurement Procedure

Put simply, the analyzer is able to "run through" a specific frequency range at sufficient resolution many times faster than a generator is able to do this. For this reason, to measure frequency response, you should have the generator increase the frequency step-by-step using the shortest possible dwell time.

To be exact, the dwell time that can be set on the generator does not determine the precise dwell time for a frequency, but rather only the rate at which the frequency changes. In reality, the actual dwell time at which the generator delivers a stable output signal is the specified dwell time minus the settling time.

Once the generator has settled, it indicates *Signal Valid* with a LOW level at the output. This signal triggers the analyzer, which then records the current DUT output with the MAX HOLD detector.

Consequently, in order to reliably acquire every frequency step, the analyzer's sweep time should be shorter than the generator's dwell time by at least a factor of two.

For a typical dwell time of 10 ms and a frequency increment of 1 MHz, the generator sweep requires 10 sec. per GHz.

- ▶ Begin by performing an orientation measurement.
- ▶ Allow for a relatively large frequency range on the generator and analyzer. Begin with a low level and measure the frequency response of the test setup, which has not yet been calibrated.
- ▶ Reset the generator and analyzer (*Reset*).
- ▶ On the generator, enter the *Start Frequency* and *Stop Frequency*, the frequency step size *Step Lin* and the *Dwell Time*:

State	On
Mode	Auto
Reset Sweep	
Start Freq	5.000 000 00 MHz
Stop Freq	3.000 000 000 00 GHz
Center Freq	1.260 000 000 00 GHz
Span	2.480 000 000 00 GHz
Current Freq	450.000 000 00 MHz
Spacing	Linear
Shape	Sawtooth
Step Lin	1.000 000 00 MHz
Dwell Time	10.0 ms
Global Trigger/Clock Settings...	
User Marker / AUX I/O Settings...	

Fig. 42: Generator settings for the frequency sweep.

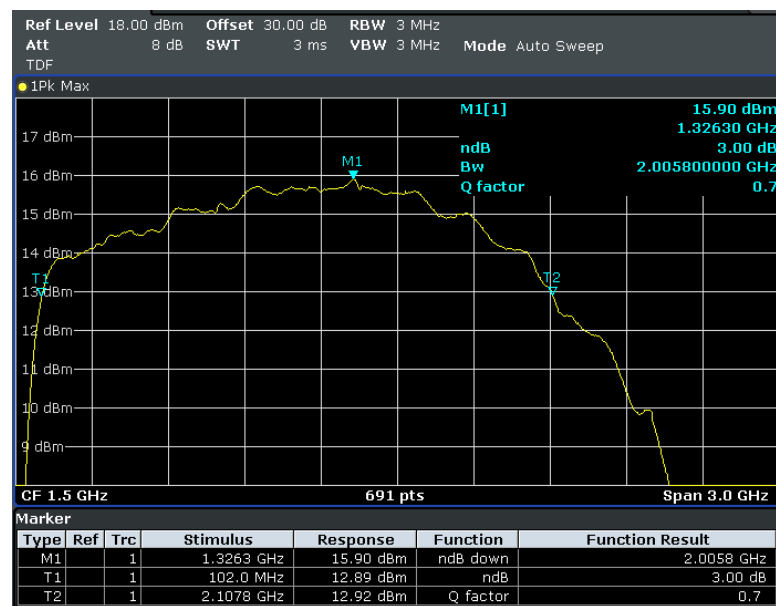
- ▶ At first, work approx. 10 dB below full-scale amplitude (10 dB below the 1 dB compression point).

The 1 dB compression point for the example DUT is at 27 dBm (data sheet). The gain is 24 dBm to 27 dBm; therefore, the generator level is -10 dBm.

- ▶ Compensate for the attenuator in the test setup with a corresponding *Ref. Level Offset* on the analyzer.
- ▶ Set the detector to *Max. Hold*.
- ▶ Set the trigger to *Ext, neg. Polarity. Trigger Offset = 0*

After a complete sweep:

- ▶ Set a *Marker to Peak* and use the *n dB down* marker function to start an initial 3 dB frequency response evaluation.



**Fig. 43: Orientation measurement on the DUT.**

Although this measurement was performed on an uncalibrated setup, it still delivers fairly accurate 3 dB frequencies and information on the level that is to be expected. Nevertheless, the purpose of this measurement was to limit the frequency range that is to be calibrated and then later measured (approx. 20 MHz to 2.5 GHz).

- ▶ Note the evaluated 3 dB frequencies.

## 9.2 Calibration

The calibration starts with a reference measurement performed without a DUT. The test setup's frequency response is used to determine the compensation values used for the measurements to be taken later with a DUT.

In this case, it can be assumed that the frequency response for cables and attenuators is independent of the level. When high-end components are used and the measurement is only to cover a small frequency range, compensating the path requires only a few corrections. In some cases, however, compensating the path involves a great effort.

In individual cases, establishing the correct path compensation might be the real challenge when it comes to measuring the frequency response. There are two basic methods for path compensation:

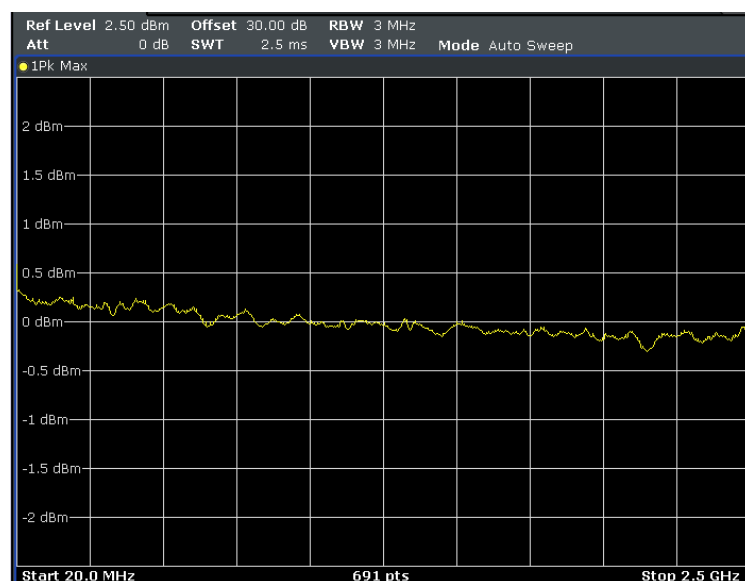
Internal or external software for the device stores the results of the reference measurement and corrects the DUT's trace. Network analyzers and analyzers with a tracking generator offer special device functions to support this.

Correction values in the analyzer transducer tables weight the trace data in the stand-alone analyzer. This method is described in the section below.

### Entering correction values into a transducer table

- ▶ Disconnect the amplifier power supply.
- ▶ Replace the amplifier's RF path with a direct connection.
- ▶ Measure the measuring path's insertion loss.

To do this, use start and stop frequencies that are somewhat below or above the 3 dB frequencies that you acquired during the orientation measurement. Using a generator level of 0 dBm is advisable. Fig. 44 shows the results for a measurement of this type:



**Fig. 44:** Frequency response for the measurement path.

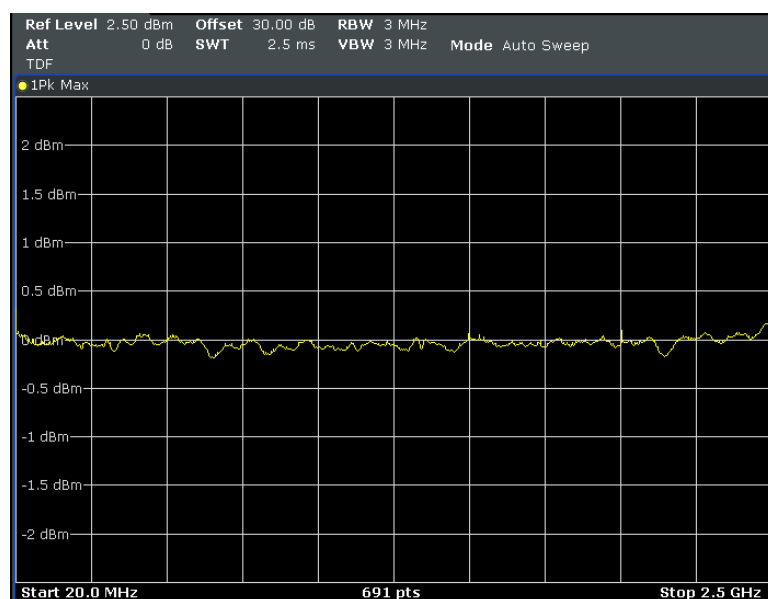
Here the measurement error caused by the test setup's frequency response is definitely lower than 0.5 dB. However, even this frequency response can still be improved:

When, as in Fig. 44, it becomes clear that the level tends to fall or rise monotonously, it is easy to compensate for this behavior by making a few transducer entries:

In this case, a compensation line was defined by two table entries:

–0.25 dB at 20 MHz      and      0.2 dB at 2.5 GHz.

Fig. 45 shows the measurement result after activating the compensation; the error is generally less than 0.2 dB:



**Fig. 45: Frequency response for the measuring path with compensation.**

The analyzer being used here can make use of eight transducer tables with 625 data points each; as a result, it is able to handle complicated traces, too.

- ▶ Define simple transducer correction values and observe the results they produce! (Don't forget to activate each table after editing it.)

## 9.3 Measuring

- ▶ Set the signal generator to the following drive level:

$$P_{\text{Generator}} = \text{Amplifier's output power at the 1 dB compression point} \\ - \text{nominal gain} - 10 \text{ dB}$$

- ▶ Insert the amplifier into the measuring path as shown in Fig. 41 on page 64.

- ▶ Connect the amplifier's power supply.
- ▶ Take a measurement as done with the orientation measurement – but with the new start/stop frequencies – and note the 3 dB bandwidth.



Fig. 46: Frequency response at an instantaneous amplitude for maximum power – 10 dB.

**Note:**

You can store the current trace on the screen by switching the trace detector that is being used to *View*.

You can put an additional measurement on the screen with another trace.

You can delete the current trace by briefly switching the detector to *Clear Write*.

- ▶ Repeat the frequency response measurement at full-scale input and note the 3 dB bandwidth.

(To do this, you must increase the generator level by approx. 10 dB.)

**Questions:**

Does the 3 dB bandwidth depend on the instantaneous amplitude?

What happens when you set the RBW to 1 MHz?

What happens when you use the RMS detector?

What does the Q-factor mean?

- ▶ Summarize the goals, execution and results of this experiment in your own words.

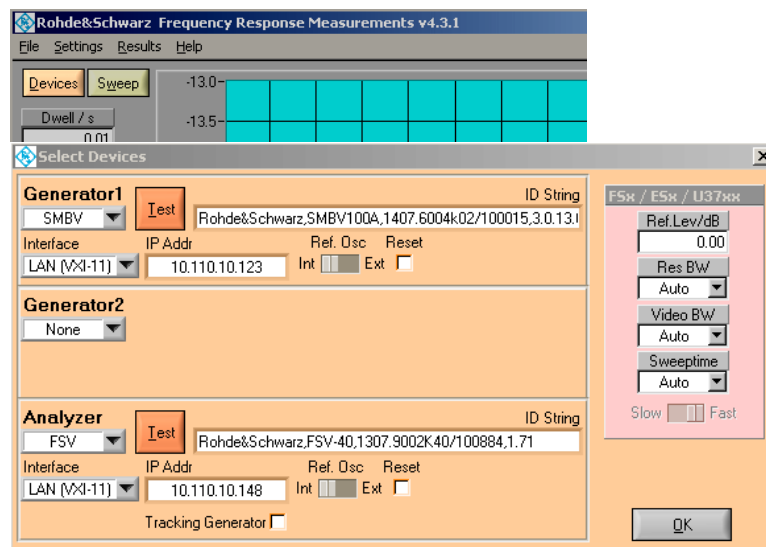
## Measurement software

The following task is only to be performed if you are working with Rohde & Schwarz instruments!

To support frequency response measurements performed with its generators and analyzers, Rohde & Schwarz makes the 1MA09 *FreRes* (Program for Frequency Response Measurements) application note available. This application note is available as a free download at: <http://www.rohde-schwarz.com/appnote/1MA09>.

*FreRes* controls frequency response measurements remotely. The instruments are addressed via GPIB or LAN.

- ▶ Download *FreRes* and install it on your test computer. Operating this software also requires a VISA library; see <http://www.ni.com/VISA>.
- ▶ Hook up the remote-control cabling.
- ▶ Start the program.
- ▶ First specify these *Devices*: *Generator1* and *Analyzer*.

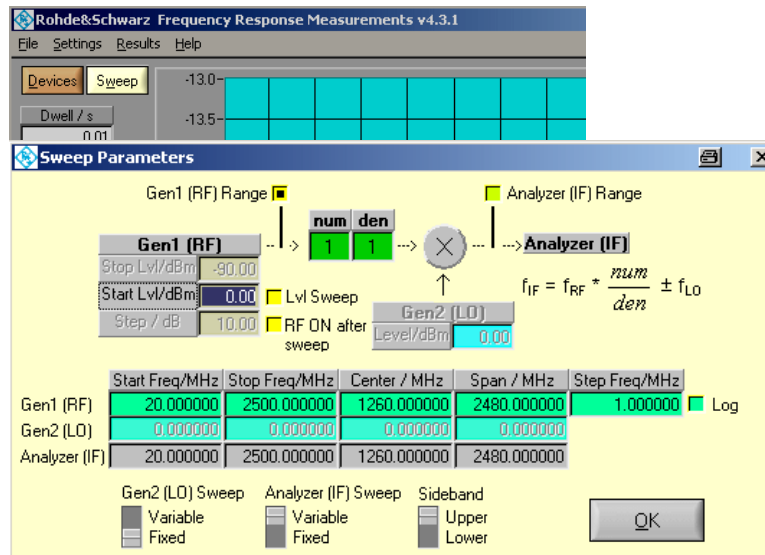


**Fig. 47: Entering the remote-control parameters.**

*FreRes* can also test frequency-converting modules. In such cases, a second generator is required. When testing amplifiers only, Generator2 is not required.

- ▶ Use the *Reset* checkbox to establish whether the corresponding instrument should be reset prior to a measurement. (The *Frequency Response Measurements* program only varies the frequency and the generator level. If *Reset* is not active, some of the additional settings that you have made, such as a *Ref Level Offset*, remain intact.)
- ▶ Click OK.

- ▶ Then specify the Sweep settings; see Fig. 48.



**Fig. 48: Entering the measurement parameters.**

When frequency shifting modules are used, use the checkboxes *Gen1 (RF) Range* or *Analyzer (IF) Range* to establish whether the input frequencies or the output frequencies should be used for the axis labeling. When performing a pure amplifier measurement, both are the same; any setting is correct.

- ▶ For calibration, specify a *Start Lvl* of 0 dBm.
- ▶ Ensure that the *Lvl Sweep* is not activated.
- ▶ For *Gen1 (RF)*, specify the *Start Lvl*, the *Start Frequency* and *Stop Frequency* and the *Frequency Step*.

The *Gen2 (LO)* field cannot be edited because a second generator has not been activated.

The entries in the *Analyzer (IF)* field are similar to the generator settings.

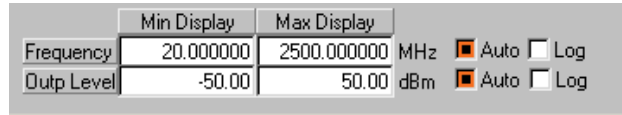
The toggle switches for *Gen2 (LO) Sweep* and *Sideband* are only significant for frequency shifting; they are not relevant for amplifier measurements (the toggle positions are unimportant).

The sweep for Generator 1 is always active.

- ▶ Ensure that *Analyzer (IF) Sweep* is also active (with the switch toggled to *Variable*).
- ▶ Click OK.

Now you are back at the main menu.

- There, first activate the auto-scale functions for the output window:



**Fig. 49: Auto-scaling for both axes shows the overall curve.**

Now perform a calibration measurement (without an amplifier) as described in section 9.2.

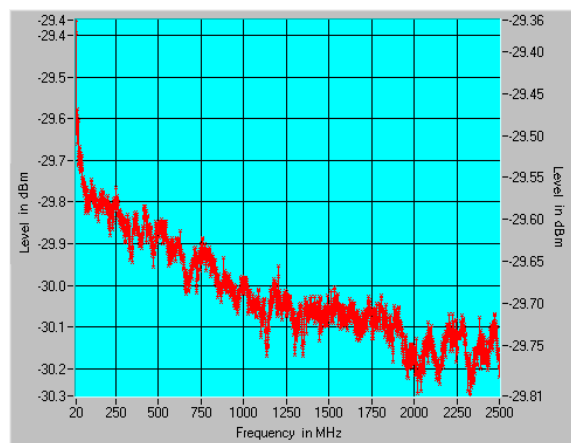
- To do this, click *Start*.



**Fig. 50: Buttons for controlling the recording.**

The *Start* button deletes the screen content and starts a new measurement (without any frequency response corrections).

Fig. 51 shows a typical result:

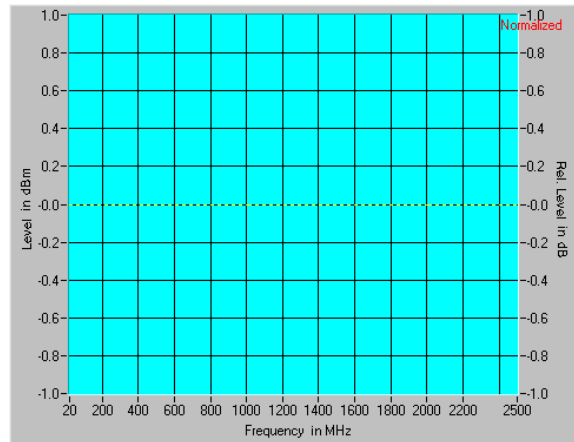


**Fig. 51: Frequency response for the measuring path.**

The insertion loss is approximately 30 dB; the level on the analyzer falls slightly as the frequency rises.



