Modulation and Signal Generation with R&S® Signal Generators
Educational Note

Products:

- R&S®SMB100A
- R&S®SMBV100A
- R&S®SMC
- HM8134-3/HM8135
- HMF2525 / HMF2550

Signal generators play a vital role in electrical test and measurement. They generate the test signals that are applied to components such as filters and amplifier, or even to entire modules, in order to ascertain and test their behavior and characteristics. The first part of this educational note presents the applications for and the most important types of signal generators. This is followed by a description of the construction and functioning of analog and vector signal generators. To permit a better understanding of the specifications found on data sheets, a closer look at the most important parameters for a signal generator is provided.

Beyond the output of spectrally pure signals, a key function of RF signal generators is the generation of analog- and digitally modulated signals. The second part of this note therefore presents the fundamentals for the most significant analog and digital modulation methods.

The third part of this educational note contains exercises on the topics of analog and digital modulation. All described measurements were performed using the R&S®SMBV100A vector signal generator and the R&S®FSV spectrum analyzer.

Note:

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Educational Note

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Contents

1 What Is a Signal Generator?

A signal generator generates electrical signals with a specific time characteristic. Depending on the type of signal generator, the generated signal waveforms can range from simple sine-wave, sawtooth and square-wave to analog-modulated signals such as AM, FM and PM and even complex digitally modulated signals such as those used in mobile radio (GSM, UMTS, LTE, etc.). The frequency domain can extend from several kHz into the two-digit GHz range. By adding external frequency multipliers, it is even possible to obtain signal frequencies up to several hundred GHz. The frequency of the output signal can typically be set in very small steps (< 1 Hz).

The RF generators used in the automated test setups for production can be remotely controlled via a LAN connection, a USB port or a GPIB port, depending on the available equipment.

RF signal generators are classified into two main types:

- Analog signal generators and
- Vector signal generators

1.1 Why Do We Need Signal Generators?

Signal generators are primarily used in the development and production of electronic modules and components. The signal produced by the signal generator is fed to an RF module (amplifier, filter, etc.) under test. The output signal from the module is then analyzed using an appropriate test instrument such as a spectrum or signal analyzer, oscilloscope, power meter, etc. (Fig. 1-1). Based on the results of this analysis, it is possible to determine whether the module exhibits the expected behavior. In addition to the typical settings for frequency, amplitude and modulation mode, state-of-the-art signal generators also offer the ability to add noise to the test signal or to simulate multipath propagation (fading) of the input signal. This makes it possible to study the behavior of a receiver with very noisy receive signals that also reach the receiver via multipath propagation.

Although signal generators are not a test instrument in the strictest sense, they are still considered to be test transmitters because of the use case described above.

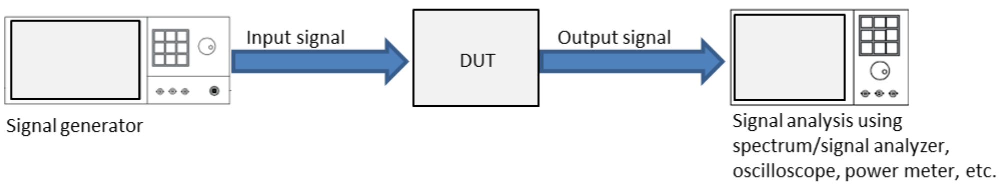


Fig. 1-1: The signal generator supplies the test signal for the functional test of the connected DUT.

1.2 What Types of Signal Generators Are Available?

The selection of the correct signal generator is always dependent on the application. Important criteria include frequency domain, level range, spectral purity, available modulations (analog, digital) and functionality for adding specific interference to signals (noise, simulated multipath propagation).

Simple signal generators for the low-frequency range are called **tone generators**. Typically, a tone generator produces a sine-wave, square-wave or sawtooth electrical signal in the human-audible frequency range, i.e. 20 Hz to 20,000 Hz. The output signal can be set manually both in frequency and in amplitude. Tone frequency generators are most often used for acoustical test and measurement and for electro-acoustics in combination with an appropriate audio level meter. This makes it possible to determine whether the generated test tone passes through a tone transmitter system at the desired level, for example. For sweep measurements, a suitable level meter (e.g. oscilloscope, spectrum analyzer, power meter) can additionally be used to acquire the full frequency response of a transmission system by continually shifting the output frequency or by using a spectrum analyzer to supply the frequency.

Signal generators that produce only simple, periodic signals are called **function generators**. This typically includes sine-wave, square-wave and triangle oscillations. Today's function generators are overwhelmingly digital. These operate using direct digital synthesis (DDS) (see also section 1.2.1.1) and can produce a variety of periodic signal traces. Example function generators applications include the assessment of electronic circuits such as amplifiers and filters.

HF-signal generators used to produce high-frequency sine-wave oscillations. These often include a sweep function that permits repetitive sweeps through an adjustable frequency domain. The frequency domain can extend from several kHz into the two-digit GHz range. RF generators are categorized as either **analog or vector signal generators**. Analog signal generators permit analog modulation of the output signal in frequency and amplitude. They are also capable of producing pulsed signals. Vector signal generators are additionally capable of producing digitally modulated signals for a variety of standards from mobile radio, digital radio and TV and so on. Like analog signal generators, vector signal generators also modulate a high-frequency carrier signal.

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. The output signal is always a baseband signal. The advantage of these generators is that they can produce and repeat almost any conceivable waveform as often as necessary. ARBs are used as a universal signal source in development, research, test and service.

The following section of this educational note provides a closer look at RF signal generators (analog, vector-modulated) and at arbitrary waveform generators.

1.2.1 Analog Signal Generators

With analog signal generators, the focus is on producing a high-quality RF signal. They provide support for the analog modulation modes AM / FM and φM. Many instruments can also produce precise pulsed signals with varying characteristics.

Analog generators are available for frequencies extending up to the microwave range. In addition to a frequency sweep over an adjustable frequency range, a few instruments also can do a level sweep through the output levels within predefined limits. This makes it possible to produce some of the same waveforms as a function generator, such as sawtooth or triangle.
As mentioned, levels with a previously defined pulse/pause ratio can output a repetitive sequence of square-wave amplitude characteristics. Analog generators typically exhibit the following characteristics:

- Very high spectral purity (nonharmonics), e.g. –100 dBc
- Very low inherent broadband noise, e.g. –160 dBc
- Very low SSB phase noise of typ. –140 dBc/Hz (at 10 kHz carrier spacing, f = 1 GHz, 1 Hz measurement bandwidth)

Analog signal generators are used:

- As a stable reference signal for use as a local oscillator, i.e. as a source for phase noise measurements 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 electromagnetic compatibility (EMC) testing
- As automatic test equipment (ATE) during production
- For avionics applications (such as VOR, ILS)
- For radar tests
- For military applications

Fig. 1-2 shows an example of a special impulse sequence for radar applications:

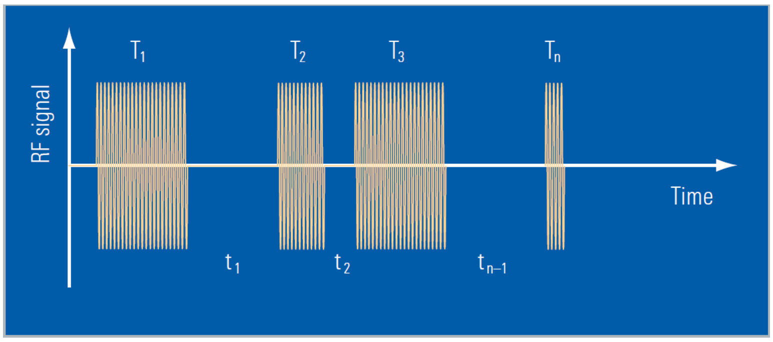


Fig. 1-2: Combination of impulses with different widths and interpulse periods for radar applications.