- Now click *Normalize*. You receive the following result:



**Fig. 52: Frequency response for the normalized measuring path.**

The top right of Fig. 52 shows that normalization was performed: The analyzer values were compensated so that they result in 0 dBm at every frequency (this is the generator's *Start Lvl*, which you entered in Fig. 48).

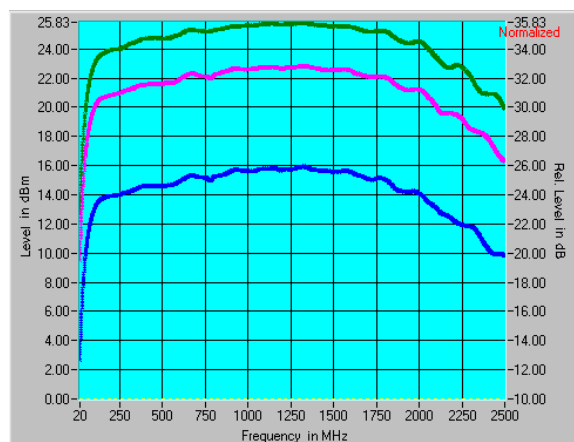
Now perform the measurement with the DUT:

- Put the amplifier back into the test setup, and switch its power supply on.
- In the *FreRes* program, click *Repeat* (not *Start*).

This starts a new measurement. The analyzer results are now corrected based on the calibration measurement.

You can stop a running measurement by clicking *Stop Meas*, and you can delete the most recent trace with *Del Last Trace*.

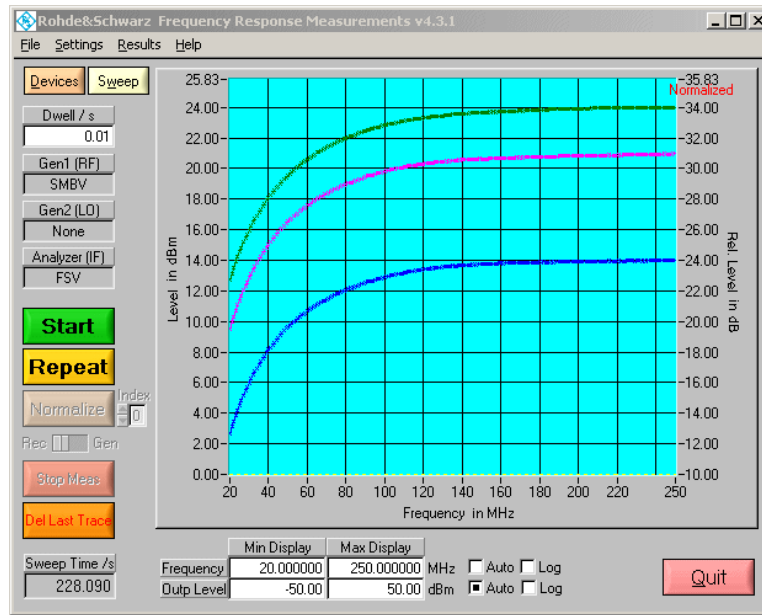
By clicking *Repeat*, you can start additional sweeps (for example with different *Start Lvl* settings); the results are corrected again. Fig. 53 shows traces for full-scale input (green), -3 dB (pink) and -10 dB (blue).



**Fig. 53: Frequency response for three instantaneous amplitudes.**

The **Start** button **deletes the normalization**, deletes the screen content and starts a new measurement.

If you turn the automatic scaling off, you can zoom in on interesting section of a completed recording, such as the edges of the passband characteristic:

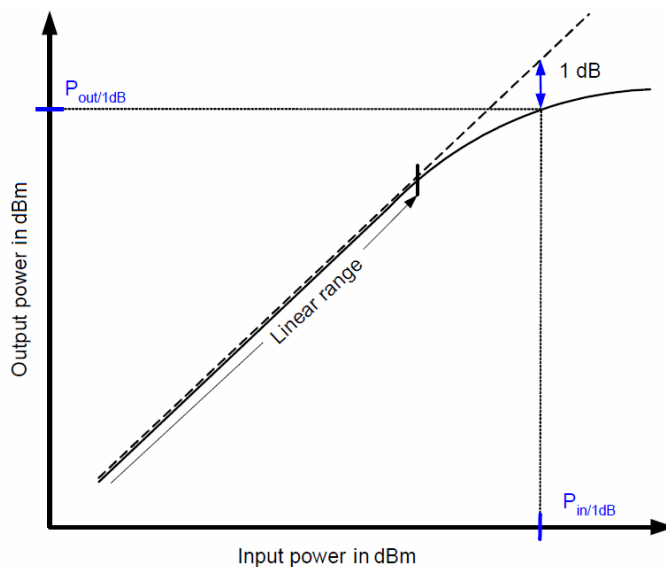


**Fig. 54:** Zoom-in display of the area near the starting frequency.

## 10 Determining the 1 dB Compression Point

The following explanations examine amplifier behavior when the frequency is constant and the input-signal level is variable.

With drive levels up to a certain limit, the relationship between the output and input powers remains constant. This range is referred to as the amplifier's linear range, and the ratio of the output to the input power is called the nominal gain. If the excitation is increased further, the amplifier reaches saturation, and the gain decreases. Fig. 55 illustrates this principle.



**Fig. 55: Transmission curve for an amplifier.**

Within the linear range, the amplifier's characteristic curve is a straight line, and the line's slope is a measure of the gain. A reduction in gain becomes visible as a flattening of the characteristic: The output power falls further and further behind the expected nominal output power. In general, the same holds true for all linear AF and RF amplifiers.

### Definition:

The "1 dB compression point" is that output power at which the output power has decreased by 1 dB compared to the nominal output power.

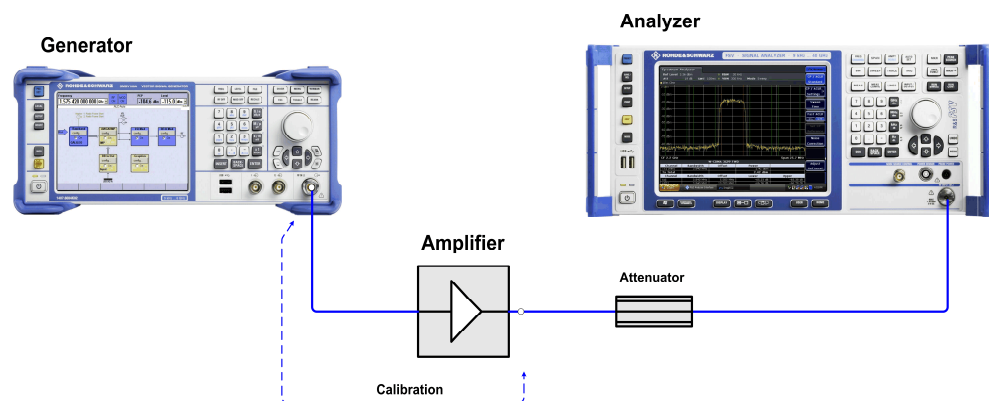
In Fig. 55, the 1 dB compression point is at  $P_{out/1dB}$ .

As the drive level continues to rise, the amplifier's output signal becomes increasingly distorted. This means that increasing numbers of harmonics arise. To keep these undesired components from falsifying the measurement results, the input signal used to drive the amplifier for this measurement is a sine signal, and the frequency is measured selectively at the amplifier output. This is best done with a spectrum analyzer.

It is also possible to perform these measurements with a broadband power meter; however, it is then necessary to suppress the harmonics using a lowpass filter at the amplifier output.

## 10.1 Test Setup

Fig. 56 shows the test setup for determining the 1 dB compression point.



**Fig. 56: Test setup for determining an RF amplifier's 1 dB compression point.**

The test setup consists of a signal generator and a spectrum analyzer. The amplifier is the device under test.

For this experiment, the generator supplies the amplifier with an unmodulated RF drive signal at defined levels. The spectrum analyzer measures and analyzes the amplifier's output signal.

To prevent the analyzer from being overloaded or even damaged, an attenuator is inserted into the measurement path.

To calibrate the test setup (described further below), connect the attenuator directly to the generator.

(The amplifier requires a power supply. That device is not shown in Fig. 56.)

- ▶ Never operate the amplifier with an open input or open output. The input is terminated when you connect the amplifier with the generator, even if the generator is still turned off. The components that are connected to the output must be able to cope with the amplifier's power.

### Configuring the attenuator:

All analyzer families have their own specified maximum input power. This power must never be exceeded.

For this reason, RF power should always be attenuated to a large extent outside the analyzer. In addition, high attenuation between the DUT and analyzer provides a certain level of protection against spikes, which can, for example, arise when an amplifier is switched on or off. These peaks are sometimes significantly above the nominal power.

- ▶ Operate the analyzer with a nominal power of approx. 0 dBm at the input.

**Example:**

The amplifier's output power at the 1 dB compression point: 27 dBm. The measurement should be taken somewhat beyond the 1 dB compression point.

-> Attenuator: 30 dB for approx. 0 dBm at the input jack.

- ▶ Use an attenuator that can reliably cope with the amplifier's output power. Keep in mind that the output power beyond the 1 dB compression point can continue to rise: It just does not do this linearly!

**Example:**

Max. output power: 30 dBm -> Attenuator power rating > 1 W, preferably 2 W.

## 10.2 Calibration

This experiment focuses on the flattening of the amplifier's **output** signal, i.e. the deviation from the linear extrapolation of the **output** signal. For this reason, compensation is only performed for the path "behind" the amplifier.

What influence does the path's insertion loss ahead of the amplifier have on the characteristic in Fig. 55?

- ▶ Disconnect the amplifier power supply.
- ▶ Replace the amplifier's RF path with a direct connection.
- ▶ Measure the measuring path's insertion loss.

**Procedure:**

Example data: The measurement should be taken at the frequency of 900 MHz. Your amplifier's 1 dB compression point is 27 dBm. The nominal gain is approx. 26 dB. A 30 dB attenuator has been added to the circuit ahead of the analyzer.

- ▶ Make the following settings on the analyzer:

*Reset* (return the instrument to the default settings)

<i>Center Frequency</i>	900 MHz
<i>Span</i>	50 MHz
<i>Reference Level</i>	-20 dBm

- ▶ Make the following settings on the generator:

*Reset* (to the default settings)

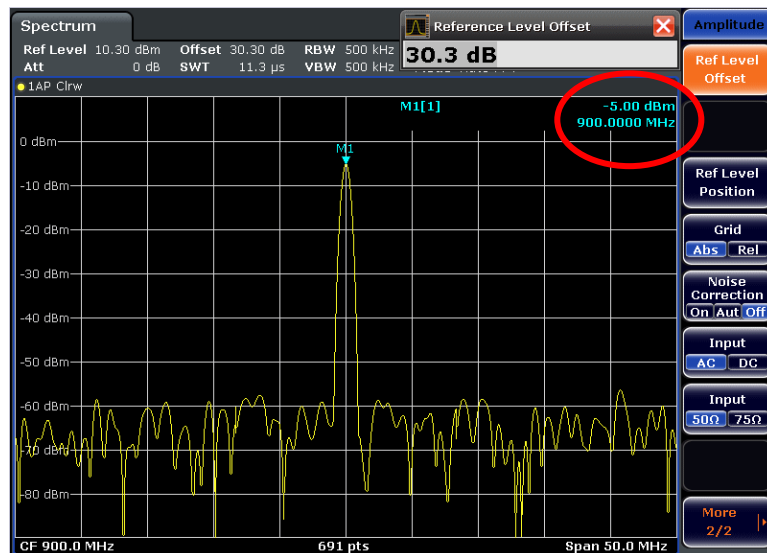
*Frequency* 900 MHz

*Level* -5 dBm

*RF* Off

- ▶ Measure the power of the generator signal on the analyzer. Use the *Marker to Peak* function for this.
- ▶ Determine the path attenuation by taking the difference between the power delivered by the generator and the power measured on the analyzer (for example: -5 dBm - (-35.3 dBm) = 30.3 dB).
- ▶ Compensate for the path attenuation: On the analyzer, enter the value that you have determined as the *Reference Level Offset*.

At the marker position, the analyzer now shows you the same level that is displayed on the generator: -5 dBm ± 0.1 dB.



**Fig. 57:** Measurement results with exactly the right level after calibration of the measuring path.

### To prepare the measurement

- ▶ On the generator, switch the *RF* to *Off*.
- ▶ Reduce the level on the generator to -20 dBm.
- ▶ Set the *Reference Level* on the analyzer to +20 dBm.

## 10.3 Measuring

- ▶ Insert the amplifier into the measuring path as shown in Fig. 56 on page 75.
- ▶ Connect the amplifier's power supply.
- ▶ Set the signal generator to the following drive level:

$$P_{\text{Generator}} = \text{Amplifier's output power at the 1 dB compression point} \\ - \text{nominal gain} - 10 \text{ dB}$$

- ▶ Measure the power on the analyzer (using the marker function). Enter your results in a table.
- ▶ Increase the generator power 1 dB increments; perform approx. 15 measurements.
- ▶ Plot your results graphically and determine the 1 dB compression point.

### Measurement results (example):

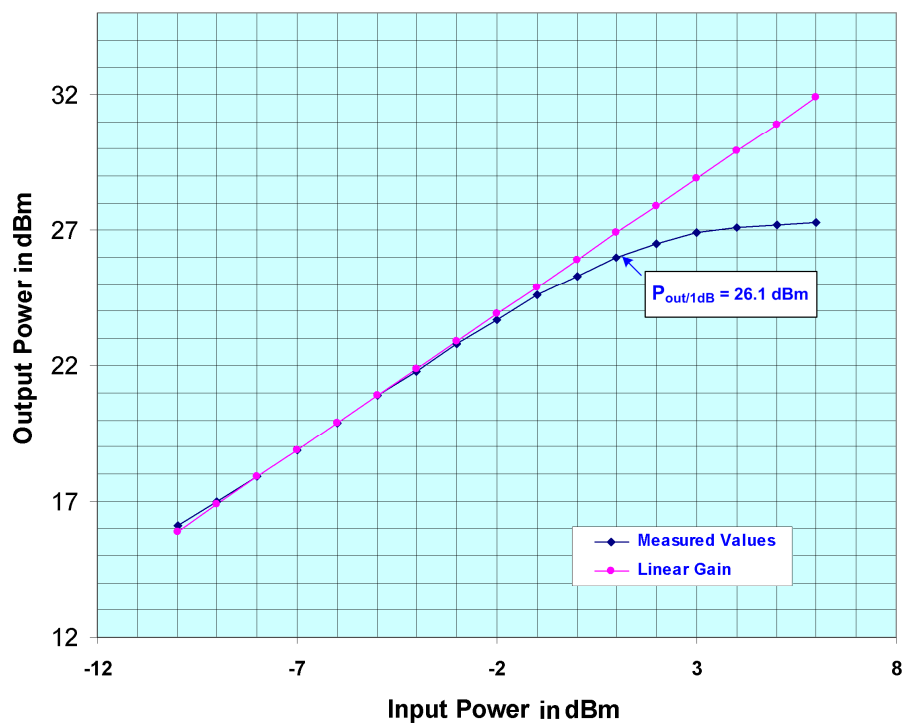
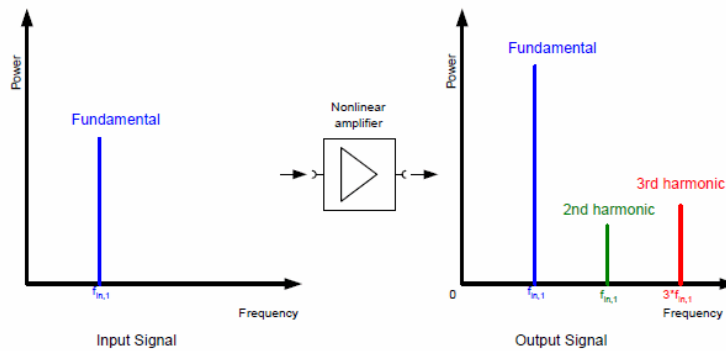


Fig. 58: Output power vs. input power, 1 dB compression point.

- ▶ Summarize the goals, execution and results of this experiment in your own words.

# 11 Measuring Harmonic Interference, IP2 and IP3

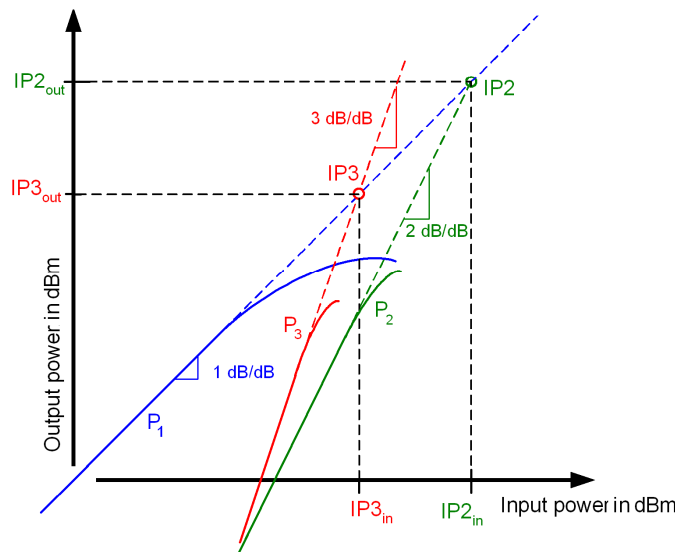
The following explanations examine how harmonics arise and grow at the output of an amplifier driven with an unmodulated signal that has almost no harmonics. The observations here will cover only the second and third harmonics, since they are the strongest and most important ones.



**Fig. 59: Harmonics of an amplifier being driven with a pure sine signal.**

Harmonics arise due to the nonlinearities of real components. As the input amplitude rises, the amplitudes of these harmonics grow in the output signal faster than the fundamental does. In general:

When the input signal of a two-port device (such as an amplifier) is raised in the linear range by 1 dB, the fundamental in the output signal also rises by 1 dB, and each n-th harmonic rises by n dB, see Fig. 60.



**Fig. 60: Basic characteristic curves for the first (fundamental, blue), second (green) and third (red) harmonic.**



With the logarithmic display that has been selected, the characteristic takes on the form of a straight line within the linear range. When these straight lines are then extended far beyond the permissible operating range, they intersect at the intercept points IP2 and IP3 (see Fig. 60).

Since these are harmonics, IP2 is called the second-order harmonic intercept point, and IP3 is called the third-order harmonic intercept point (THI).

(The intercept points for intermodulation products, which are covered later, are abbreviated as the SOI and TOI.) Further information is available, for instance, in the white paper "Interaction of Intermodulation Products between DUT and Spectrum Analyzer" [4].

The purpose of the following task is to measure the course of the first harmonic (fundamental), second harmonic and third harmonic at the output of an amplifier as a function of the excitation at the input. The measurements are only performed within the amplifier's linear range. After that, the IP2 and IP3 intercept points are to be established.

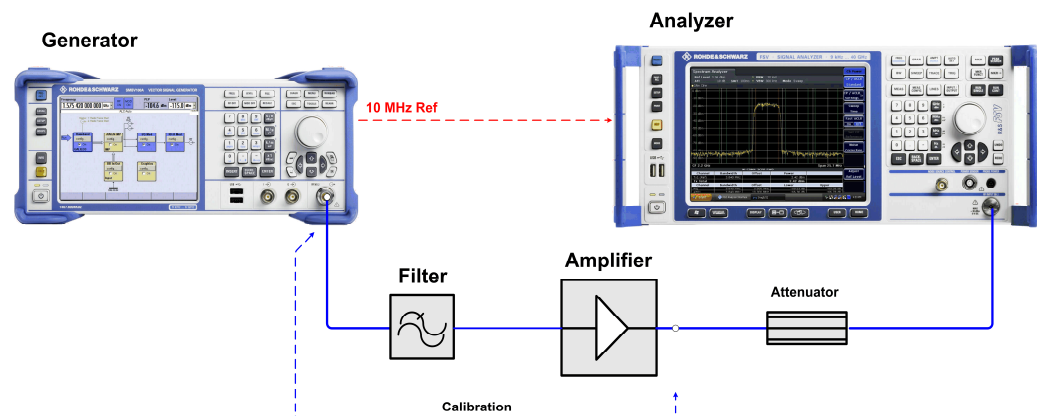
The input frequency is 900 MHz.

Measuring the harmonics is a demanding task, because a good amplifier only generates very low harmonics within its linear range. This means that there is a significant difference between the level of the fundamental and, above all, the level of the third harmonic; this requires a large dynamic range.

Consequently, the analyzer must be driven in a range in which it is ensured that the sensitivity is high and that its own generation of harmonics is minimized (ideal mixer level).

Additionally, it must also be ensured that the signal being fed into the amplifier is largely free of harmonics.

## 11.1 Test Setup



**Fig. 61: Hardware setup for measuring the harmonics.**

The test setup essentially consists of a signal generator and a spectrum analyzer. The amplifier is the device under test.

For this experiment, the generator feeds the amplifier with an unmodulated HF signal at 900 MHz. In order to sufficiently suppress the generator's harmonics, a lowpass filter with a cutoff frequency of 1 GHz is added to the circuit after the generator.

The spectrum analyzer measures and analyzes the amplifier's output signal. The attenuator between the amplifier and analyzer ensures that the analyzer operates within the optimal level range and simultaneously provides protection against overload and damage.

Since this measurement demands a high degree of accuracy, the generator receives a 10 MHz reference signal from the analyzer.

To calibrate the measuring path downstream from the amplifier (described further below), the lead at the amplifier output is connected directly to the generator.

- ▶ Never operate the amplifier with an open input or open output. The input is terminated when you connect the amplifier with the generator, even if the generator is still turned off. The components that are connected to the output must be configured for the amplifier's power.

#### Configuring the attenuator:

- ▶ Note your amplifier's output power at the 1 dB compression point. You will find this value in your amplifier's data sheet.

The measurement should be taken within the amplifier's linear range; in other words, up to a maximum of 3 dB below the 1 dB compression point.

- ▶ Refer to the analyzer manual to find the optimal mixer level for minimizing distortion.

The mixer level is the level at the analyzer input minus the attenuation that has been set on the instrument for the internal attenuator (the RF attenuation); see Fig. 20. To achieve the best possible ratio between the signal and the inherent noise, this level is set to 0 dB for this measurement.

Determine the attenuator value in the test setup shown in Fig. 61 so that, at a signal that is 3 dB below the 1 dB compression point, the mixer level is within the range that has the least amount of distortion.

#### Example:

The amplifier's output power at the 1 dB compression point: 27 dBm  
 The amplifier's linear range up to approximately: 27 dBm – 3 dB = 24 dBm  
 Optimal level on the analyzer's mixer: –22 dBm to –17 dBm, according to the data sheet

-> Optimal attenuator:  $24 \text{ dBm} - (-17 \text{ dBm}) = 41 \text{ dB}$

The value used is 40 dB.

- ▶ The attenuator must be able to reliably cope with the amplifier's output power.

In this example:  $27 \text{ dBm} \cong 0.5 \text{ W}$  -> Attenuator power rating, ideally 1 W.

## 11.2 Calibration

- ▶ Disconnect the amplifier power supply.
- ▶ Connect the cable that is to lead to the amplifier output directly to the generator. The measuring path consists of the cabling and attenuator(s). The attenuation of the path "ahead" of the amplifier is unimportant.
- ▶ Now measure the insertion loss at 900 MHz, 1.8 GHz and 2.7 GHz.

### Procedure:

Example data: Your amplifier's 1 dB compression point is 27 dBm. The nominal gain is approx. 26 dB. A 40 dB attenuator has been added to the circuit ahead of the analyzer.

- ▶ Make the following settings on the analyzer:

Reset (to the default settings)

<i>Reference</i>	<i>Internal</i>
<i>Center Frequency</i>	900 MHz
<i>Span</i>	10 kHz
<i>Reference Level</i>	-20 dBm

- ▶ Make the following settings on the generator:

Reset (to the default settings)

<i>Reference</i>	<i>External</i>
<i>Frequency</i>	900 MHz
<i>Level</i>	-5 dBm
<i>RF</i>	<i>On</i>

- ▶ Measure the power that arrives at the analyzer. Use the *Marker to Peak* function to do this (in the example, approx.: -45.4 dBm).
- ▶ Determine the path attenuation by taking the difference between the power delivered by the generator and the power measured on the analyzer, and note this value (in the example: -5 dBm - (-45.4 dBm) = 40.4 dB). This value will later be set on the analyzer as the *Ref Level Offset*.
- ▶ Repeat this measurement at 1.8 GHz and 2.7 GHz.

### Preparing the measurement on the amplifier

- ▶ On the generator, switch the *RF* to *Off*.
- ▶ Reduce the level on the generator to -20 dBm.
- ▶ Set the *Reference Level* on the analyzer to +30 dBm.

## 11.3 Measuring

- ▶ Connect the test setup as depicted in Fig. 61 on page 81.
- ▶ Supply the amplifier with operating voltage.

First, the fundamental should be measured in ten steps of 1 dB each; in the example, this would then be for output powers of 15 dBm to 24 dBm. For 15 dBm, the amplifier used here, with a gain of approx. 26 dB, must be driven at approx. –11 dBm. Therefore, set the generator as follows:

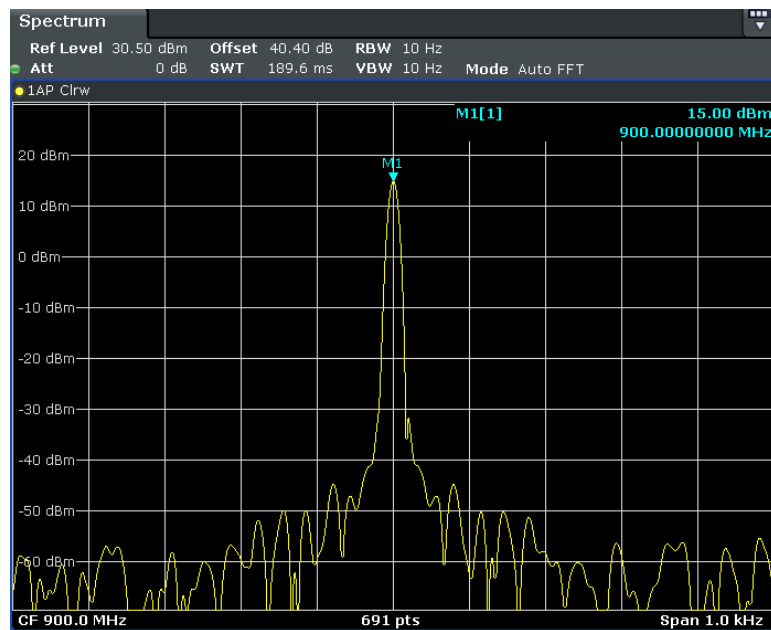
<i>Frequency</i>	900 MHz
<i>Level</i>	–11 dBm in the example
<i>RF</i>	<i>On</i>

Set the analyzer as follows:

<i>Center Frequency</i>	900 MHz
<i>Reference Level</i>	30 dBm
<i>Ref. Level Offset</i>	as described above; in the example: 40.4 dB at 900 MHz
<i>Span</i>	1 kHz
<i>Res Bandwidth RBW</i>	10 Hz

By synchronizing the generator and analyzer using the 10 MHz reference signal, you achieve very high frequency accuracy. In this example, the mixer level is at approx. –25 dBm, which is nearly ideal. Due to the narrow span, the display's frequency resolution (pixel resolution) is very good. Consequently, the measurement is very precise; the results are stable. The narrow RBW ensures low noise while maintaining a fast sweep time.

- ▶ Set the Marker *to Peak*.
- ▶ Now adjust the level on the generator so that the analyzer displays exactly 15 dBm; see Fig. 62 on page 85.



**Fig. 62: Adjusting the fundamental's level.**

Since the filter(s) and cabling on the amplifier-input side are lossy, the required level is somewhat higher than originally estimated ( -10.26 dBm in the example).

- ▶ Enter the level that is displayed on the generator, along with the measurement value from the analyzer, in a table.
- ▶ Raise the generator level by exactly 1 dB.
- ▶ Repeat the measurement until you reach an amplifier output of 24 dBm.

Then measure the second and third harmonic in the same way:

- ▶ To do this, change the generator and analyzer frequency; if necessary, correct the *Ref Level Offset* on the analyzer.
- ▶ Use exactly the same generator level as with the fundamental!

The figures below show measurement results taken at 1.8 GHz and 2.7 GHz. This shows the analyzer's large dynamic range.

How do the levels of the analyzer mixer differ for the three frequencies?

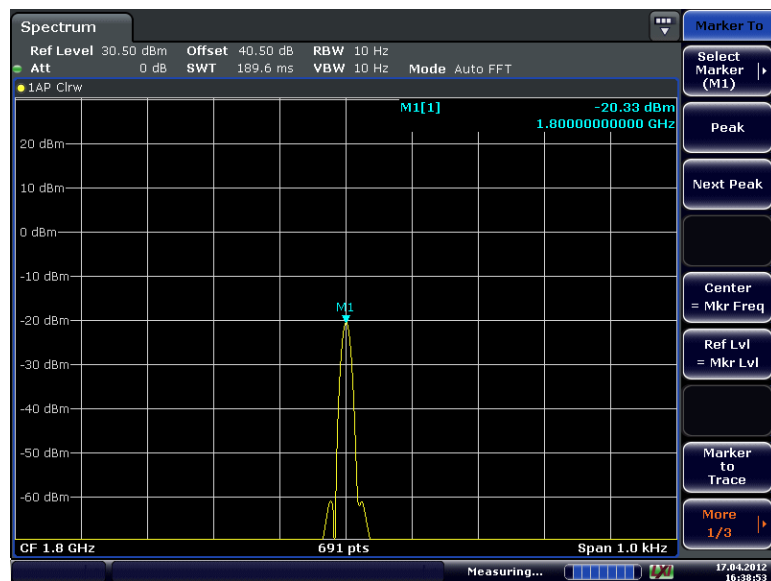


Fig. 63: Second harmonic on the amplifier output at an output power of 15 dB.

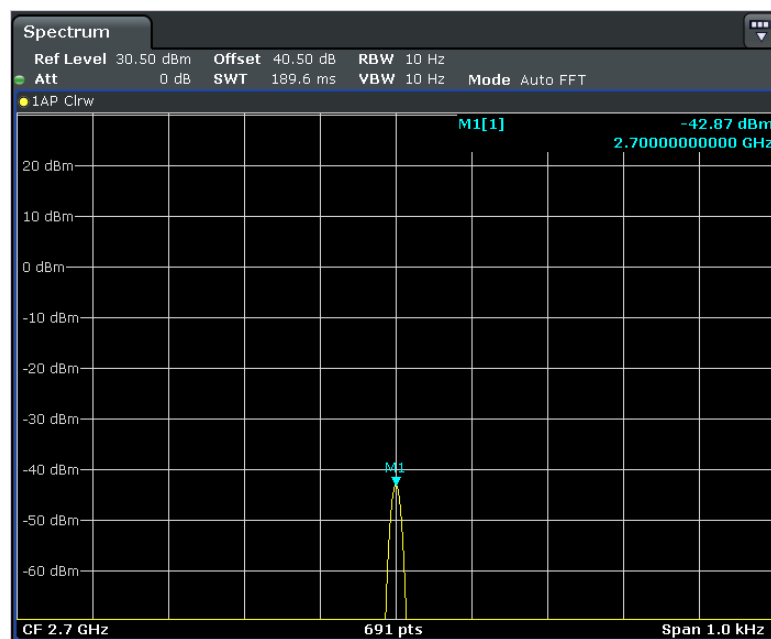


Fig. 64: Third harmonic on the amplifier output at an output power of 15 dB.

- Use the "fundamental / 2nd harmonic / 3rd harmonic" value triplet for an output power of 17 dBm and for 21 dBm to calculate the IP2 (SHI) and IP3 (THI) intercept points in each case. For this, use Eq. (32) on page 53.

In the example, this results in the following values for

17 dBm:  $SHI = 17 \text{ dBm} + 33.5 \text{ dB} = 50.5 \text{ dBm}$

$THI = 17 \text{ dBm} + 26.4 \text{ dB} = 43.4 \text{ dBm}$

And for

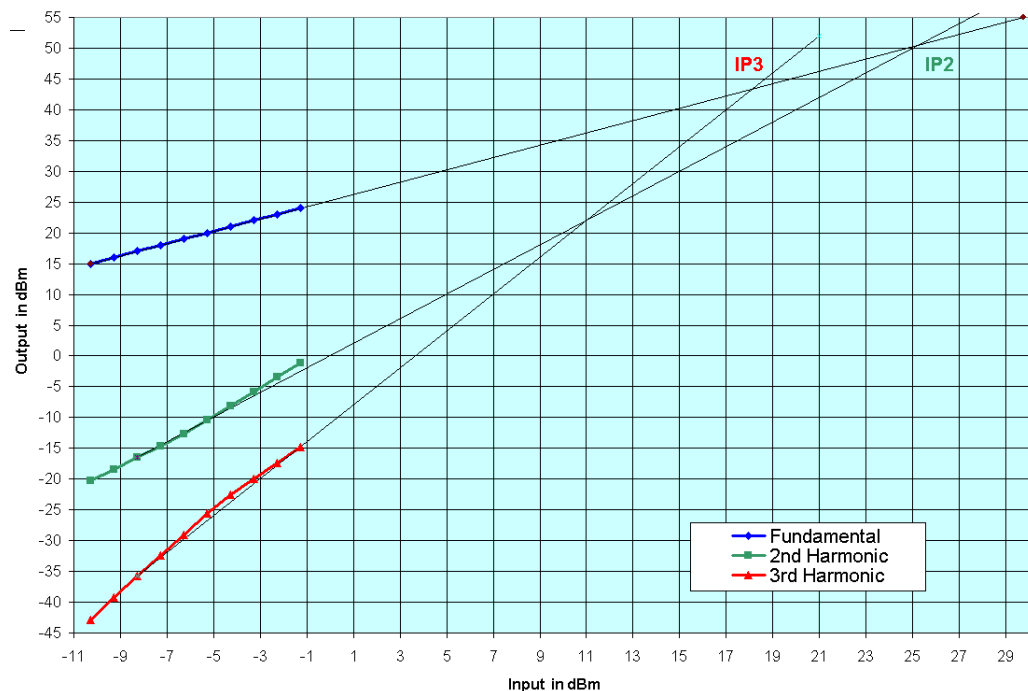
21 dBm:  $SHI = 21 \text{ dBm} + 29.2 \text{ dB} = 50.2 \text{ dBm}$

$THI = 21 \text{ dBm} + 21.8 \text{ dB} = 42.8 \text{ dBm}$

- Plot your measurement results on a graph (output power vs. generator level) and determine IP2 and IP3 by extending the recorded curves with straight lines of the correct slope. For the base point of the straight lines, use the measurement results for the 17 dBm output power.

How steep are the slopes for the individual curves (theoretical and measured)?

Fig. 65 shows a graph of the results for a measurement series of this type.