Analog signal generators are available with different specifications in all price classes. The criteria listed in section 1.3 Key Signal Generator Characteristics can be used to select an appropriate model. Some typical characteristics include a specific output power level, rapid frequency and level settling, a defined degree of level and frequency accuracy, a low VSWR or even the design and weight of the generator.

1.2.1.1 Design and functioning of an analog signal generator

Fig. 1-3 shows the basic design of an analog RF signal generator.
A signal generator consists of three main functional blocks:

- A synthesizer for producing the oscillation
- An automatic level control for maintaining a constant output level
- An output stage with amplifiers and step attenuators for controlling the output power

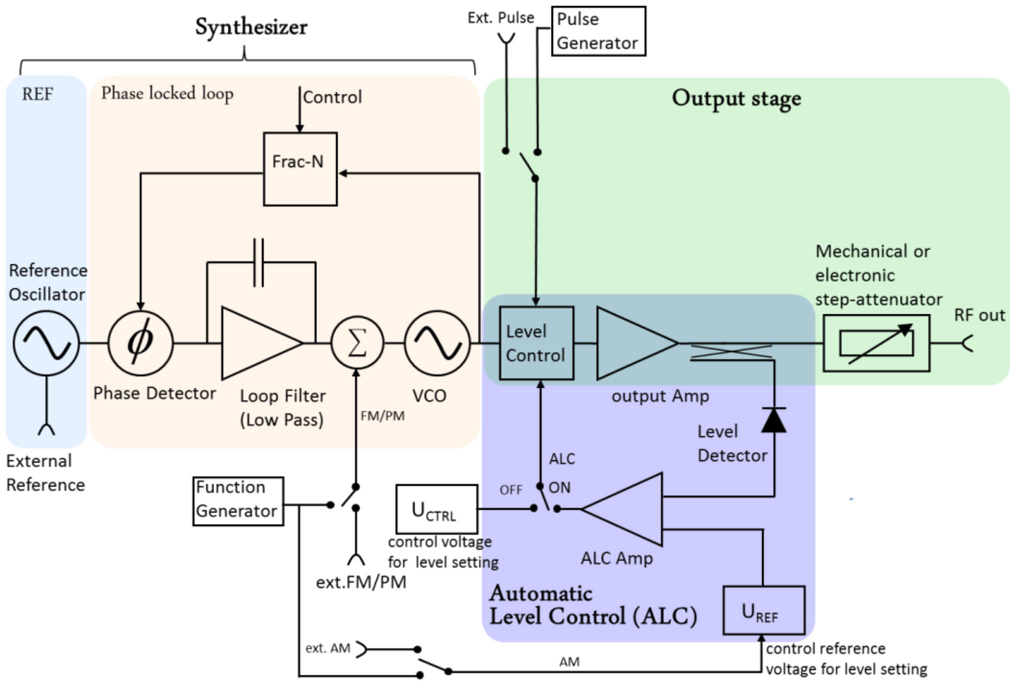


Fig. 1-3: Schematic diagram of an analog RF signal generator.

Synthesizer:

The core of the generator is the synthesizer used to produce the oscillations. The term synthesizer describes a functional principle in which oscillations are synthetically derived by means of division and multiplication from the oscillations of a crystal with a fixed reference frequency (reference oscillator). Usually a signal generator can also be synchronized to an applied reference frequency via a reference frequency input. This permits synchronization with other test instruments, such as a spectrum analyzer. Because a frequency from a crystal provides long-term stability and low dependence on temperature, crystal oscillators are used to generate the reference frequency. The quality of an oscillator is typically assessed based on the stability of its amplitude, frequency and phase. The stability can be increased further by using a temperature-controlled crystal oscillator (TCXO). This consists of a voltage-controlled oscillator (VCO) and a compensation circuit that compensates for the oscillator's temperature dependency. The frequency accuracy of these oscillators is under 100 ppm (ppm:

parts per million, i.e.10^{-6}), making them up to ten thousand times more accurate than an LC resonant circuit built from discrete components. Nevertheless, a TCXO always retains a small thermal frequency drift that is disruptive for precision applications. To further improve the frequency accuracy, most signal generators also offer an oven-controlled crystal oscillator (OCXO) as an option. In an OCXO, the crystal oscillator is located in a heated, temperature-controlled chamber. By regulating the temperature of the oscillating crystal and the oscillator circuit to a value above room temperature, it is possible to further stabilize the oscillation frequency, providing a higher degree of accuracy than without heating. Depending on the type of oscillator used and the application, the operating temperature lies in the range from +30 °C to +85 °C. This provides a level of accuracy a thousand times better than that of a TCXO, in the range of 0.001 ppm, or 10^{-9}. The frequency drift caused by aging of the oscillator is also improved. In this case, the improvement is a factor of 100, from 10^{-6} / year to 10^{-8} / year. Because aging of the reference oscillator is unavoidable, a regular adjustment of the reference frequency might be necessary in order to ensure a high absolute accuracy, depending on the application.

Frequency multiplication is typically managed by using a phase-locked loop (PLL).The basic concept behind a PLL is illustrated in Fig. 1-4.

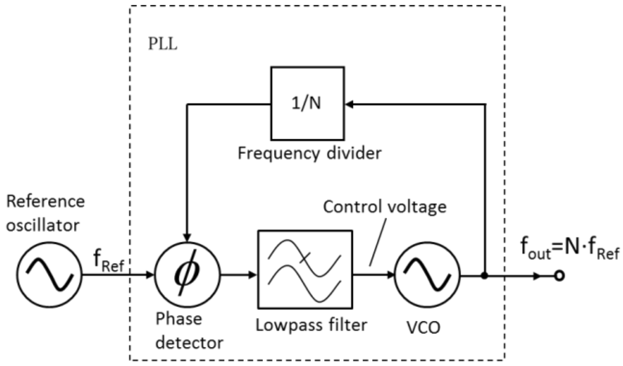


Fig. 1-4: PLL block diagram.

The circuit is structured as a locked loop. It consists of three functional groups, a phase detector for the phase comparison, a loop filter with lowpass characteristics for filtering high-frequency signal components out of the control voltage from the phase detector and a voltage-controlled oscillator (VCO). The input and output values for a PLL represent the frequency. In the locked loop, a phase detector compares the phase angle of a voltage-controlled oscillator (VCO) against the reference phase angle of the reference oscillator. The difference is used to obtain a control voltage, which in turn is used to adjust the phase angle of the oscillator several times a second. The PLL principle is used for the frequency multiplication in a signal generator. This is done by dividing the VCO output frequency by the integer divider N using a digital programmable frequency divider and then applying the frequency to the phase detector via the feedback path (integer-N PLL). In a locked PLL, the output frequency f_{out} is then exactly N-times the input frequency f_{in} of the reference oscillator. The greatest advantage to this method is that the generated frequency can be incrementally varied over a large frequency range while still retaining the same stability and accuracy of the reference crystal oscillator.

The disadvantage is that as a result of the integer frequency multiplication, the step size for the frequency resolution is equal to the reference frequency. To implement the smallest frequency resolution steps possible, the frequency of the reference oscillator can be divided down accordingly before the PLL by using a frequency divider. In order to produce high frequencies, the resulting small steps make it necessary to select an appropriate lager dividing factor N (works as a multiplier at the output of the PLL). This in turn negatively affects the phase noise (see also section 1.3.1) of the output signal, which increases by 20 log (N) dB. In the case of a signal generator with a step size of 100 kHz and a maximum generator frequency of 1 GHz, the phase noise up to 1 GHz increases by 20 log (10000) dB = 80 dB. Lower reference frequencies additionally require a narrow control bandwidth. The lock time depends primarily on the control bandwidth. The smaller this is, the slower the PLL becomes.

In order to counteract the above-mentioned disadvantages of lager dividing factors and the associated worsening of the phase noise, synthesizers can be equipped with multiple PLLs as shown in Fig. 1-5.

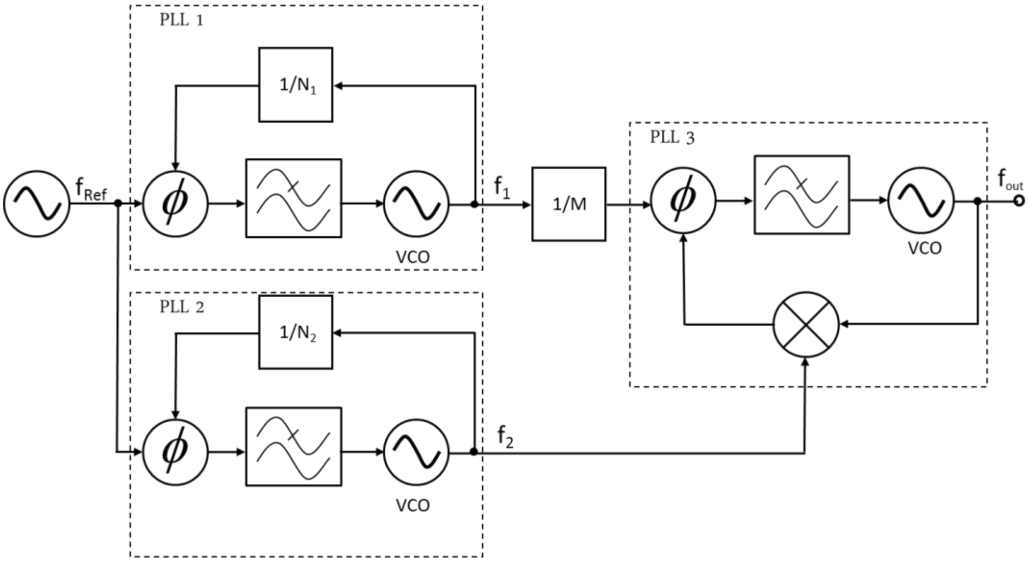


Fig. 1-5: Multiloop synthesizer (multi-PLL).

To implement fine frequency steps, a frequency divider is placed in the upper signal path at the output of the first PLL (PLL_1). Larger frequency steps are generated in the lower path by means of a second PLL (PLL_2). At the output of yet another PLL (PLL_3), which contains a mixer in the feedback path instead of a frequency divider, the output frequency with the dividing factors N_1, N_2 and M is described by the following equation:

Equation 1-1:

$$f_{out} = f_{ref} \cdot \left(\frac{N_1}{M} + N_2\right)$$

For example, if $N_1 = 100$, $M = 100$ and $N_2 = 99$, at a reference frequency $f_{ref} = 10$ MHz the above equation generates an output frequency of 1 GHz ($f_{out} = 1$ GHz). The minimum step size is 100 kHz. In comparison to the simple PLL shown in Fig. 1-4, the significantly smaller divider values result in a lower phase noise for the output signal. In addition, from Equation 1-1 it becomes clear that with a multi-PLL, the appropriate selection of N_1 and M can also result in fractional factors for the frequency multiplication, and therefore very small frequency steps. The disadvantage of a multi-

PLL is that a stable output signal requires that all PLLs be locked, which makes extremely short frequency settling times impossible.

In summary, when dimensioning an integer-N PLL, a high frequency resolution stands in opposition to the requirements for spectral purity and short lock times. A multi-PLL delivers small step sizes with low phase noise. However, the frequency settling time is not adequate for many applications, such as production environments.

In order to achieve a faster settling time, modern signal generators therefore add a frequency divider to the feedback path of a PLL, which divides the VCO frequency using fractional factors (see Fig. 1-3).

To do this, the dividing factors of various integer values are varied over time so that they average value is equal to the desired fractional value. For example, to achieve an divider of 100.5, one half of the time is divided by 100 and the remaining time by 101. This averages out to the desired value of 100.5. Control loops that operate on this principle are called fractional-N PLLs or fractional-N synthesizers. However, fractional spurs are generated at the output of the phase detector that must be compensated for or filtered out by using appropriate countermeasures (Delta-Sigma method). Fractional-N synthesizers can provide infinitely fine frequency resolutions with simultaneously short settling times and very high spectral purity. For further reduction of the phase noise, fractional-N synthesizers are also implemented as multiloop fractional-N synthesizers following the principle shown in Fig. 1-5.

Particularly low phase noise synthesizers are implemented based on the direct digital synthesis (DDS) concept. This concept has its foundation in a method taken from digital signal processing for the generation of periodic, band-limited signals with infinitely fine frequency resolution. Next to PLL, DDS is the most important method today for generating signals with finely adjustable frequencies into the mHz range, and as such has found wide usage in the test and measurement field.

Because D/A converters are not yet available for generating high frequencies, the DDS concept is limited to clock frequencies up to around 1 GHz. In signal generators (function generators) for the frequency range up to several hundred MHz, DDS is often the only signal generation method in use. To utilize the advantages offered by PLL synthesizers and DSS, signal generators for higher frequencies often use a combination of the two methods.

Fig. 1-6 shows the basic setup for DDS.

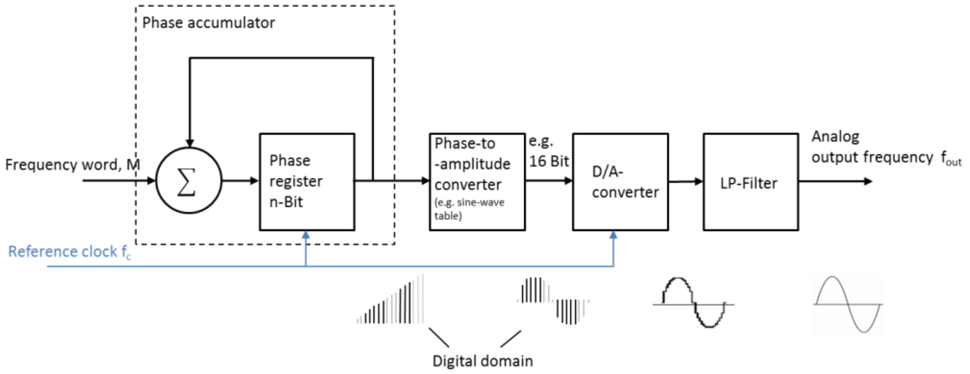
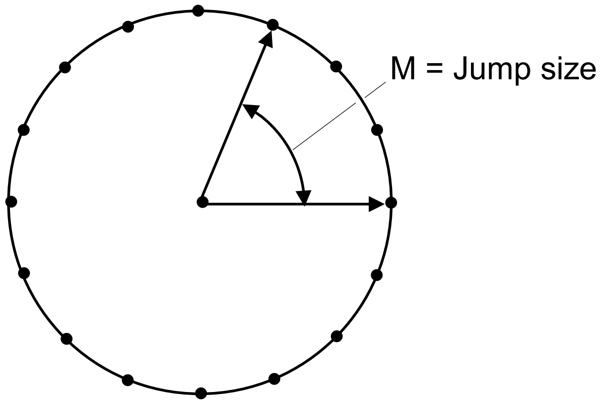


Fig. 1-6: DDS block diagram.