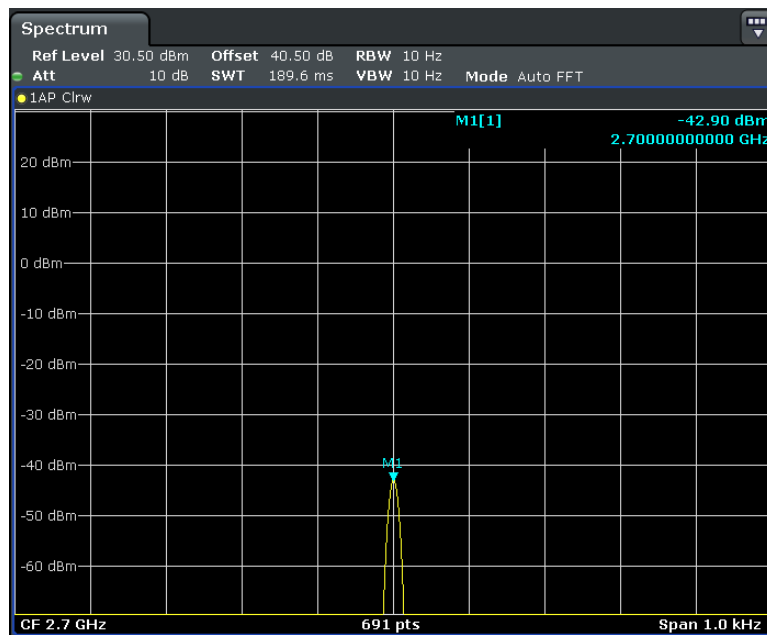
**Fig. 65:** Using a graph to determine the (harmonic) intercept points.

In this case, IP3 is at approx. 43 dBm and IP2 at approx. 50 dBm. In this respect, Fig. 65 shows good agreement between the calculation and the interpolation. The characteristic curves for the harmonics deviate somewhat from ideal straight lines. Nevertheless, the tendency of the quadratic and cubic slopes is clearly visible as the level rises. It can be seen that results achieved when determining the intercept points graphically depend heavily on which point on the characteristic curve that the approximation line is to go through. Here, the measurement results for an input signal of 10 dB below the 1 dB compression point were used as the base points.

To be certain of your results, you should check to see if the measured harmonics are actually arising in the amplifier, or if they are being caused by undesired intermodulation products in the analyzer itself. If the measured harmonics are arising in the analyzer, the measurement results for the harmonics ought to go down when the signal at the mixer is reduced. This can easily be checked by increasing the attenuation (Att) in the analyzer.

If the measured harmonics were arising in the analyzer, how many dB would the second or third harmonic drop, if the RF attenuation was raised by 10 dB?

- ▶ Repeat one measurement each at 1.8 GHz and 2.7 GHz, but this time with 10 dB RF attenuation. Fig. 66 shows a measurement of that type for the third harmonic. The measurement result remains the same. That means that the harmonics are arising in the amplifier, not in the amplifier.



**Fig. 66: Checking the analyzer linearity.**

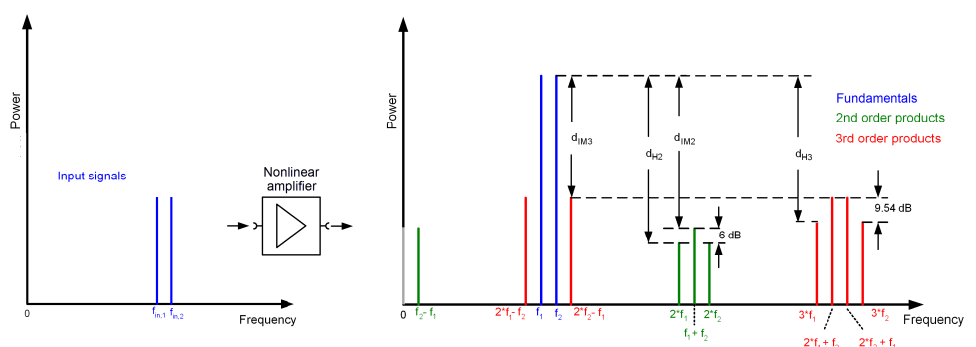
Measuring an amplifier's harmonics is a demanding task for test and measurement equipment. In order to achieve the required dynamic range on the analyzer, it is necessary for the measurements to be performed with a very narrow bandwidth. The longer settling times that narrowband filters have drastically limit the permissible sweep speed. Consequently, in practice, measurements are limited to small observation range within the spectrum (with a span of just a few kHz). The fundamental and harmonics must then be measured individually.

- ▶ Summarize the goals, execution and results of this experiment in your own words.



## 12 Intermodulation Measurements, Two-Tone Excitation

The following task employs the two-tone measurement method. Here, two sinusoidal signals on adjacent frequencies are feed into an amplifier. Then, besides the fundamental and harmonics, intermodulation products of the fundamental and the harmonics also arise at the output; see Fig. 67. In practice, these intermodulation products cause much more interference than the harmonics themselves. Above all, some of the third-order intermodulation products at an amplifier's output arise near the operating frequencies; therefore, they generally cannot be filtered out. They are theoretically 9.54 dB above the third harmonic.



**Fig. 67: Input and output spectrum with fundamentals as well as second and third-order products. (Higher-order products are not shown here.)**

Fig. 67 shows the relationship of intermodulation products and harmonics to the fundamental. The distances for the individual intermodulation products  $d_{IMv}$  are lower than the distances for the corresponding harmonics  $d_{Hv}$ . For this reason, instead of the measuring the harmonic level, which can be very low in some case, it is advantageous to measure the significantly higher level of the intermodulation products.

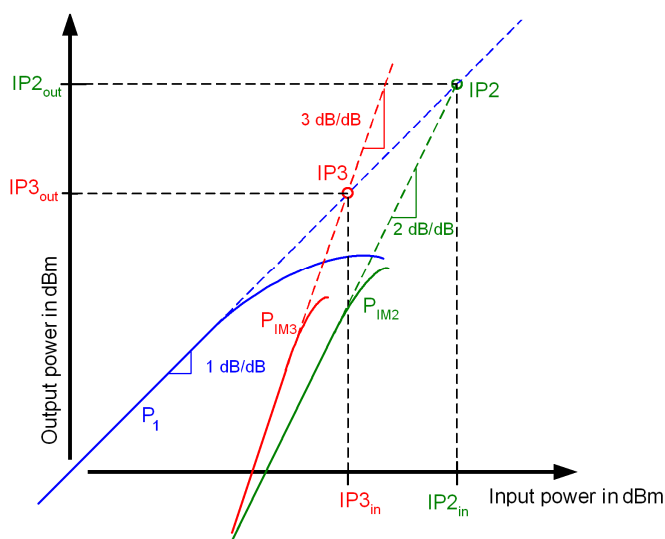
The current level ratios indicated for the spectral components are only valid for the current excitation. If the input signal is changed by 1 dB (within the linear range of the amplifier), each of the  $n$ th-order components change by  $n$  dB; that holds true for both the harmonics and the intermodulation products.

The characteristic curves for output versus input of harmonics and intermodulation products are straight lines (within the amplifier's linear range). If these straight lines are extended beyond the permissible operating range, the points at which the lines intersect determine the intercept points (IPs; see Fig. 68).

The names for the intermodulation products are simplified to second-order intercept point (SOI) and third-order intercept point (TOI).

The intercept points for the harmonics are referred to as the second-order harmonic intercept point (SHI) and third-order harmonic intercept point (THI).

Further information is available, for instance, in the white paper "Interaction of Intermodulation Products between DUT and Spectrum Analyzer" [4].



**Fig. 68:** Basic characteristic curves for the first (fundamental, blue), second (green) and third (red) harmonic.

The next task is to determine the third-order intercept point (TOI) using the two-tone method.

Therefore, it is necessary to measure the third-order intermodulation product. This product can be found to be to the "left" and "right" of the input frequencies by the amount equal to the difference between the input signals (see Fig. 67 on page 89). One disadvantage in practice proves to be an advantage when it comes to performing this task: Unlike the process for measuring harmonics, it is possible to acquire the fundamental and the third-order intermodulation with a single span. In practice, this intermodulation product is an interfering signal; for this task, however, it is the wanted signal. The measurement accuracy is good, because these lines are very pronounced. To perform this task, the analyzer only requires an average dynamic range.

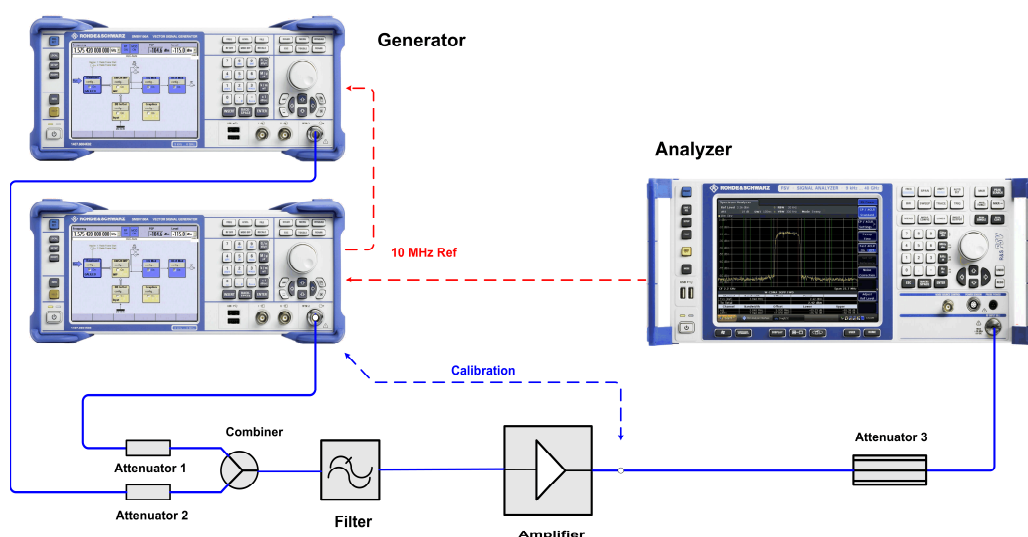
Nevertheless, it is still important to carefully ensure that the analyzer is being operated in its optimal level range and that it is not delivering intermodulation products of its own.

In addition, the amplifier must be supplied with a signal that is largely free of harmonics.

## 12.1 Test Setup

Arbitrary waveform generators are usually able to generate a multicarrier signal (and thus also a two-tone signal). In general, however, the output signal is not sufficiently free of intermodulation products. The generator's third-order products cannot be suppressed with a filter.

For this reason, the test setup uses two generators (see Fig. 69). Having attenuators 1 and 2 in the paths makes it possible to decouple the instruments before a combiner brings the two signals together. A low-pass filter after the combiner suppresses possible harmonics. This makes a spectrally pure two-tone signal available.



**Fig. 69: Hardware setup for the two-tone measurement.**

The spectrum analyzer measures and analyzes the output signal from the amplifier (DUT). The attenuator between the amplifier and analyzer ensures that the analyzer operates within the optimal level range and simultaneously provides protection against overload and damage.

The first generator receives a 10 MHz reference signal from the analyzer, and second generator receives one from the first generator.

To calibrate the measuring path after the amplifier (described further below) the cable at the amplifier output is connected directly to a generator.

- ▶ Never operate the amplifier with an open input or open output. The input is terminated when you connect the amplifier with the generator, even if the generator is still turned off. The components that are connected to the output must be able to cope with the amplifier's power.

#### Configuring the attenuators:

- ▶ Note your amplifier's minimum gain and its output power at the 1 dB compression point. You will find this value in your amplifier's data sheet.
- ▶ Refer to the analyzer manual to find the optimal mixer level for minimizing distortion.

The "mixer level" is the level at the analyzer input minus the attenuation that has been set on the instrument for the attenuator (the RF attenuation); see Fig. 20 on page 32. To achieve the best possible ratio between the signal and the inherent noise, this level is set to 0 dB for this measurement.

The measurement should be performed in the amplifier's linear range; to be safe, it should be taken 6 dB below the 1 dB compression point.

Determine the attenuator value 3 in the test setup shown in Fig. 69 so that, at a signal that is 6 dB below the 1 dB compression point, the mixer level is within the range that has the least amount of distortion.

**Example:**

The amplifier's output power at the 1 dB compression point: 27 dBm

The range that is reliably linear for this amplifier: 27 dBm – 6 dB = 21 dBm

Optimal level on the analyzer's mixer (data sheet): about –19 dBm

-> Optimal attenuator: 21 dBm – (–19 dBm) = 40 dB

- ▶ The attenuator must be able to reliably cope with the amplifier's output power.

In this example: 27 dBm  $\cong$  0.5 W -> Load capacity of Attenuator 3: ideally 1 W.

The example DUT's minimum gain is 24 dB. Having 21 dBm at the output then requires at least –3 dBm at the amplifier's input. The cabling, combiner and filter on the input-signal side add approx. 5 dB of additional attenuation. If 10 dB is selected for Attenuator 1 and Attenuator 2, the generators must supply approx. 12 dBm. This is ensured.

(At 10 dB each, attenuators 1 and 2 sufficiently decouple the signal sources.)

## 12.2 Calibration

- ▶ Disconnect the amplifier power supply.
- ▶ Connect the cable that is to lead to the amplifier output directly to a generator. The measuring path consists of the cabling and attenuator(s). The attenuation of the path "ahead" of the amplifier is unimportant.
- ▶ Now measure the insertion loss at 899.5 MHz (or 900.5 MHz).

**Procedure:**

Example data: The amplifier's 1 dB compression point is 27 dBm; the minimum gain is approx. 24 dB. A 40 dB attenuator has been added to the circuit ahead of the analyzer.

- ▶ Make the following settings on the analyzer:

<i>Reset</i> (to the default settings)	
<i>10 MHz Reference</i>	<i>Internal</i>
<i>Center Frequency</i>	900 MHz
<i>Span</i>	10 kHz
<i>Reference Level</i>	–20 dBm

- ▶ Make the following settings on the generators:

*Reset* (to the default settings)

<i>10 MHz Reference Frequency</i>	<i>External</i>
<i>Level</i>	899.5 MHz and 900.5 MHz
<i>RF</i>	–5 dBm
	<i>On</i>

- ▶ Measure the power that arrives at the analyzer. To do this, use *Marker to Peak* function (in the example, –45.4 dBm was measured).
- ▶ Determine the path attenuation by taking the difference between the power delivered by the generator and the power measured on the analyzer, and note this value (in the example: –5 dBm – (–45.4 dBm) = 40.4 dB).
- ▶ Set this value as the *Ref Level Offset* on the analyzer.

#### Preparing the measurement on the amplifier

- ▶ On the generators, switch the *RF* to *Off*.
- ▶ Set the *Reference Level* on the analyzer to +30 dBm.

## 12.3 Measuring

- ▶ Connect the test setup as depicted in Fig. 69 on page 91.
- ▶ Supply the amplifier with operating voltage.

The measurement should first be taken at an output power that is 6 dB below the 1 dB compression point, which is 21 dBm. For this, the example amplifier – with a min. gain of 24 dB – must be driven at approx. –3 dBm. Attenuator 1 or 2, the cable, combiner and filter at the amplifier input attenuate the signal by approx. 15 dB. For this reason, the generators must deliver approx. 12 dBm to achieve the desired output power.

Set the generators as follows:

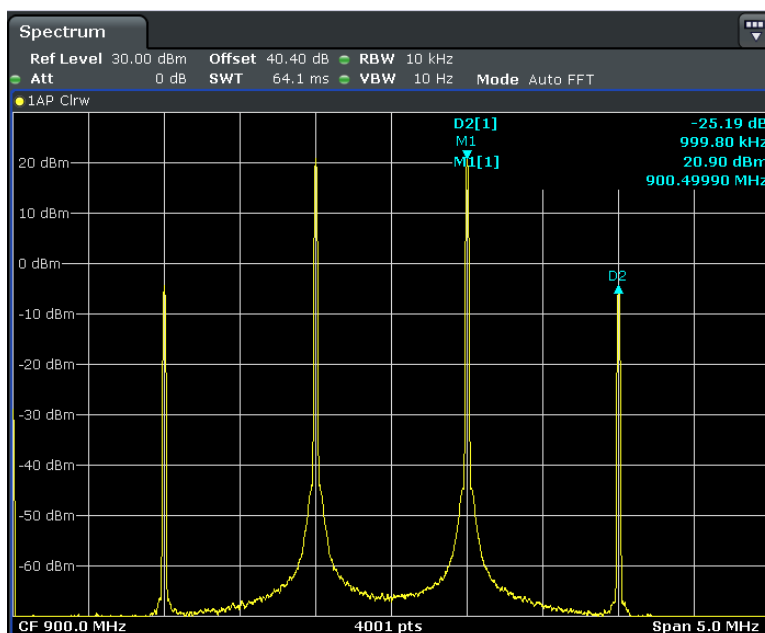
<i>Reference Frequency</i>	<i>External</i>
<i>Level</i>	899.5 MHz and 900.5 MHz
<i>RF</i>	10 dBm (in the example)
	Both <i>On</i>

Set the analyzer as follows:

<i>Center Frequency</i>	900 MHz
<i>Reference Level</i>	30 dBm
<i>Ref. Level Offset</i>	As described above; 40.4 dB in the example
<i>RF Attenuation, Manual</i>	0 dB
<i>Span</i>	5 MHz
<i>Sweep Points</i>	4001
<i>Res. Bandwidth</i>	10 kHz
<i>Video Bandwidth</i>	10 Hz

The high number of sweep points and the synchronization of the generator and analyzer using the 10 MHz reference result in good marker accuracy. The mixer level is in the optimal range. Due to the narrow span, the display's frequency resolution (pixel resolution) is very good. Consequently, the measurement is very precise; the results are stable. The narrow RBW ensures low noise while maintaining a fast sweep time.

- ▶ Set a marker to *Peak*.
- ▶ Now adjust the level on the corresponding generator so that the analyzer displays  $21 \text{ dBm} \pm 0.1 \text{ dBm}$ .
- ▶ Set the marker on the peak of the second generator.
- ▶ Now adjust the level on the other generator so that the analyzer also displays exactly  $21 \text{ dBm} \pm 0.1 \text{ dBm}$ .
- ▶ Now set a delta marker on an adjacent intermodulation product.
- ▶ Note (the power for the fundamental and) the intermodulation distance.



**Fig. 70: Third-order intermodulation distance at a maximum power of  $-6 \text{ dB}$ .**

- ▶ Calculate the TOI.

The following can be derived from Fig. 70:

$$20.9 \text{ dBm} + \frac{25.19}{2} \text{ dB} = 33.5 \text{ dBm}$$

- ▶ On one of the generators, switch *RF* to *Off*.