DDS is based on a phase accumulator that consists of a digital accumulator combined with a register, including a feedback path. The register functions as storage for the current phase angle. For each clock step, this phase accumulator adds the applied frequency word via the feedback path and the phase register, and thus functions as a counter. The digital frequency word, at a length ranging from 24 bits to 48 bits, is equivalent to the desired output frequency. The instantaneous count corresponds to a specific phase angle. The phase accumulator can also be viewed as a digital phase angle wheel as shown in Fig. 1-7, divided into n equal parts. The phase angle wheel includes a vector that is rotated by a defined number of points or angle units with every pulse of the reference clock. The output frequency is changed by the variation of frequency word M, which can also be interpreted as step sizes. At a vector step size of M = 1 and a 32 bit wide phase accumulator, 2^{32} clock pulses are necessary for the phase vector to cover 360°. Once the phase accumulator overflows, a new cycle is started at phase angle 0°. If M = 2, the vector skips every other phase point and therefore rotates twice as fast, doubling the output frequency. The frequency resolution relative to the clock frequency that can be achieved with DDS depends entirely on the word width of the phase accumulator. A typical word width is 32 bits, which at 200 MHz system clock is equal to a frequency resolution of about 0.046 Hz at a maximum output frequency of theoretically one-half the clock frequency, so 100 MHz.



n	Number of points 2^n
8	256
12	4096
16	65536
20	1048576
24	16777216
28	268435456
32	4294967296
48	281474976710656

Fig. 1-7: Digital phase angle wheel.

The digital output value of the phase accumulator is then sent to a phase-to-amplitude converter, which converts the phase information based on the waveform saved there into an amplitude (Fig. 1-8).

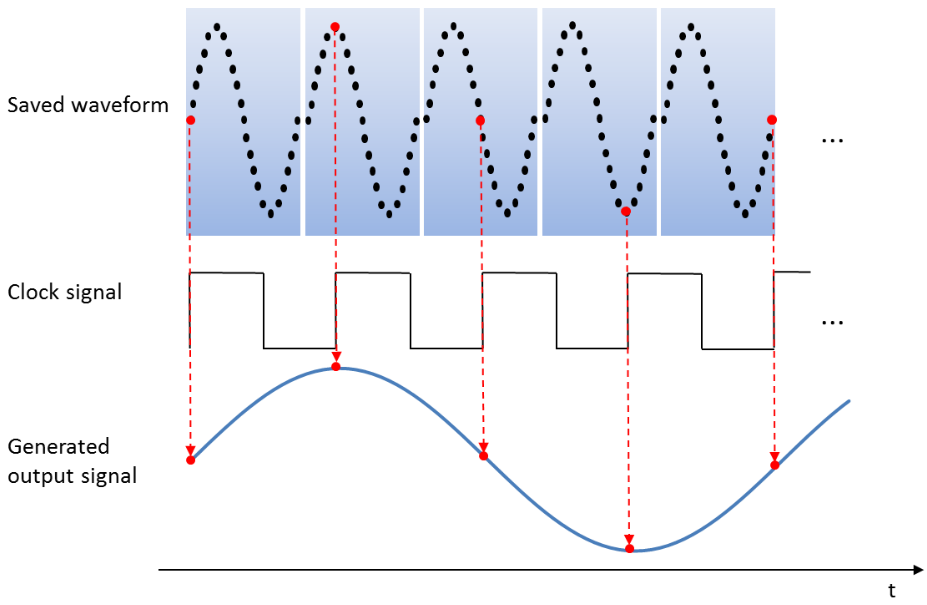


Fig. 1-8:

The frequency produced in this manner is calculated as follows:

Equation 1-2:

$$f_0 = \frac{M \cdot f_c}{2^n}$$

where:

M: Frequency word (step size)

f_c: Clock frequency

2^n: Number of points in the saved waveform

The maximum possible frequency, taking the sampling theorem into account, is

$f_0 = \frac{f_c}{2}$. As a result, the condition $M \leq 2^n/2$ must be observed.

After the phase/amplitude converter, a D/A converter converts the sequential digital data values to an analog, step-shaped signal, which is then formed into a sine-wave signal using a lowpass filter.

The spectral purity of the output signal is primarily dependent on the D/A converter. The accuracy of the generated frequency depends on the quality of the clock signal and the resolution (number of points in the quantized sine-wave oscillation). Because DDS does not require a settling time for a loop control, this concept in combination with the D/A converter offers clock frequencies with very fast frequency settling times into the ns range.

Fig. 1-9 shows the phase noise characteristic for a signal generator with and without DDS. In the lower frequency range to about 250 MHz (blue curve), the use of DDS (here in comparison to the conventional synthesizer concept using PLL) provides significantly lower phase noise.

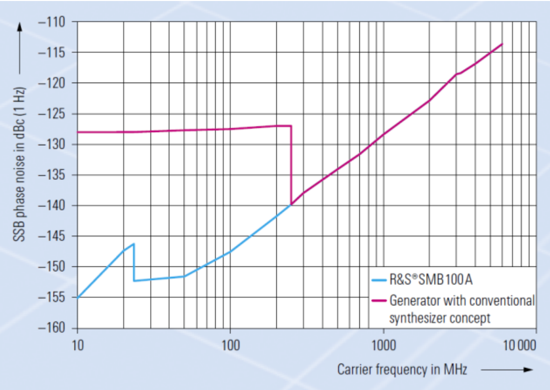


Fig. 1-9: Phase noise with and without DDS.

In addition to a sine-wave signal, DDS can also relatively easily be used to generate frequently required signal waveforms. Because the phase within each period increases linearly and then jumps back to zero, the phase value can be directly output as a signal value in order to generate a sawtooth signal. A triangular signal can be generated just as easily. In this case, the phase value is interpreted as a signed binary number, after which the absolute value is calculated. DDS is therefore frequently used in digital function generators.

DDS can very easily be implemented in a field programmable gate array (FPGA), in a digital signal processor (DSP) or even in an application-specific integrated circuit (ASIC). By adding appropriate functional enhancements to DDS, it is easy to implement FM/PM modulation, or amplitude modulation, of the output signal. DDS circuits with an integrated D/A converter are currently available with clock frequencies of higher than 1 GHz.

Independent of the synthesizer types described above, it is also possible to use a signal generator as a sweep generator by adding a generator control. In other words, the output signal transverses a frequency band defined by a start and stop frequency at the defined frequency resolution. Together with a spectrum analyzer, the signal generator can be used as a tracking generator, for example for measuring the filter transfer function. Many signal generators and analyzers offer corresponding functionality for synchronization, frequency definition, etc. via a remote control interface.

In addition to the frequency accuracy and the quality of the generated signal, the level accuracy and the size of the adjustable level range are important characteristics of a signal generator. As shown in Fig. 1-3, the signal generated by the synthesizer is applied over the output stage to the output jack on the signal generator. The output stage essentially functions as a level control. It therefore includes an automatic level control (ALC) as well as step attenuators and, depending on the generator model, also a power amplifier for generating high power levels.

ALC:

The automatic level control (ALC) is located in the output stage before the step attenuator. Fig. 1-10 shows the setup and functioning. The ALC has three tasks:

- Level setting at small step sizes (up to 0.01 dB)

- Constant maintenance of the level over temperature and time
- Amplitude modulation (AM) through changes to the input variable

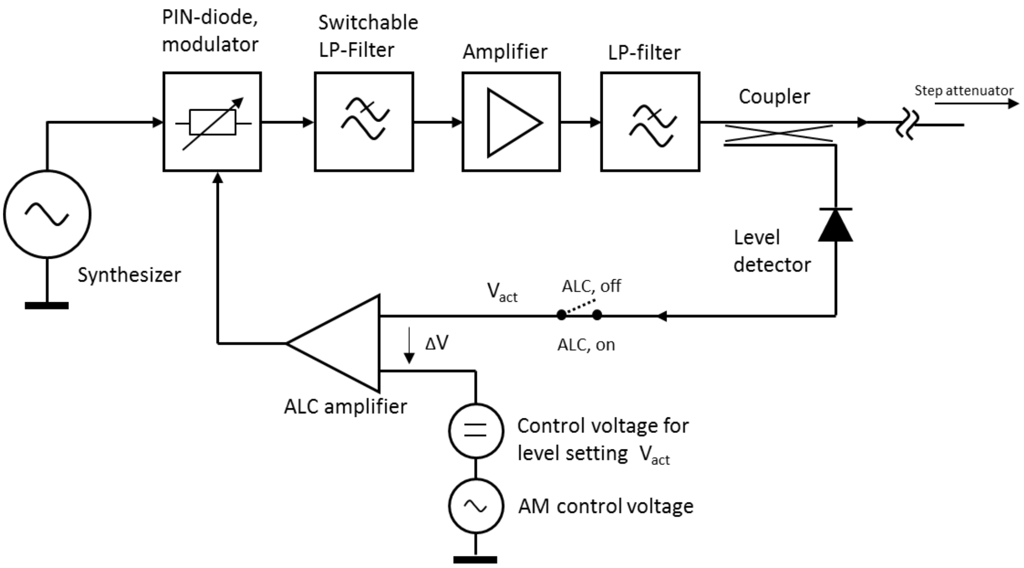


Fig. 1-10: Automatic level control (ALC) setup.

For the level control the synthesizer signal is supplied to a pin-diode modulator. The PIN diode represents a DC-controlled resistance for higher frequencies that is used to control the signal amplitude. The subsequent switchable lowpass filters serve to suppress harmonics. The following amplifier compensates the insertion loss of the modulator. Harmonics that lie above the maximum output frequency of the synthesizer are subsequently suppressed with a lowpass filter. In the case of an ALC, a control loop constantly maintains the output signal at a defined value. The control is implemented by using a directional coupler with a known coupling loss behavior to uncouple a portion of the output signal, which is then applied to a detector diode. An ALC amplifier in the feedback path compares the reference voltage V_{ref} that is equal to the defined level against the actual voltage V_{act}. From the difference between the two voltages, the correction DC voltage is derived in order to control the PIN diode. An ALC makes a very precise electronic level attenuation up to 40 dB possible, for example from –15 dBm to +25 dBm, along with small step sizes down to 0.01 dB. With an ALC, a change in the control voltage (V_{ref}) based on the wanted signal can very easily generate an amplitude-modulated oscillation.

For most applications, the ALC is typically switched on. In the case of pulse-modulated signals, however, an ALC would lead to level variations in the output pulse sequence. Therefore, signal generators with ALC include the option of switching it off. Another set of applications in which ALC negatively affects the measurements are multitone measurements with two signal generators. Fig. 1-11 shows a multitone measurement for determining the third harmonic intercept point (IP3) of an amplifier. Two signals of equal level are sent to the amplifier with different frequencies f_1 and f_2 via an RF combiner. The output spectrum of the amplifier is determined using a spectrum analyzer. From the third intermodulation ratio a_{IM3} (level offset between the first harmonic and the third intermodulation product), it is possible to calculate the IP3.

Many spectrum analyzers use a special measurement function that calculates and displays the IP3.

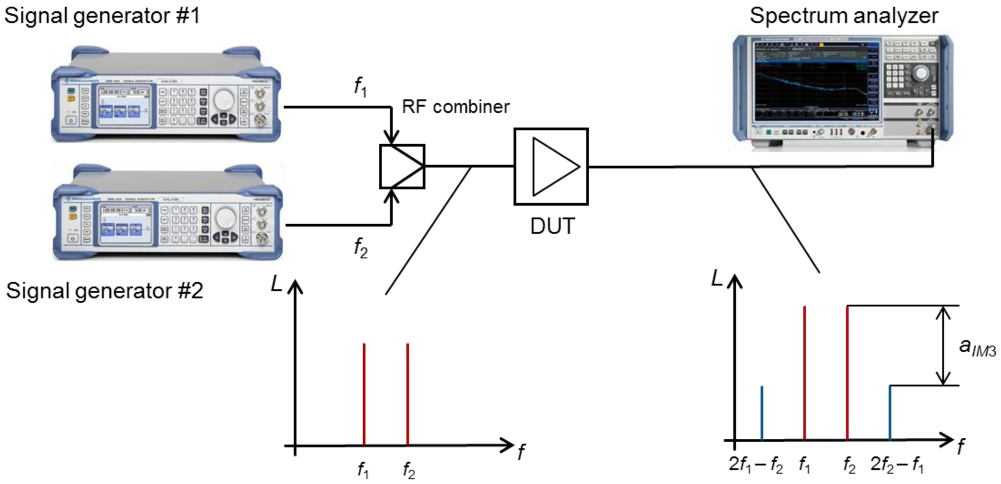


Fig. 1-11: Test setup for an IP3 measurement on an amplifier.

Fig. 1-12 shows the influence of the ALC on the measurement result during a multitone measurement. Although no DUT is included in the test setup, the output of the RF combiner shows not only the first harmonics with the frequencies f_1 and f_2 but also intermodulation products. The reason for this lies in the limited isolation of the RF combiner as well as the limited directivity of the directional coupler in the ALC. As long as the spacing between the two generator signals lies within the control bandwidth of the ALC, a portion of the output signal from generator 2 or from generator 1 will reach the ALC control loop of the other generator as interference (Fig. 1-12 shows only the influence of generator 2 on generator 1, not vice versa).