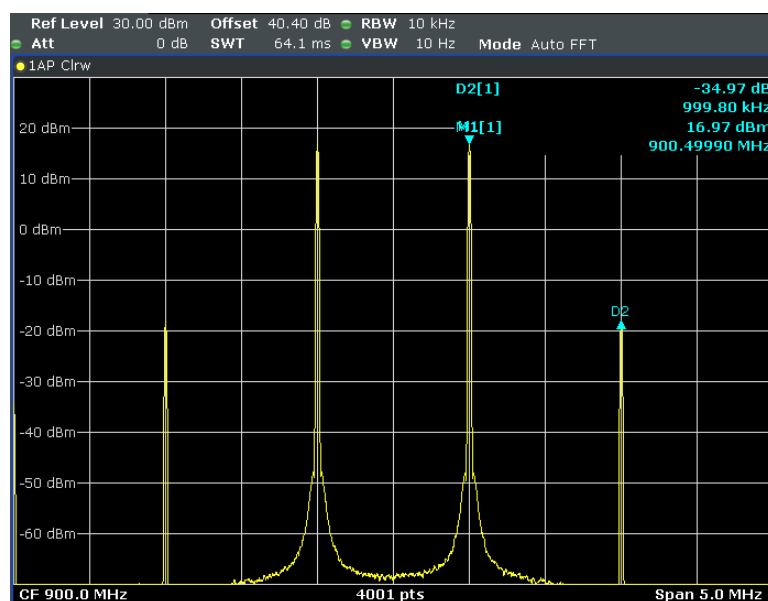
How is it possible to explain what is displayed on the analyzer?

Theoretically, intercept points are independent of the measurement point, as long as the measurement is performed within the amplifier's linear range.

- ▶ Switch the *RF* back to *On*.
- ▶ Note the current generator level (for the 21 dBm amplifier output power).
- ▶ First reduce the output level for one of the generators, then for the other, until the fundamental peaks on the analyzer indicate 17 dBm  $\pm$  0.1 dBm.

Is the current generator level exactly 4 dB lower?

- ▶ Calculate the TOI.



**Fig. 71: Third-order intermodulation distance at a maximum power of -10 dB.**

The following can be derived from Fig. 71:

$$16.97 \text{ dBm} + \frac{34.97}{2} \text{ dB} = 34.45 \text{ dBm}$$

- ▶ Compare your measurement result with the value specified in the amplifier's data sheet.
- ▶ Where is the third-order harmonic intercept point?

In the event that the intermodulation products arose in the analyzer, how many dB would the corresponding line then have to go down, if the RF attenuation is raised by 10 dB?

- ▶ Repeat the measurement with an RF attenuation of 10 dB. Fig. 72 shows a measurement of that type; the intermodulation distance has remained unchanged. That means that the intermodulation products are arising in the amplifier, not in the analyzer.

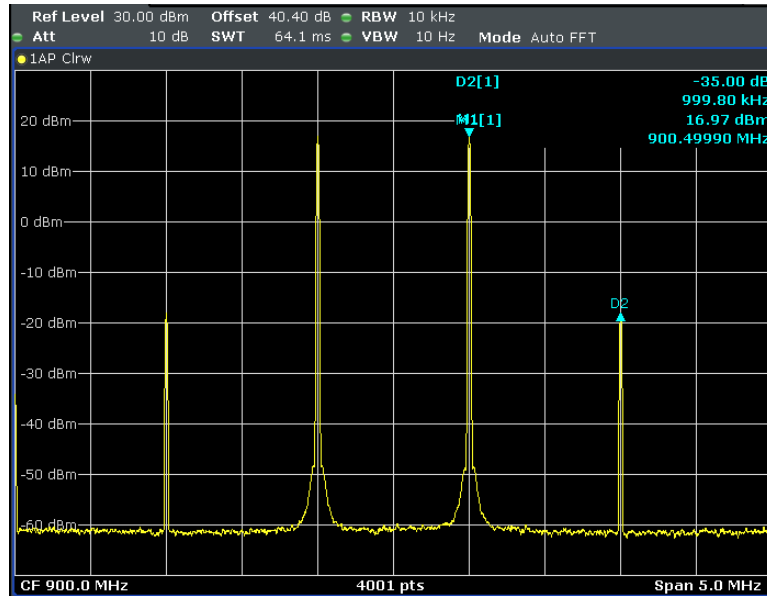


Fig. 72: Verification measurement with an RF attenuation of 10 dB.

State-of-the-art spectrum analyzers offer many automatic measurement functions. One of these functions searches the screen section for peaks that would be plausible for a two-tone measurement and calculates the TOI with every sweep (see Fig. 73).

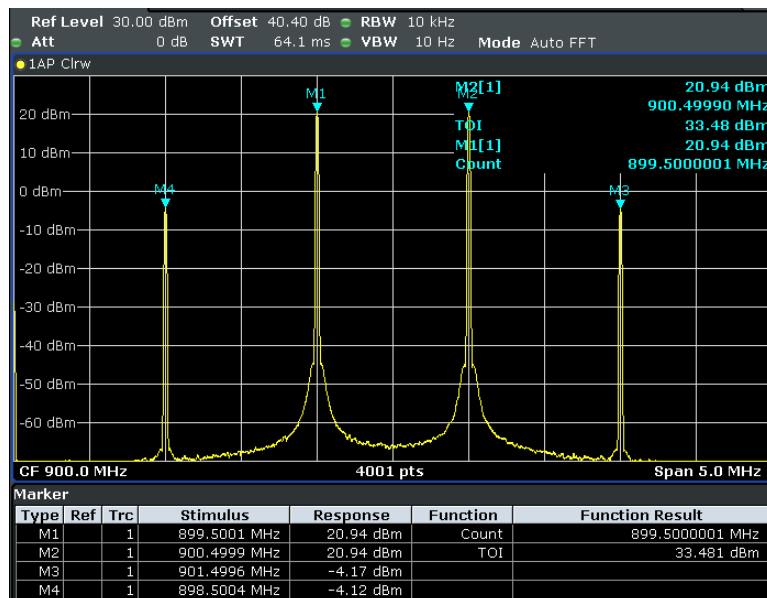


Fig. 73: TOI measurement function.

- ▶ Summarize the goals, execution and results of this experiment in your own words.



## 13 Crest Factor, ACLR in Wireless Communications

When drive-signal peaks extend into an amplifier's nonlinear range, powerful, undesired intermodulation products arise. In mobile (cellular) networks, for example, these intermodulation products then also interfere with the neighboring channels being used by other subscribers. For this reason, all cellular telephony standards require tests in order to ensure that these interference components remain below the useful-channel power by a certain minimum amount, which is known as the minimal "adjacent channel leakage ratio" (ACLR).

The purpose of this lab exercise is to observe how the ACLR changes at an amplifier output depending on the amplifier gain. The main goal here is to determine the quality of the amplifier's behavior.

To do this, you employ a signal generator to create a typical 3GPP-FDD cellular network signal (the signal from a UMTS/W-CDMA base station) that has a crest factor of 10.55 dB. First, you need to perform measurements on the generated signal:

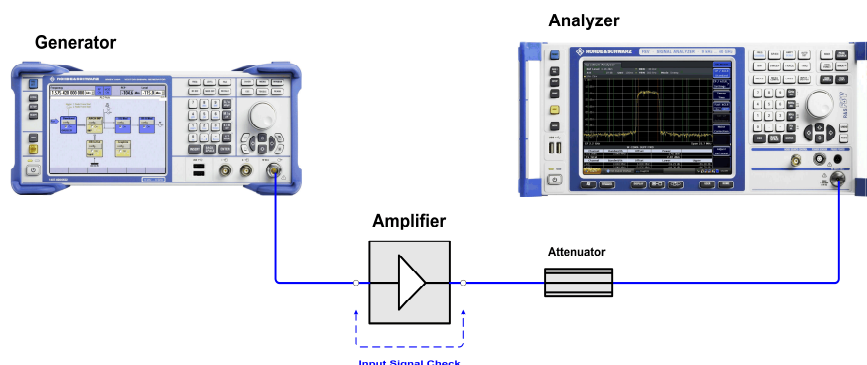
- Using the CCDF to determine the probabilities of levels that exceed the average RMS value and
- Using the ACLR to determine the differences between the power in the adjacent radio channels and the power in the subscriber channel.

After that, you excite an amplifier with this signal. In the spectrum, you then observe how the powers in the useful and adjacent channels change at the output depending on the characteristics of the amplifier input.

Finally, you examine the effect that clipping has on the crest factor and on the ACLR.

### 13.1 Test Setup

The measurement hardware needed to perform the CCDF and the ACLR measurements on an amplifier are a signal generator and a spectrum analyzer, see Fig. 74.



**Fig. 74:** Hardware setup for measuring the amplifier's input and output signals.

In this lab exercise, the generator supplies a 3GPP FDD cellular network signal (UMTS/W-CDMA) that has a 10.55 dB crest factor. The quality of this drive signal will be determined in advance in the first half of this exercise. To do that, a direct connection replaces the amplifier. The spectrum analyzer measures the generator signal's CCDF, crest factor and ACLR. This is accomplished at exactly the generator level that will later be used to drive the amplifier at the 1 dB compression point.

After that, the amplifier is inserted again. Beginning at a low generator level, the drive signal is increased incrementally until the ACLR fails to maintain the minimum distance that the UMTS/W-CDMA wireless communications standard demands.

To prevent the analyzer from overloading and being damaged, and to work within the analyzer's ideal operating range, an attenuator is inserted into the measurement path. Analyzing the input and output signal requires different attenuators with different power ratings (and different  $Att_{ext}$  settings).

(The amplifier requires a power supply. That device is not shown in Fig. 74.)

- ▶ Never operate the amplifier with an open input or open output. The input is terminated when you connect the amplifier with the generator, even if the generator is still turned off. The components that are connected to the output must be able to cope with the amplifier's power.

### Configuring the attenuator rating:

At first, the measurement signal is the generator's drive signal; later it is the amplifier's output signal. Analyzing the drive signal and the output signal requires different levels of attenuation in the measurement path.

- ▶ Note your amplifier's gain and its output power at the 1 dB compression point. You will find this value in your amplifier's data sheet. Calculate the drive level required for this and write down the results.

Refer to the analyzer manual to find the optimal mixer level for minimizing distortion. The mixer level is the level at the analyzer input minus the attenuation that has been set on the instrument for the attenuator (the RF attenuation); see Fig. 20 on page 32. In order to achieve the greatest possible dynamic range, you should first attempt to perform these measurements at a mixer level that is 10 dB higher. If, however, you determine that this drives the analyzer into saturation, attenuation must be increased at the expense of decreasing the dynamic range.

Calculate the required attenuation  $Att_{ext}$  in line with the standard:

$$Att_{ext} = \text{meas. Signal} - (\text{optimal mixer level} + 10 \text{ dB})$$

### Example:

According to the specification sheet, the optimal level on the mixer for the analyzer being used is  $-22 \text{ dBm}$  to  $-17 \text{ dBm}$ .

The amplifier's output power at the 1 dB compression point is 26 dB.

The attenuation for measuring the amplifier's *output* signal is calculated as follows:

$$Att_{\text{ext}} = 26\text{dBm} - (-17\text{dBm} + 10\text{dB}) = 33\text{dB}$$

Typically, the DUT gain is 26 dB, and the output power for the 1 dB compression point is 26 dBm. This results in the following amplifier *drive* power at the 1 dB compression point:

$$(26\text{dBm} + 1\text{dBm}) - 26\text{dB} = 1\text{dBm}$$

(The fact that the 1 dB compression point is at a 1 dBm input power in this case is a coincidence!)

$$Att_{\text{ext}} = 1\text{dBm} - (-17\text{dBm} + 10\text{dB}) = 8\text{dB}$$

In this example, 10 dB is selected.

- ▶ Calculate the value that the attenuators must have for the two measurements in your case.

## 13.2 Calibration

This lab exercise focuses more on the qualitative results than on the quantitative results. For this reason, no calibration of the measurement path is performed in this case.

## 13.3 Measuring

### Drive Signal:

- ▶ Make the following settings on the generator:

*Preset*  
*Frequency*                      900 MHz

- ▶ Make the following settings on the analyzer:

*Reset*  
*Level Offset*      *Value for the attenuator in the measurement path (10 dB in the example)*

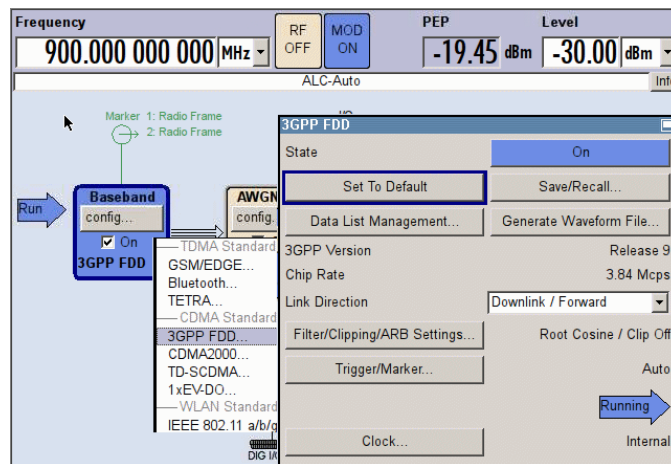
- ▶ Connect the generator to the analyzer via the appropriate attenuator as shown in Fig. 74 (without amplifier).

Configure the generator to generate a standard 3GPP FDD signal (downlink).

On the example devices, this is performed as follows:

- ▶ Select the following settings on the generator:

<i>Baseband</i>	<i>Config.</i>
<i>CDMA Standards</i>	<i>3GPP-FDD</i>
<i>State</i>	<i>Set to Default</i>
	<i>On</i>



**Fig. 75: Sequence of generator settings for a 3GPP FDD cellular network signal.**

- ▶ Now switch to:

*RF*    *On*

This makes the cellular network signal available at the generator output.

At the top right of Fig. 75, you can see two power values. The peak (maximum value), given as the *peak envelope power*, or "*PEP*" (−19.45 dBm in this case), and the average value, given as the *Level* (−30 dBm here).

- ▶ Calculate the crest factor.

The *PEP* value indicated on the generator comes from the internal calculation of the signal, not from a measurement. Consequently, that value is not dependent on the length of a particular observation period.

**In this lab exercise, the *PEP* is the relevant value.** You can only make the settings for this indirectly: When you vary the *Level*, the *PEP* automatically changes along with it.

- ▶ Now set the *peak envelope power* to the value that you have determined for later excitation of the amplifier at the 1 dB compression point; in the example, this would be a PEP of 1 dBm.
- ▶ Calculate the crest factor.
- ▶ Familiarize yourself with the drive signal. Use the analyzer to perform measurements in the time and frequency domains. To do that, select the RMS detector.

3GPP-FDD signals have a bandwidth somewhat below 5 MHz. If measurements are performed with a smaller resolution bandwidth, the analyzer only acquires a portion of the overall power. The trace shown is then lower by this amount:

$$10\lg\left(\frac{RBW}{SignalBandwidth}\right) \text{ dB}$$

(That is only the case for signals that have the same amplitude distribution above the frequency.)

- ▶ Observe the signal in the frequency domain with these different resolution bandwidths: 5 MHz, 300 kHz and 30 kHz (with the RMS detector active).

By how many dB does the displayed result go down when the RBW is 1/10th or 1/100th the signal bandwidth?

### Signal statistics / CCDF of the generator signal

- ▶ Leave the generator signal unchanged.
- ▶ Enter these settings on the analyzer:

<i>Reset</i>	
<i>Frequency</i>	900 MHz
<i>Level Offset</i>	<i>Value of the attenuator in the measurement path (10 dB in the example)</i>
MEAS	CCDF

Due to the relatively short acquisition time (AQT) that the system automatically selects, the display is very unsteady. You can lengthen the AQT indirectly by increasing the number of samples that are to be taken into consideration.

- ▶ To do that, set the analyzer to this value:

*# of Samples* 1000000

- ▶ Check the RBW setting. If the RBW setting is < 3.84 MHz, the analyzer only acquires a portion of the overall power.
- ▶ At what point on the trace can you read the crest factor?
- ▶ Make a note of the probability that the RMS value will be exceeded by 3 dB, 6 dB, and 10 dB during your acquisition.

Fig. 76 shows typical measurement results.



Fig. 76: CCDF of a 3GPP signal (yellow) compared to the noise (red).

The display shows the probability (yellow trace) at which instantaneous levels (greater than the RMS value) will occur.

As a reference, the red trace describes the CCDF for white noise. By using the percent marker, you can move to specific probabilities on the trace. Numeric values supplement the graphical display.

### Adjacent channel leakage ratio (ACLR) for the generator signal

- ▶ Leave the generator signal unchanged.
- ▶ Enter the following values on the analyzer:

<i>Reset</i>	
<i>Frequency</i>	900 MHz
<i>Ref Level Offset</i>	Value of the attenuator in the measurement path (10 dB in the example)
<i>RF Atten., Manual</i>	0 dB
<i>MEAS</i>	Ch Power ACLR
<i>CP / ACLR Standard</i>	W-CDMA 3GPP FDD
<i>RBW, Manual</i>	30 Hz (to achieve a steady image.)

*If necessary, optimize the reference level (Ref level) in order to lift the generator signal out of the analyzer noise across the entire span.*

You now see a display similar to the one shown in Fig. 77. The processing used in the analyzer for this has been tailored to match the needs of the UMTS/W-CDMA wireless communications standard (3GPP FDD).