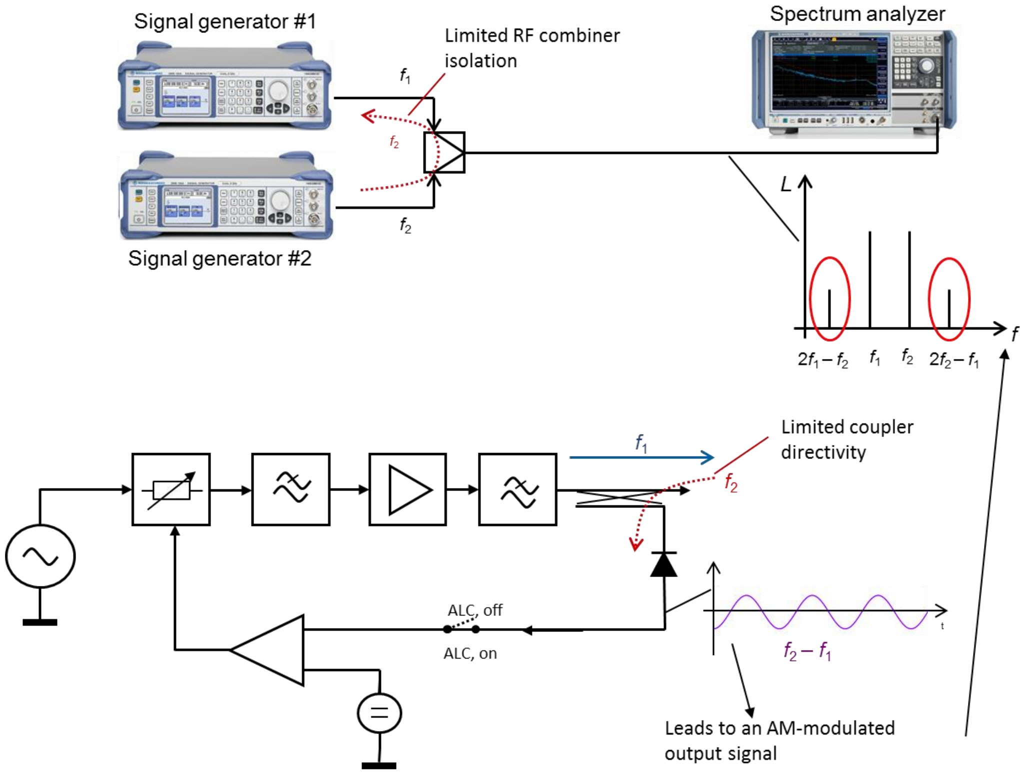


Fig. 1-12: Influence of the ALC in a multitone measurement.

This interference causes AM modulation on both test signals, resulting in the following lower and upper sidebands:

Sidebands around the carrier with frequency f_1:

$f_1 \pm (f_2 – f_1)$

→LSB: $2f_1 – f_2$

→USB: f_2

Sidebands around the carrier with frequency f_2:

$f_2 \pm (f_2 – f_1)$

→LSB: f_1

→USB: $2f_2 – f_1$

These make it clear that the sidebands fall precisely on the third harmonic intercept point for the IP3 measurement, thereby falsifying the measurement. To prevent this, the ALC must be switched off on both signal generators during this type of measurement. If the frequency spacing of the two generator signals is greater than the control bandwidth of the ALC, the interference component from generator 1 or 2 will not affect the ALC, which can therefore remain active.

Level attenuation using step attenuators:

Because the level attenuation available from the ALC is not sufficient for many applications, a step attenuator is inserted between the ALC and the signal generator output. Several highly accurate attenuators at varying attenuation values are combined via RF switches (Fig. 1-13). A step attenuator is set up so that it ideally will have no effect on the signal other than the signal attenuation.

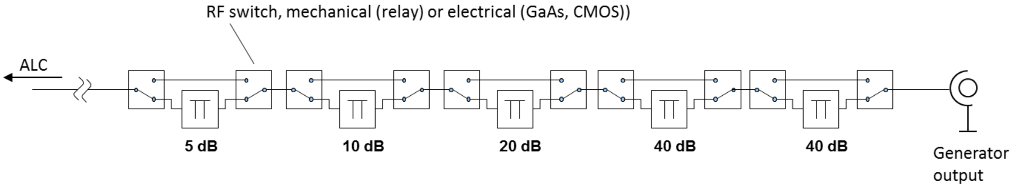


Fig. 1-13: Step attenuator with maximum 115 dB attenuation.

Both mechanical and electronic step attenuators are used in signal generators (Fig. 1-15). Relays function as the switches on mechanical step attenuators. The advantage lies in the low insertion loss of the switches, the low temperature drift and the very good voltage standing wave ratio (VSWR). In addition, only mechanical step attenuators can be used for frequencies in the higher two-digit GHz range. Their disadvantage lies in the relatively slow switching times and the limited lifespan of several million switching cycles. They therefore have only limited suitability for applications in production.

Electronic step attenuators use semiconductor switches with GaAs or CMOS technology. They offer significantly better switching times and have no mechanical wear. Their disadvantage is the greater temperature drift as compared to mechanical step attenuators. An electronic step attenuator also has significantly greater insertion loss than a mechanical one. To compensate, a power amplifier is connected before the signal generator output in order to achieve higher output levels (Fig. 1-14). This negatively affects the spectral purity of the output signal. Because the insertion loss always increases with the frequency, electronic step attenuators can in practice be used only up to about 6 GHz for GaAs technology and up to about 12 GHz for CMOS technology.

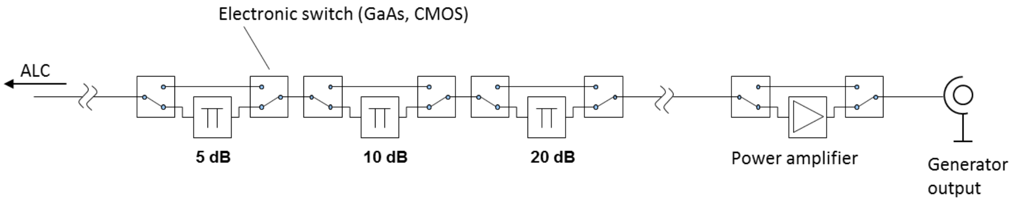


Fig. 1-14: Electronic step attenuator with downstream power amplifier.

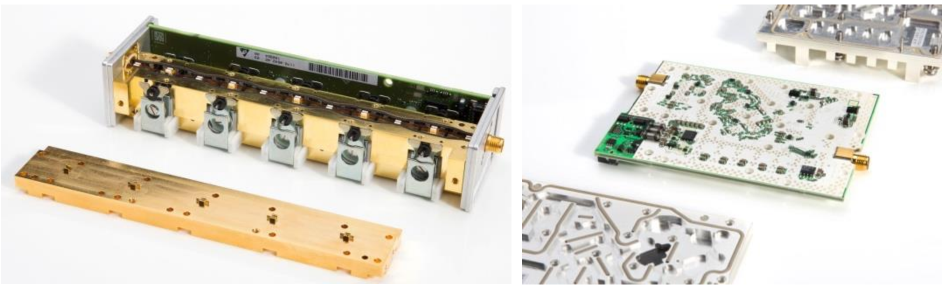


Fig. 1-15: Mechanical step attenuator (left), electronic step attenuator (right).

The traces in Fig. 1-16 show the level deviation of a signal generator for four different frequencies. The comparison between the mechanical and the electronic step attenuator shows that, on average, the total deviation from a defined reference level is higher for the electronic step attenuator. However, in certain level ranges, the level accuracy is fully comparable to that of a mechanically switched attenuator. The traces for the electronic step attenuator at the bottom of Fig. 1-16 also show a consistent level versus time over the entire dynamic range, making any compensation easier.