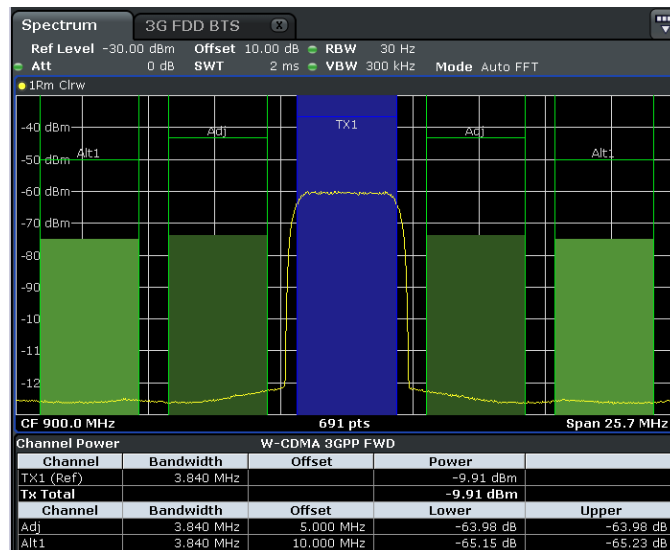


Fig. 77: Adjacent channel leakage ratio for the generator signal.

In the middle, you see the currently used channel (TX1, blue) with the adjacent channels (Adj) to the left and right (green) and the next-closest channel on each side (alternate channel, Alt1). The channel spacing is 5 MHz. The 3.84 MHz signal bandwidth is shown in each channel.

The analyzer software measures the (average) power in each channel and indicates the measurement results numerically.

In order to ensure that radiocommunications in the adjacent channels remain largely undisturbed, the ETSI approval tests for base and mobile stations require verification of the ACLR as a mandatory measurement. For 3GPP base stations, the ACLR limits are 45 dB for the adjacent channels and 50 dB for the alternate channels.

- Note your measurement results.

Does the generator comply with the ETSI limits? What do the colored bars indicate?

### Adjacent channel leakage ratio (ACLR) at the amplifier output

Your measurements should begin 12 dB below the 1 dB compression point.

- Reduce the peak power on the generator by 12 dB.

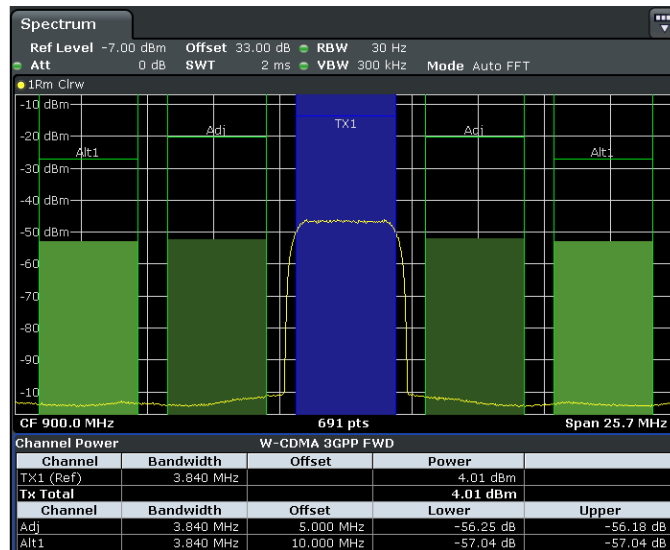
In the example, this results in PEP = -11 dBm at Level = -21.55 dBm.

- Make the following settings on the analyzer:

<i>Ref Level Offset</i>	Value of the attenuator in the measurement path
<i>RF Atten., Manual</i>	0 dB

- ▶ Insert the amplifier with the corresponding attenuator (in the example: 33 dB; keep an eye on the power rating) into the measurement path; see Fig. 74 on page 97.
- ▶ Connect the amplifier's power supply.
- ▶ Measure the ACLR. If necessary, optimize the analyzer settings (reference level?).

You now see a display similar to the one shown in Fig. 78.



**Fig. 78: ACLR at the amplifier output with a weak drive signal.**

This ACLR is significantly worse than the generator signal's ACLR!

- ▶ Enter your results for the adjacent channels in a table.
- ▶ Increase the generator power in increments of 1 dB; measure up to 2 dB beyond the 1 dB compression point. As you do that, observe how the power in the adjacent channels grow.

What causes this increase (spectral regrowth)?

- ▶ Plot your results on a graph.
- ▶ Mark the maximum permissible amplifier drive level with the cellular network signal (PEP and RMS) at which the ACLR reaches 45 dB. Reset the generator level to this value.
- ▶ Note the RMS power at the amplifier output.

What would be the maximum RMS power (at the output) at which the amplifier could be driven, if the transmitter signal were a frequency modulated signal?

With the example amplifier, the point at which the ACLR reaches 45 dB is a drive signal with approximately 0 dBm (PEP) or -10.55 dBm (RMS) at an amplifier output power of 14.9 dBm; see Fig. 79 and Fig. 80.



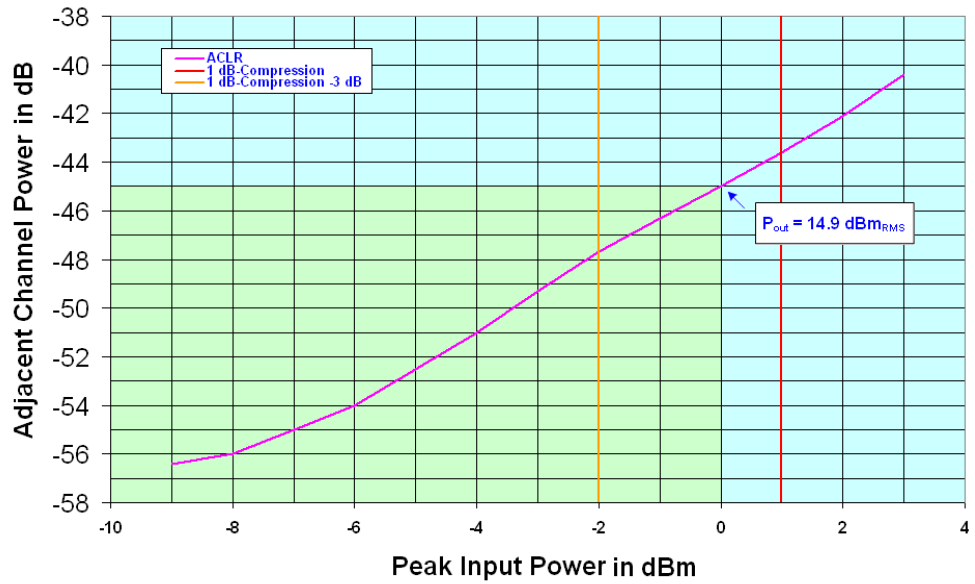


Fig. 79: ACLR versus input power.

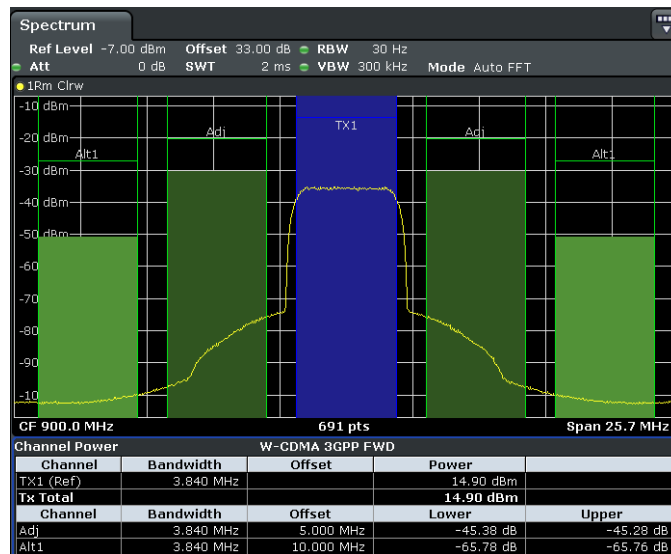


Fig. 80: Spectral regrowth; max. permissible power in the adjacent channel.

Why is the difference in power between the alternate channels and the wanted signal better here than in Fig. 78?

The fact that the increase in the adjacent channels does not come from the generator was already demonstrated by measuring the maximum generator signal's ACLR; see Fig. 77 on page 103.

How could you check to see if the growth in the adjacent channels is caused by analyzer overload?

By how many dB would third-order intermodulation products that are being caused by analyzer overload have to go down, if you were to increase the internal attenuation (Att) by 10 dB?

- ▶ Check to see if the increase in the adjacent channels is being caused by overloading of the analyzer.

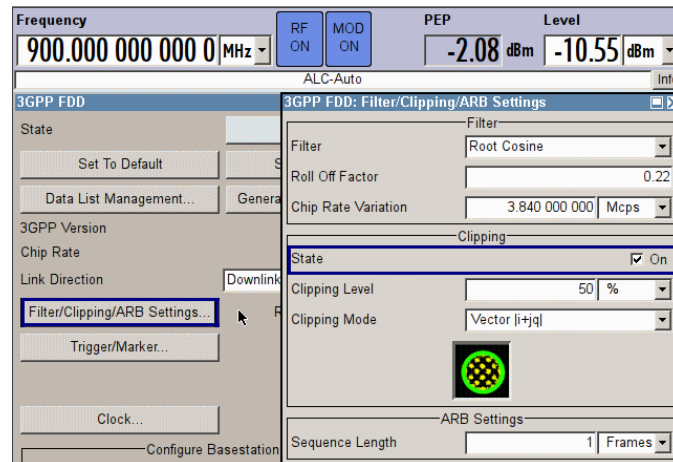
## Clipping

Clipping limits the maximum permissible instantaneous voltage or instantaneous power.

What peak values (for power and voltage) arise at an RMS power of 1 W at 50 Ohm and a crest factor of 10 dB?

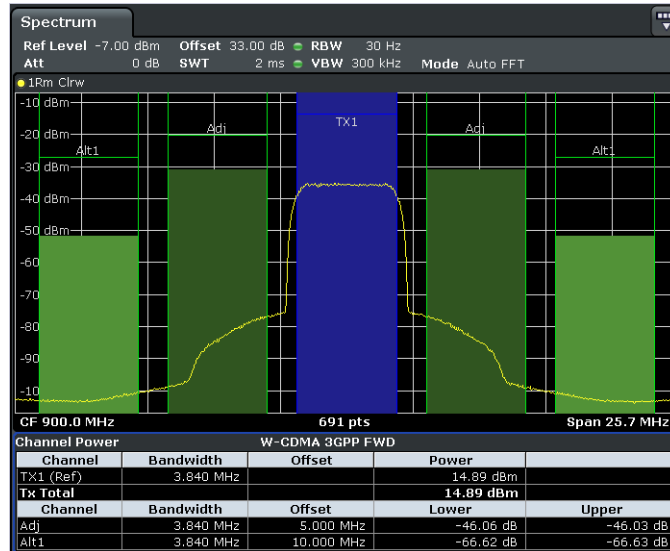
Is it possible to make a statement about the number of dBs that a signal's crest factor decreases by when 50 % clipping is used?

Set the generator to 50 % clipping; see Fig. 81 on page 106.



**Fig. 81: Setting the generator's clipping level.**

- ▶ Calculate the crest factor.
- ▶ Check to see how this clipping affects the ACLR.



**Fig. 82:** ACLR at the same RMS value, with a 50 % clipping.

Why does this clipping influence the ACLR to a certain extent?

What peak values (for power and voltage) arise at an RMS power of 1 W at 50 Ohm and a crest factor of 8 dB?

Note: Clipping reduces signal quality. The error vector magnitude (EVM) increases; the symbols form "clouds" around the exact positions in the constellation diagram.

The percentage of the signal that can be clipped without exceeding a specific EVM has to be examined in each individual case. This lab exercise does not cover that aspect.

With the 3GPP-FDD (UMTS/W-CDMA) standard, the EVM limit is 17.5 % for pure QPSK carriers and 12 % for carriers that include 16 QAM. With the signal used in this exercise, 50 % clipping results in a total EVM of 11.6 %.

- Summarize the goals, execution and results of this experiment in your own words.

## 14 Measuring the Phase Noise<sup>4</sup>

The purpose of this lab exercise is to determine the phase noise for a DUT, such as a generator, using a spectrum analyzer directly, without the aid of additional devices. The fundamental frequency is to be 1 GHz. The goal is to determine the phase noise at an offset of 1 MHz from the carrier.

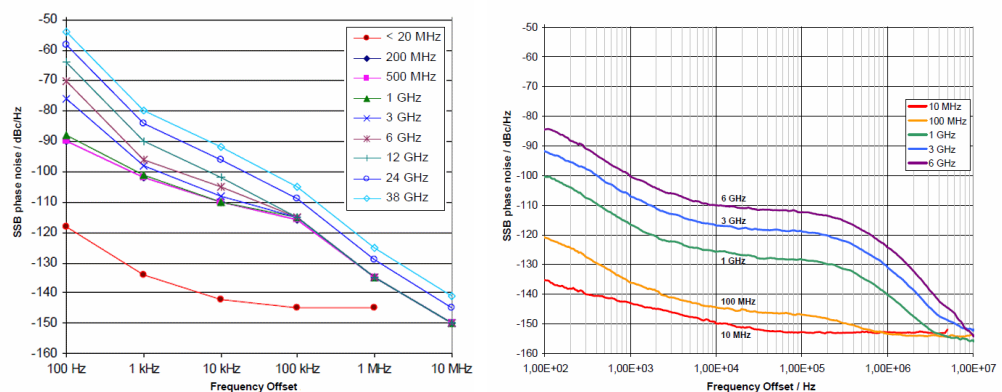
To perform a direct measurement, the following conditions must be met:

- The DUT's frequency drift must be small relative to the spectrum-analyzer sweep time.

Generally, this is always ensured. If you synchronize the DUT and analyzer via the 10 MHz reference ports, no (long-term) drift will arise.

- The spectrum analyzer's phase and amplitude noise (AM noise) should be significantly below the DUT's expected phase noise. Achieving a measurement accuracy of 0.5 dB requires a spacing of approx. 10 dB.

With the two test and measure instruments proposed in chapter 7, it is practically impossible to meet the latter of those two requirements. Fig. 83 shows the typical SSB phase noise values for these instruments.



**Fig. 83: Typical SSB phase noise for the analyzer (left) and the generator (right).**

As you can see, the generator's phase noise is lower than that of the analyzer. For this reason, at a fundamental frequency of 1 GHz and a 1 MHz offset, you will not measure phase noise below the level of approx.  $-135$  dBc, the analyzer's base level. For this reason, the measurements taken below were performed with a different generator (DUT).

<sup>4</sup> The information provided in this chapter has been taken largely from the application note "Phase Noise Measurement with Spectrum Analyzers of the FSE Family" by Josef Wolf [6].

- ▶ If possible, you, too, should use a generator that does not achieve such excellent results to serve as the DUT!

If, however, you have precise knowledge of the analyzer's phase noise, you can also derive reliable noise values for the DUT from measurements that are in the range of the analyzer's phase noise. Here, the following points need to be taken into consideration:

The measurement value that the analyzer indicates is composed of the following components:

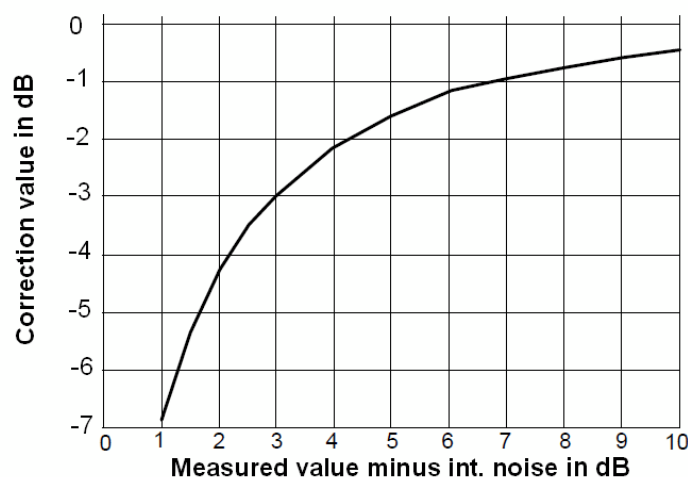
- The DUT's phase noise
- The DUT's amplitude noise
- The analyzer's phase noise
- The analyzer's amplitude noise

The amounts that the DUT and analyzer contribute are not correlated; these three power values add up to form the measurement result.

At a distance of 1 MHz from the carrier, the DUT's amplitude noise is much lower than the phase noise and can, therefore, be ignored. When that is the case, the analyzer's amplitude noise is, for the time being, not taken into consideration either; as a result, only two summands remain.

To recover the DUT's phase noise (one summand) from the sum (measurement result) and the analyzer's known phase noise (second summand) you can take an antilogarithm, subtract, and then take a logarithm again.

An easier way to do this is to take a correction value from the following chart:



**Fig. 84: Correction value for a sum measurement value when the measurement value offset is known. [6]**

For instance, if the measurement value is 10 dB above the analyzer's phase noise, that DUT's actual phase noise is as follows: Measured value – 0.5 dB.

If, for example, the measured value is 3 dB above the analyzer's phase noise, the DUT's actual phase noise is: Measured value – 3 dB.

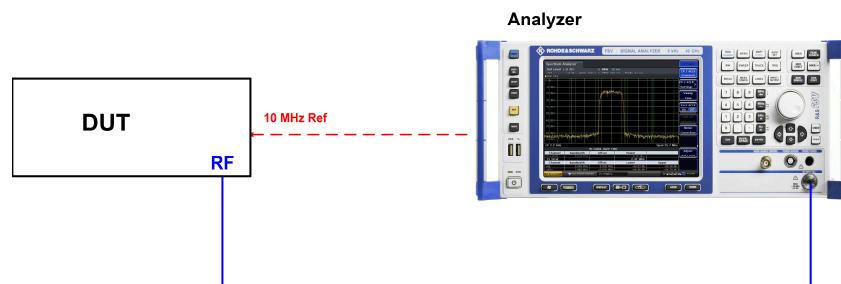
(If the measurement value theoretically supplies exactly the analyzer's phase noise, the DUT's theoretical phase noise is: Measured value – inf dB.)

After that, you determine the analyzer's amplitude noise (by measuring without an input signal) and make the correction in the same way. At a spacing of 9 dB from the measurement trace, for example, it is necessary to subtract approximately 0.6 dB.

In this experiment, as an initial step, first measure the phase noise using the analyzer's phase-noise marker. With this function, SSB phase noise is indicated directly in dBc/Hz. When calculating the measurement results, the analyzer automatically takes correction factors into account for the IF filters' noise bandwidth, for the logarithmic amplifier and for the weighting of the detector.

After that, take a second measurement – this time utilizing the support of the analyzer's the noise measurement option.

## 14.1 Test Setup



*Fig. 85: Easy connection of the DUT with the analyzer.*

The test setup consists of the DUT, such as a suitable signal generator, and a spectrum analyzer.

The DUT's RF path is connected directly to the spectrum analyzer's input port. Coupling the components via the 10 MHz reference ensures a high level of frequency accuracy and prevents (long-term) drift between the oscillators for the DUT and for the analyzer.

## 14.2 Calibration

This lab exercise examines the relationship between two amplitudes. Consequently, calibrating the measurement path is not necessary.

## 14.3 Measuring Using the Phase-Noise Marker

### Step 1: Measure the DUT's level

In order to achieve a large dynamic range, it is necessary to work with high levels on the spectrum analyzer. Make the following settings on the DUT:

<i>Reset</i>	
<i>Frequency</i>	1 GHz
<i>Level</i>	-5 dBm
<i>RF</i>	ON

(Using -5 dBm ensures that the analyzer will definitely not be damaged.)

- ▶ Make the following settings on the analyzer:

<i>Reset</i>	
<i>Frequency</i>	1 GHz
<i>Span</i>	5 MHz
<i>Ref Level</i>	-5 dBm

The *RF Att* attenuator setting should first be made automatically. Depending on the value entered for *Ref Level*, this adjusts the level on the mixer for higher measurement accuracy.

- ▶ Reduce *Res BW* to 10 kHz and *Video BW* to 10 Hz.
- ▶ Set the *Sweep Count* to 10.

Configure Trace 1 (and, if appropriate, already configure Trace 2, as well):

- ▶ *Average, RMS Detector, Average Mode Power*
- ▶ Use the marker to measure the peak level.
- ▶ Increase the level on the generator to the point at which you measure approx. -5 dBm on the analyzer.

## Step 2: Set the phase-noise marker

- ▶ Select the *Phase Noise Marker Function*.

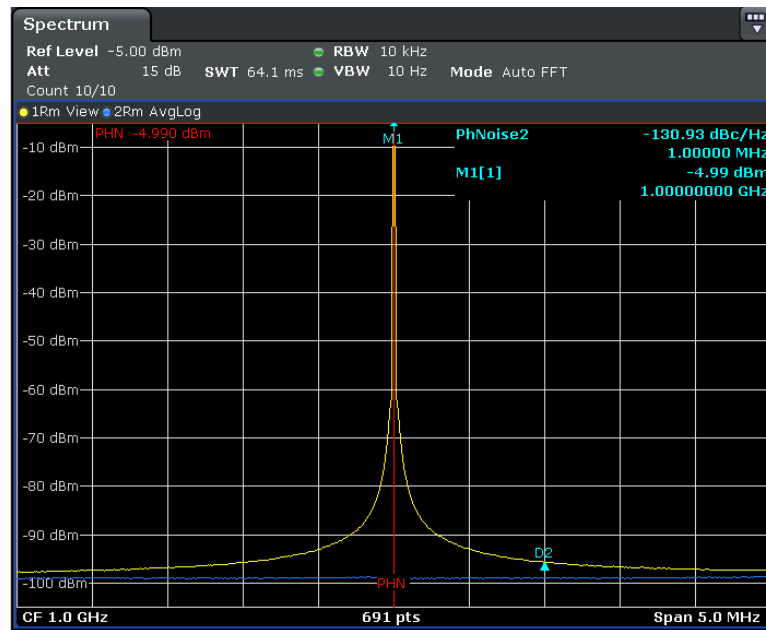
This causes the analyzer to start the phase-noise reference measurement: The reference point on the fundamental frequency (at Marker M1) is now marked with a red line; it is also indicated numerically in red and is saved internally as a reference value. Whenever you change the DUT's level, you have to turn the phase noise off and then back on to re-establish a reference!

- ▶ Select Marker D2 (softkey); turn it on as a delta marker.
- ▶ Enter the desired frequency offset (1 MHz).

In the marker field, the analyzer now indicates an initial measurement result for the phase noise directly in dBc/Hz.

- ▶ Switch Trace 1 to View.
- ▶ Turn the DUT's RF output off (or disconnect the RF connection).
- ▶ Now measure the current AM noise with Trace 2.

You now see a display similar to the one shown in Fig. 86.



**Fig. 86:** Phase-noise measurement with automatic attenuation setting (yellow). The analyzer's inherent noise (blue) limits the dynamic range.

In Fig. 86, a phase noise of  $-130.93$  dBc (1 Hz) is measured 1 MHz from the carrier. This value also contains the analyzer's AM noise (in addition to its own inherent noise). Here, at an offset of 1 MHz, the AM noise is only about 3 dB below the sum result. In the next step, you can reduce this distance considerably.

### Step 3: Minimize the AM noise

You can reach the desired increase in the dynamic range by incrementally decreasing the RF attenuation (RF Att), which had originally been selected automatically. This is permissible until a level range is reached in which the IF modules (the modules downstream from the first mixer) are already overloaded at the fundamental frequency – in other words, until the fundamental frequency is measured incorrectly. At the measurement frequency, the level has dropped to a point at which the system is able to make a correct evaluation.

Overloading the IF stages at the fundamental frequency is permissible to a certain extent, because the analyzer input itself is able to handle high levels: With the instrument used here, up to approximately  $-5$  dBm may be applied at the first mixer. It is not until the mixer – or the relevant stages upstream from the mixer – are overloaded that the analyzer reports:

*Overload.*

The measurement error caused by overloading the IF modules at the fundamental frequency has no impact on the (relative) measurement results, because the reference level was measured and recorded in advance (using the automatic attenuation setting) without overload.

The following calculation is used:

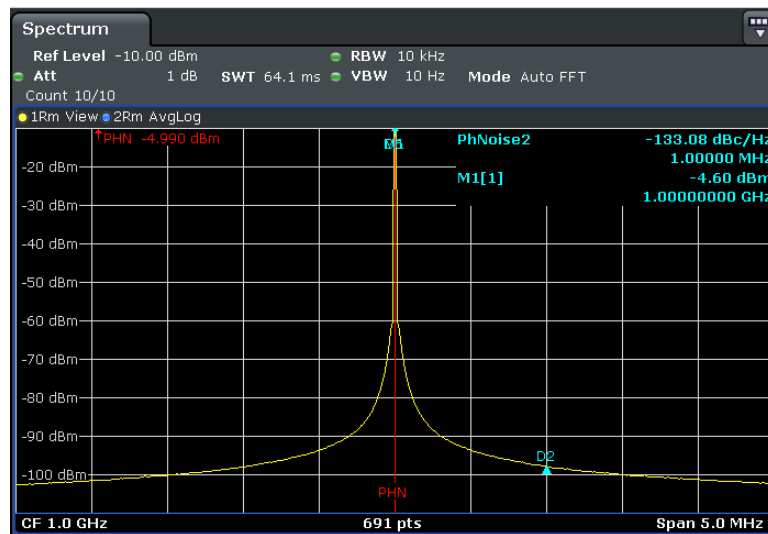
$$\text{Mixer level} = P_{\text{in}} - \text{RF Att}$$

(when no pre-amplifier is used).



- ▶ What is the minimal RF attenuation (*RF Att*) that can be employed for the measurement signal used here without overloading the analyzer input?
- ▶ Reduce the RF attenuation (*RF Att, Manual*) step-by-step until the system reports *Overload*. Then increase the attenuation by one increment.
- ▶ Perform another measurement with the DUT signal (Trace 1), and then without a signal (Trace 2).

Fig. 87 shows a typical measurement result – now optimized for AM noise.



**Fig. 87: Phase-noise measurement with manual setting of the RF attenuation. It is possible to achieve a sufficient dynamic range, because the RF input port is capable of handling overload.**

The result with the phase-noise marker is still displayed correctly here, because the RF path is not overloaded at this position in the spectrum, and because the analyzer had stored the reference value when the phase-noise function was turned on.

The distance from the AM noise in Fig. 87 is now approx. 12 dB. The measured single-sideband phase noise is approx.  $-133$  dBc (1 Hz) at 1 MHz from the carrier.

- ▶ How can you explain the improvement in the display?
- ▶ Is the numeric level display for Marker M1 at the peak still valid?
- ▶ What measurement result is influenced by the overload after the IF filters?
- ▶ Is your measurement result approximately the same size as the analyzer's inherent component?
- ▶ Is it necessary to compensate for AM noise?
- ▶ Assuming that the analyzer's phase noise is  $-135$  dBc (1 Hz), what value can then be calculated for the DUT's phase noise?
- ▶ Set the DUT's frequency to 999.9 MHz and measure again. Discuss the numeric values that are displayed.

## 14.4 Measuring with the Analyzer's Phase-Noise Option

The phase-noise option configures the most important settings automatically.

- ▶ Continue to work with the same DUT settings:

*Frequency*      1 GHz  
*Ref Level*        -5 dBm

- ▶ Configure the following basic settings on the analyzer:

*Reset*  
*Mode*              Phase Noise

You now find yourself in the main menu for the phase-noise option.

- ▶ Specify the general settings:

*Frequency*      1 GHz  
*Level*             -5 dBm

- ▶ Specify this value in the display settings:

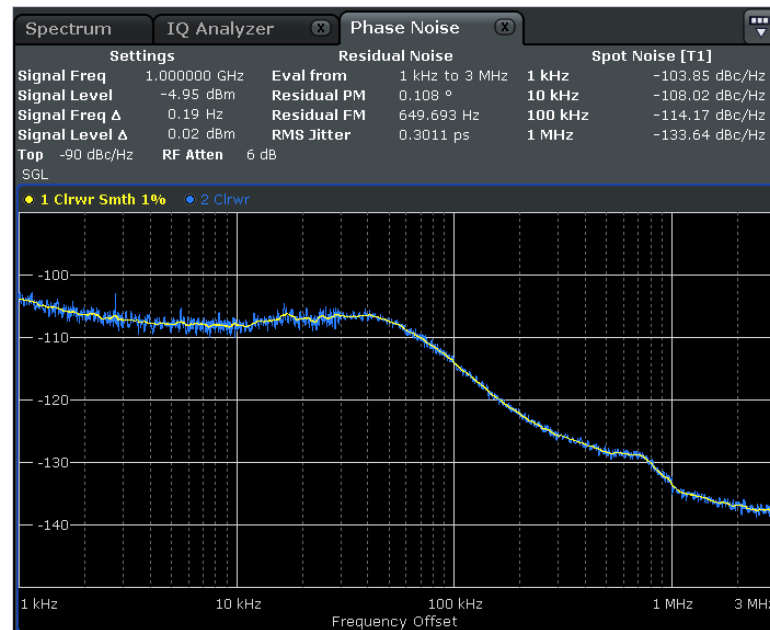
*X Axis Stop*    3 MHz

- ▶ Specify this value in the measurement settings:

*Sweep Mode*    Averaged

- ▶ Apart from the above settings, use the default settings.

- ▶ Start an individual measurement or continuous measurements.



**Fig. 88:** Measuring with the phase-noise option.

Fig. 88 shows typical measurement results. The blue line shows a result without smoothing; the yellow line shows a smoothed trace. In addition to the graphical display, the "spot values" supply numeric values at four (selectable) frequency offsets. Additional information on the residual PM and FM complements the phase-noise results.

The phase-noise option greatly simplifies operation of the device. Most of the necessary settings are preset; for instance, the right detector has been chosen and no logarithmic averaging takes place.

- ▶ Which detector is the right one?
- ▶ How do the powers have to be averaged?

Beyond this, the analyzer compensates for its AM noise internally using the phase-noise option.

- ▶ Are the measurement results shown in Fig. 88 compatible with those in Fig. 87?
- ▶ What conclusion can be drawn based on the measurement results in Fig. 88 for the analyzer's phase noise (is this value  $-135$  dBc (1 Hz))?
- ▶ Do your measurement results match those from the "manual" measurement?
- ▶ Change the individual measurement parameters. Try to shorten the measuring time. How does that affect accuracy?

## 15 Performing Measurements on a Mixer

The four most important measurements for testing a mixer to see if it is suitable for a specific application are:

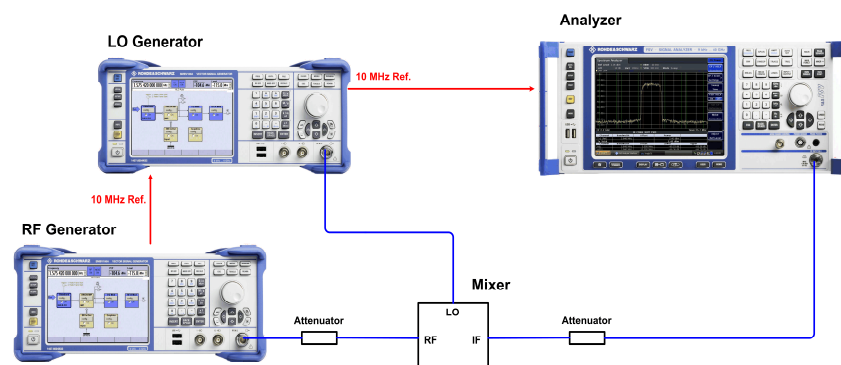
- Measurement of the conversion loss
- Measurement of the backward attenuation (isolation)
- Measurement of the voltage standing wave ratio (VSWR)
- Measurement of the third-order intermodulation products (two-tone third-order distortion)

In the following lab exercise, you will perform the first two of these tests. The goal is to determine the extent to which a certain mixer is suited for down-converting the ISM frequency band (RF) from 2.4 GHz to 2.6 GHz to an intermediate frequency (IF) of 300 MHz. The conversion loss and the isolation will only be measured for this specific application, not for the mixer's entire operating range.

As the example DUT, this description will assume the use of a ZX05-43H+ double-balanced mixer from the company Mini-Circuits ([www.minicircuits.com/pdfs/ZX05-43H+.pdf](http://www.minicircuits.com/pdfs/ZX05-43H+.pdf)). This is a Level-17 mixer, which means that the power at the local oscillator (LO) input should be approximately 17 dBm. For the overall mixer model, the manufacturer specifies a frequency range of 1 GHz to 4 GHz for the RF and LO ports and from DC to 1.5 GHz for the IF port. (If you will be performing the measurements on a different mixer, you might have to change the task to match your device.)

The exercise will begin by enabling you to first gain an overall impression of the mixer's output signals. To do that, and to measure the conversion loss, the following test setup will be used:

### 15.1 Test Setup



**Fig. 89: Setup for measuring the mixer's output signals.**

On the mixer, the RF port will serve as the input and the IF port as the output, because the goal is to convert the ISM band to a lower frequency. For up-conversion, the signal flow would be the other way around.

One generator creates the LO signal, and a second one generates the RF input signal. At its RF and IF ports, the DUT (mixer) is connected via attenuators in order to improve the VSWR somewhat. This is not the case at the LO port, because (with the mixer being used here), this task requires a power level that is close to the generator's maximum power output. The generators and the analyzer are synchronized via the 10 MHz reference in order to ensure stable frequency conditions.

- ▶ Configure the RF generator for internal reference and the other instruments for external reference.

(A setup consisting of a generator and a network analyzer would be comparable to the one used here.)

To measure the conversion loss, the RF and the LO frequency will be changed later incrementally to "run through" the ISM band. The analyzer measures the mixer's output power at the constant intermediate frequency.

**Note:** The analyzer input cannot handle any offset voltage in the DC mode. Ensure that the analyzer is operated in the AC mode. If that isn't possible, insert a DC blocker into the measurement path!

If the LO and RF signals have the same frequency while you are experimenting (for example after a reset of both generators), a DC voltage arises at the mixer's IF output!

There are two ways to achieve frequency conversion:

- With an LO frequency that is below the RF frequency.
  - With an LO frequency that is above the RF frequency.
- ▶ For both cases, calculate the LO frequencies (RF = 2.4 GHz to 2.6 GHz, IF = 300 MHz).
  - ▶ Where is the image frequency range in each case?
  - ▶ Which LO frequency is more favorable for suppressing the image frequency?

### Calibration:

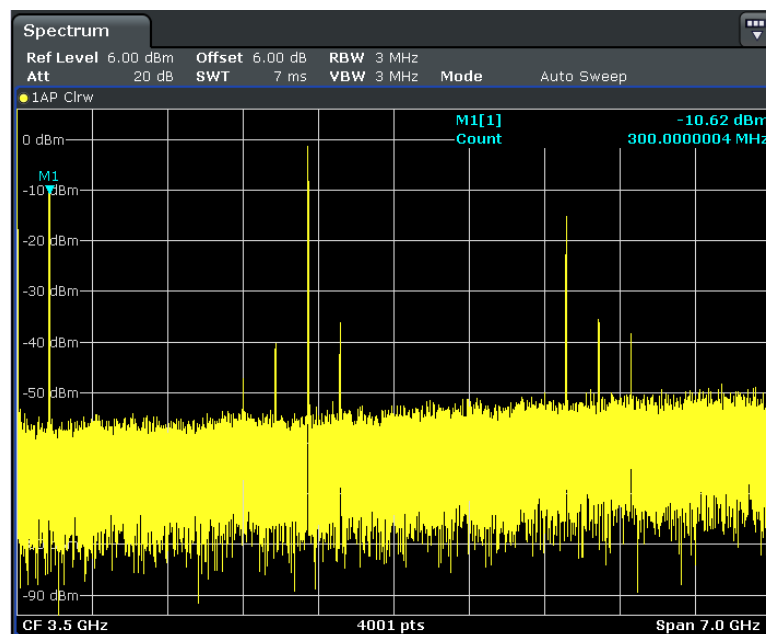
For the task of receiving a 200 MHz band, the conversion loss ripple is the primary factor, not the absolute value. Consequently, calibrating the test setup is not necessary in this case. (Any ripple in the damping from the test setup's attenuators and cables will be negligible within this narrow frequency range.)