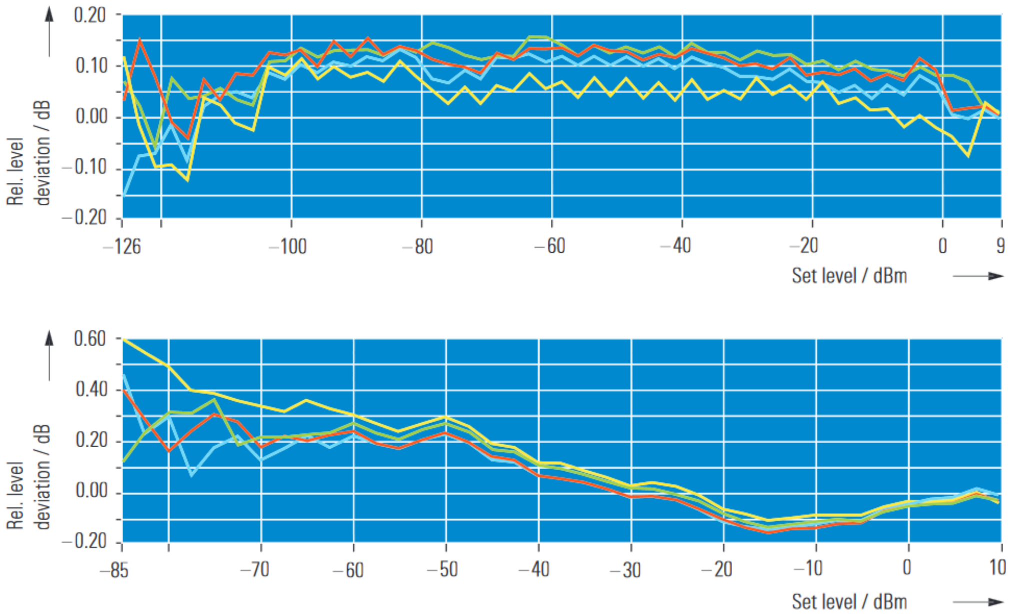


Fig. 1-16: The measured relative level deviation from the defined level for a mechanical step attenuator (top) and for an electronic step attenuator (bottom). Blue: 89 MHz; red: 900 MHz; green: 1.9 GHz; yellow: 3.3 GHz.

1.2.2 Vector Signal Generators

Vector signal generators upconvert modulation signals (external or internal, analog or digital) to the RF frequency and then output the signals. The modulation signal is digitally generated and processed as a complex I/Q data stream in the baseband. 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 add Gaussian noise into the generated signal. This is helpful for investigating the limit at which a noisy signal can still be correctly demodulated by a receiver, for example. Moreover, some generators are able to numerically simulate multipath propagation (fading, MIMO) that will later occur for the RF signal. As when adding noise, this can also be used to determine how the input signal characteristics affect the demodulation in the receiver. In general, the complete generation of the baseband signal is accomplished through realtime computation. ARB generators are an exception to this (see section 1.2.3).

The generated baseband I/Q data is then converted to an RF operating frequency (some vector generators operate only in the baseband without converting to RF signals). Vector signal generators often also include analog or digital I/Q inputs for including externally generated baseband signals.

Using I/Q technology (see also section 2.3 Principles of I/Q Modulation) makes it possible to implement any modulation types – whether simple or complex, digital or analog – as well as single-carrier and multicarrier signals. The signal demodulated by a spectrum analyzer in
Fig. 1-17 was generated by a vector signal generator. Of course, there is no practical application for this "four-quadrant beer mug", but it does illustrate how a modern vector signal generator can be used to produce just about any conceivable waveform.

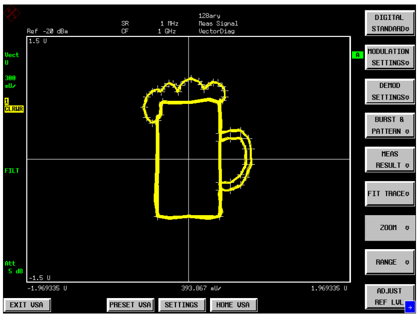


Fig. 1-17: I/Q diagram of a signal generated by a vector signal generator.

The requirements that the vector signal generators must meet are derived primarily 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 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, or MIMO), with and without phase coherence for beamforming
- Generating modulated sources of interference for blocking tests and for measuring suppression of adjacent channels

Fig. 1-18 uses the example of a digital transmission system to show the typical area of application for a vector signal generator. This includes transmitter-side generation of digital baseband signals, coding and conversion to RF. The signal sampling and displayed for the transmitter with a subsequent A/D conversion do not have to be handled by a vector signal generator because it can generate digital signals on its own. In addition to baseband signal generation, modern instruments can insert additive white Gaussian noise (AWGN) on the signal and simulate multipath propagation (fading). With these functions, it is possible to generate signals as they would appear upon reception following transmission via a realistic radio channel, complete with interference. This educational note does not go into any more detail regarding the simulation of the transmission channel. A spectrum analyzer is typically used for the signal analysis on the receiver end (see Educational Note 1MA201).

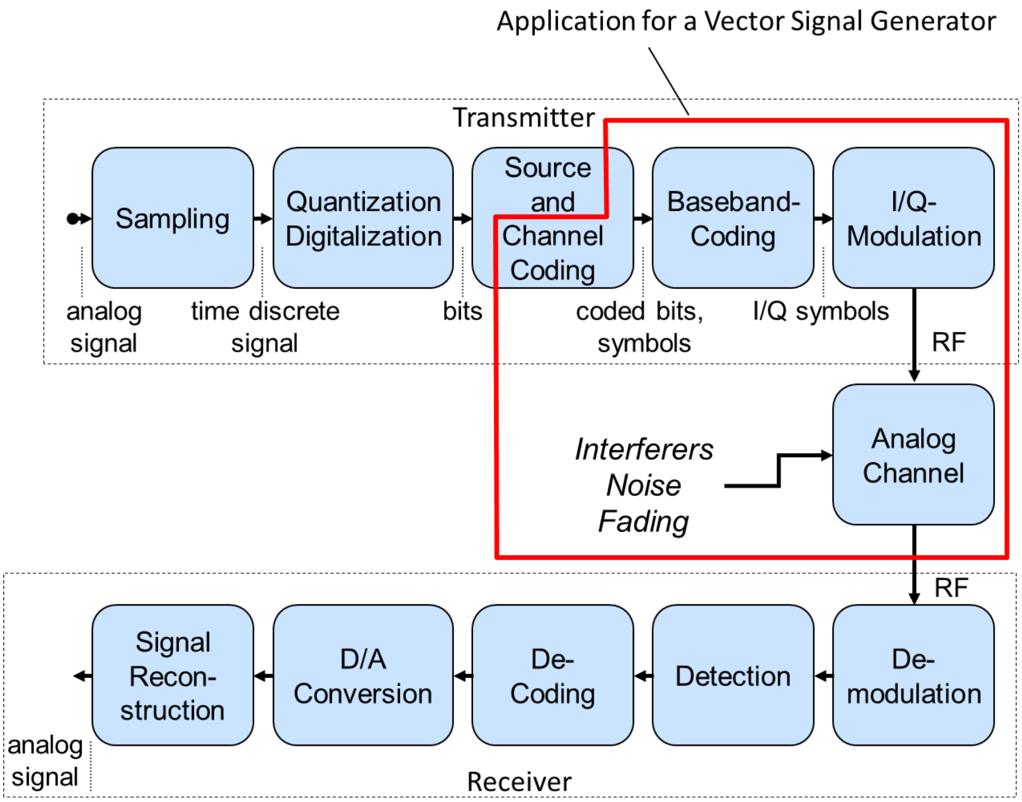


Fig. 1-18: Digital data transmission (voice, data) with the range of applications for vector signal generators.