### Overview Measurement

At the output of an ideal mixer, only one spectral component appears in the difference frequency and one at the sum frequency for the two input signals. In the next section, examine the real spectrum at the mixer output.

- ▶ Reset the analyzer (*Reset*), and use the maximum span.
- ▶ On the analyzer, compensate for the attenuator in the IF path as the *Ref-Level-Offset*.
- ▶ Set the analyzer's *Ref Level* to 10 dBm.
- ▶ Apply an (unmodulated) signal with 17 dBm / 2700 MHz at the mixer's LO port.
- ▶ At the mixer's RF port, apply an unmodulated signal with 0 dBm / 2400 MHz (compensate for the attenuator in the RF path).

You then obtain a spectrum similar to the one shown in Fig. 90:



**Fig. 90:** Spectrum at the mixer's IF port with input frequencies of 2.4 GHz (RF) and 2.7 GHz (LO).

Besides the desired IF at 300 MHz, you primarily detect a powerful spectral component at the local oscillator's frequency. Since the analyzer's frequency range ends at 7 GHz in this case, higher spectral components cannot be observed, although they are certainly present.

- ▶ Measure the frequencies of the strongest spectral lines for your DUT and explain where they come from.

Is coupling an antenna directly to the mixer's RF input permissible?

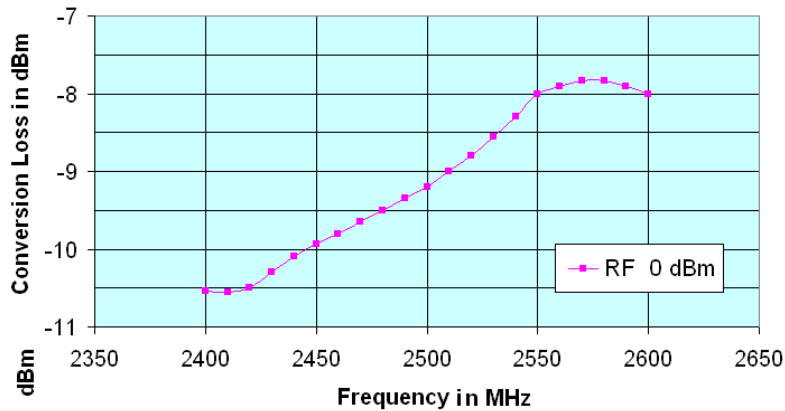
How is it possible to prevent the LO signal from penetrating "into the environment"?

If, in combination with a fixed IF, you increase the LO frequency by 10 MHz, where is the new RF reception frequency?

## 15.2 Measuring Conversion Loss

- ▶ Now set the analyzer's center frequency to 300 MHz and its span to 10 MHz.
- ▶ Set the marker to "peak."
- ▶ Measure the conversion loss across the entire ISM band in steps of 10 MHz.
- ▶ Enter the results in a diagram.

Fig. 91 shows a trace from an example mixer.

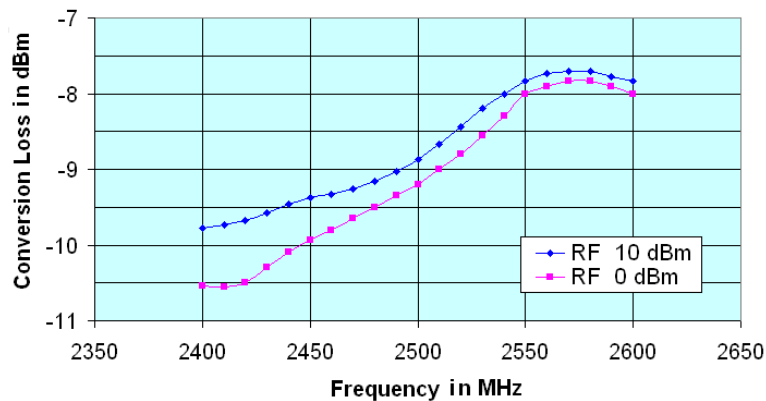


**Fig. 91: A mixer's conversion loss in the ISM band.**

Is this mixer well suited for the required task?

- ▶ Repeat the measurement at an input power of 3 dB below the 1 dB compression point (based on information from the device's data sheet).

Fig. 92 shows both traces from the example mixer.



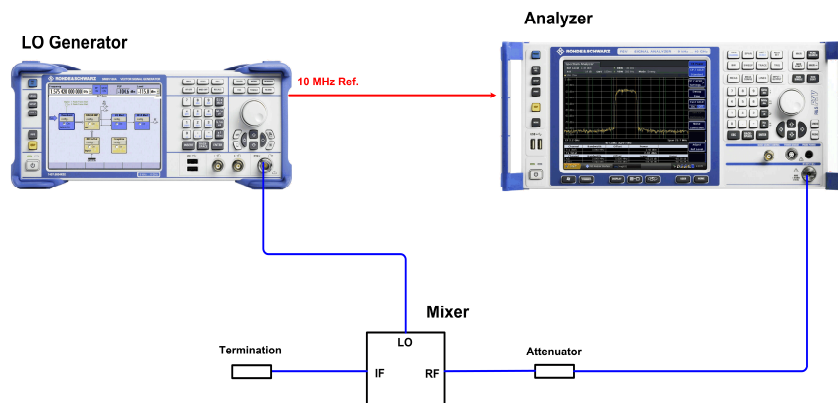
**Fig. 92: Conversion loss for a mixer at different input powers.**

Interpret your own measurement results.

## 15.3 Measuring the LO-RF Isolation

To measure the attenuation between the LO and another port, the third unused port is fitted with an RF terminator. The measurement is taken using the LO's operating power level.

- ▶ Modify the test setup to match Fig. 93.



**Fig. 93: Setup for measuring LO-RF isolation.**

Note that the LO generator must be operated with an internal 10 MHz reference in this case.

Which LO frequencies were needed in the experiments performed up until now for reception on the lower end of the ISM band? In the middle of the band? In the upper end of the band?

- ▶ Reset the analyzer (*Reset*), and use the maximum span.
- ▶ On the analyzer, compensate for the attenuator in the IF path as the *Ref-Level-Offset*.
- ▶ Set the analyzer's *Ref Level* to 10 dBm.
- ▶ At the mixer's LO port, apply an unmodulated signal with 17 dBm at the LO frequency for receiving at the lower end of the band.

What spectral components do you observe?

- ▶ Note your measurement results and repeat the measurement at LO frequencies for the middle and upper end of the band.

Is the LO-RF isolation constant?

- ▶ Check to see if (at a certain frequency) the harmonics arise as the result of analyzer overload.
- ▶ Replace the mixer with a 20 dB attenuator. Examine the generator's harmonics.
- ▶ In your own words, summarize the knowledge gained from this experiment.



## 15.4 Measurement Software

The following task is only to be performed if you are working with Rohde & Schwarz instruments!

To support frequency response measurements performed with its generators and analyzers, Rohde & Schwarz makes the 1MA09 *FreRes* (Program for Frequency Response Measurements) application note available. This application note is available as a free download at: <http://www.rohde-schwarz.com/appnote/1MA09>.

*FreRes* controls frequency response measurements remotely. The instruments are addressed via GPIB or LAN.

First return the measurement setup to the configuration used for measuring conversion loss, see ► Fig. 94.

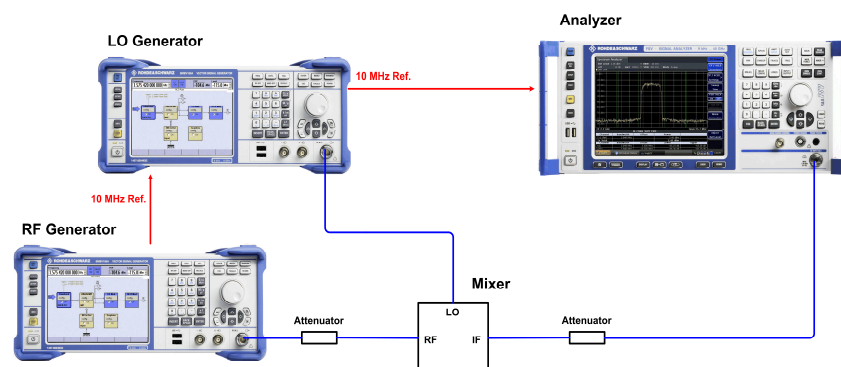


Fig. 94: Setup for measuring the mixer's output signals.

- Connect the remote control cabling between the instruments and your test computer (LAN or GPIB).
- Download the *FreRes* program and install it on your test computer. Operating this software also requires a VISA library; see <http://www.ni.com/VISA>.
- Start the program.
- First specify these *Devices*: *Generator1* and *Analyzer*.

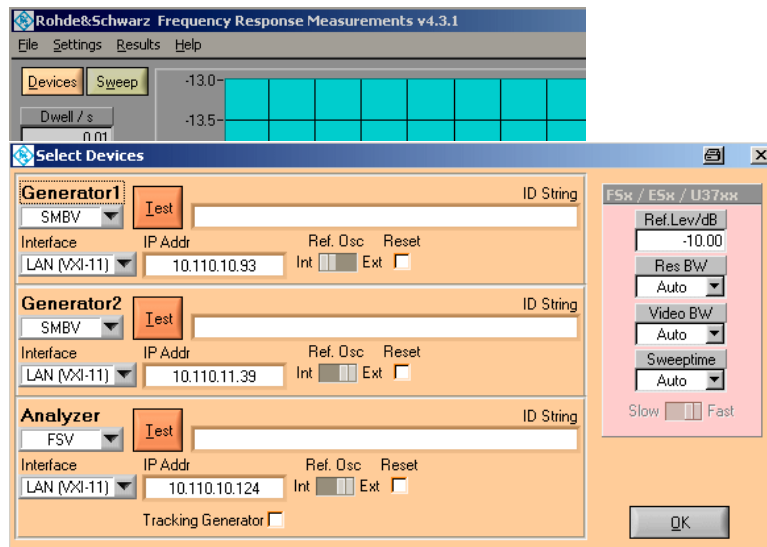


Fig. 95: Entering the devices' remote control parameters.

FreRes can test frequency-conversion modules. Doing that requires two generators. Generator1 supplies the measurement signal, and Generator2 supplies the LO signal.

- ▶ Specify these *Devices*: *Generator1*, *Generator2* and *Analyzer*.
- ▶ Do not activate the three *Reset* check boxes. That keeps a *Reset* from causing both generators to set to the same (default) frequency. Furthermore, this ensures that some of the additional settings that have been configured, such as a *Ref Level Offset*, remain intact.
- ▶ Click OK.

- ▶ Then specify the *Sweep* settings; see Fig. 96.

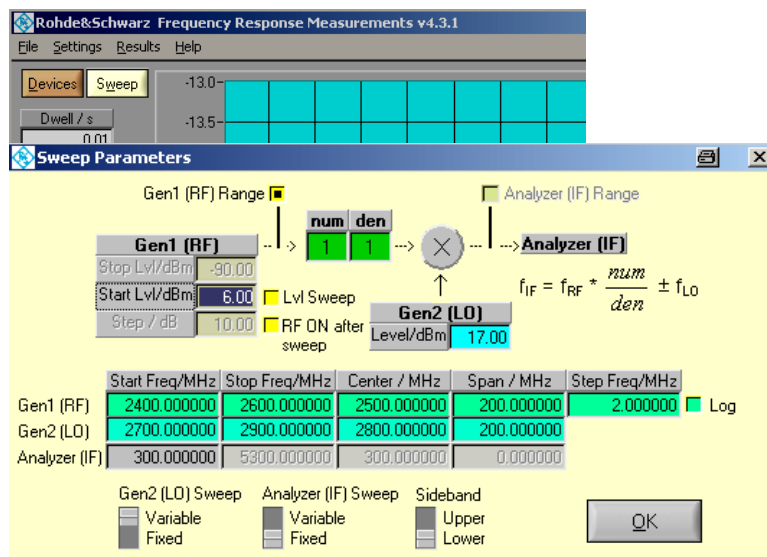


Fig. 96: Entering the sweep parameters.

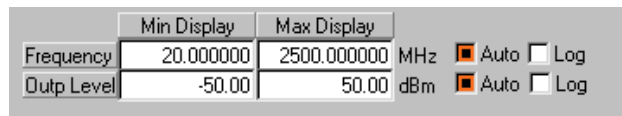
- ▶ Use the checkboxes to select the *Gen1 (RF) Range* or *Analyzer (IF) Range* to establish whether the input frequencies or the output frequencies should be used for the axis labeling.

The measurement is to be taken with a 0 dBm RF input signal. Due to the 6 dB attenuator on the generator:

- ▶ For *Gen1 (RF)*, specify a *Start Lvl* of 6 dBm.
- ▶ Ensure that the *Lvl Sweep* is not activated.
- ▶ For *Gen1*, specify the *Start* and *Stop Frequencies* and the *Frequency Step*.
- ▶ For *Gen2 (LO)*, specify the *Start Lvl (17 dBm)*, as well as the *Start* and *Stop Frequency*.
- ▶ Set the *Gen2 (LO)* sweep switch to *Variable*, the *Analyzer (IF)* switch to *Fixed* and the *Sideband* switch to *Lower*.
- ▶ Click OK.

Now you are back at the main menu.

- ▶ There, first activate the auto-scale functions for the output window:



**Fig. 97: Auto-scaling for both axes shows the overall curve.**

The values shown in the *Frequency* and *Outp. Level* windows now have no significance. No calibration measurement (normalization) is needed, because the focus is one the ripple in the ISM band in this case, and the cables and attenuators can be considered nearly ideal in this narrow range.

Use the following buttons to control the recording:



**Fig. 98: Buttons for controlling the recording.**

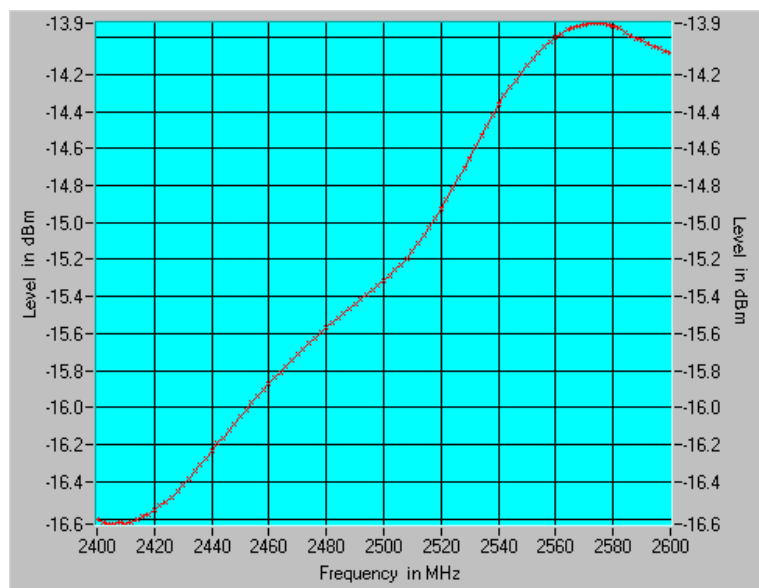
The *Start* button **deletes all normalizations that have been performed (calibration)**, clears the screen content and starts a new measurement.

By selecting *Repeat*, you can start additional sweeps (for example with different values for the *Start Lv*). Unlike the procedure used when the *Start* button is selected, any normalization that has been performed remains in effect when this button is chosen.

You can stop a running measurement by clicking *Stop Meas*, and you can delete the most recent trace with *Del Last Trace*.

► Click *Start*.

Fig. 99 shows the measurement result for the sample mixer with 0 dBm at the input.



**Fig. 99: Conversion loss in the ISM band.**

Did you obtain similar results?  
How large is the conversion loss?

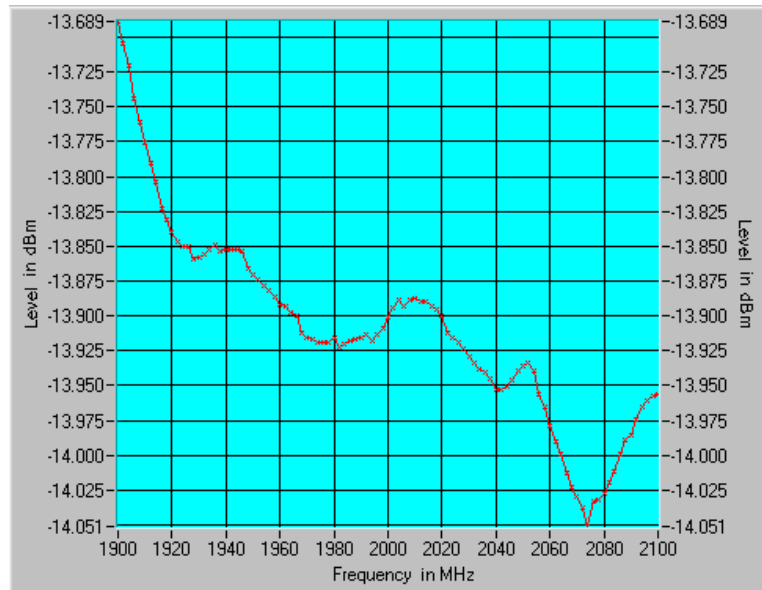
Do these results match the results from the manual measurement (Fig. 91)?  
Explain the difference.

Is the mixer suitable for the task of down-converting the ISM band to an IF of 300 MHz?

► Select other start and stop frequencies in order to find a suitable area of application for the mixer.

Fig. 100 shows measurement results for the example mixer in a different frequency range.

Could this mixer be used in this frequency range?



**Fig. 100: Conversion loss in a different frequency range.**

- Summarize the goals, execution and results of this experiment in your own words.

## 16 Reference

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## 17 Additional Information

This educational note is updated from time to time. To download the latest version,  
please visit:

<http://www.rohde-schwarz.com/appnote/1MA201>.

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