As an example, Fig. 1-19 shows a portion of the preprogrammed standards that a vector signal generator supports:

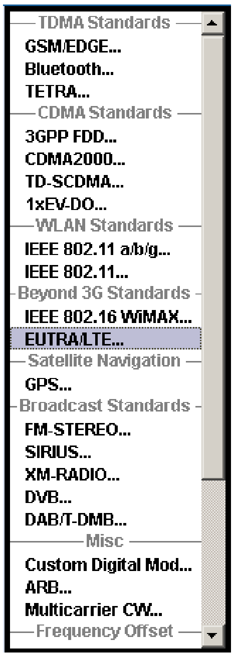


Fig. 1-19: Preprogrammed standards for a vector signal generator.

For most of the communciations standards shown in the upper figure, predefined test signals are provided by test models coupled with a configuration specified in the standard. The test models ensure that the measurements defined in the standard are always performed using the appropriate signals. This also makes it easier to compare measurements against one another. Vector signal generators can store the preprogrammed test models for the various standards. Fig. 1-20 shows a selection of these test models for the LTE standard.

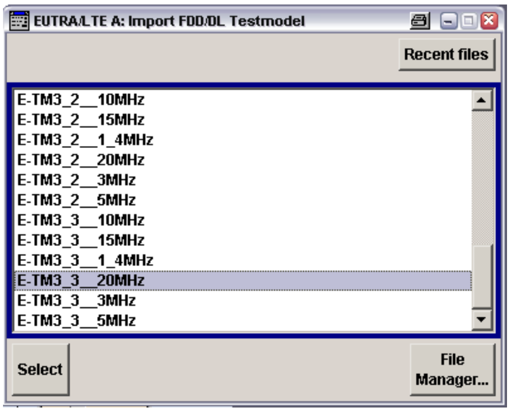


Fig. 1-20: Some of the preprogrammed test models for the LTE wireless communications standard.

Fig. 1-21 shows the spectrum for the selected E-TM3_3__20MHz test model.

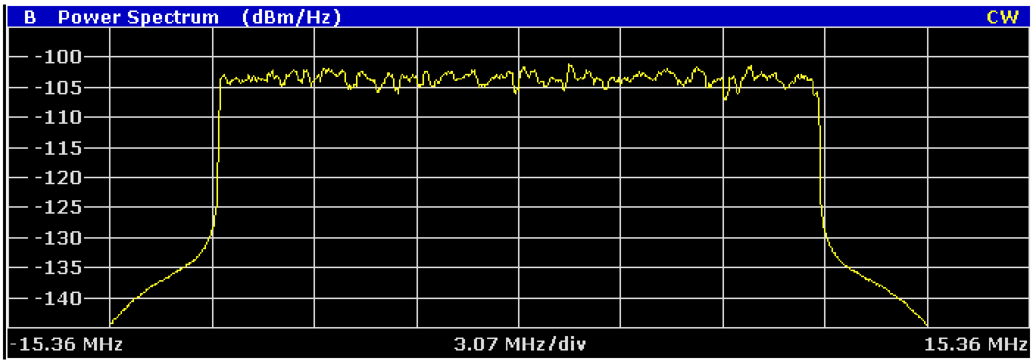


Fig. 1-21: Multicarrier spectrum for the E-TM3_3__20MHz test model from the LTE standard.

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

Fig. 1-22 shows the constellation diagram (I/Q display) for this test model.

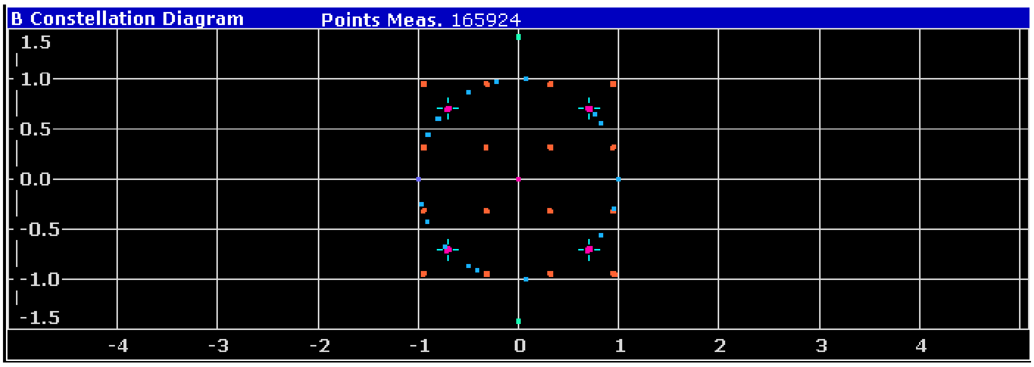


Fig. 1-22: Overall constellation for the E-TM3_3__20MHz LTE test model.

The individual channels of this signal are modulated differently. All of the modulation types being used are summarized in a single display:
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.

In addition to the predefined test models and settings, today's vector signal generators allow nearly all parameters of a digital standard to be modified. This makes it possible to generate signals for almost any conceivable scenario that might be required for specialized testing in development and production. For example, Fig. 1-23 shows the input window for configuring an LTE frame.

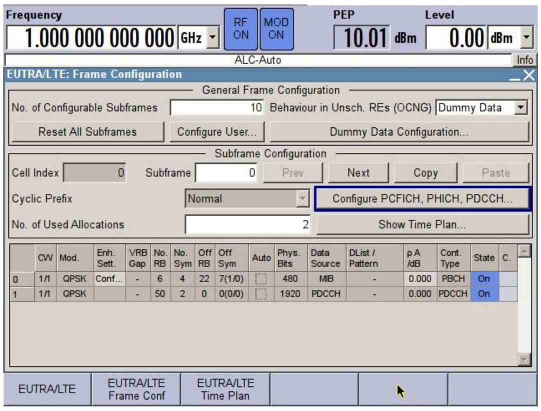


Fig. 1-23: Configuration of an LTE frame.

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 more readily cause harmonics and intermodulation products. 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) or adjacent channel leakage ratio (ACLR). For more information, see section 1.3 Key Signal Generator Characteristics1.3. State-of-the-art generators far exceed the requirements of the current mobile radio standards, making them ready for the future. They also support a large number of international standards.

As for analog generators, general criteria for selecting a vector signal generator include the required output power, the settling time and accuracy for frequency and level, plus a low reflection coefficient or VSWR. Additional considerations include low EVM values, high dynamic range for ACP measurements and the number of supported standards.

1.2.2.1 Design and functioning of a vector signal generator

The essential components of a vector signal generator are identical to that of an analog generator. The synthesizer and output stage with ALC and step attenuator are found here as well. As shown in Fig. 1-24, a baseband generator and an I-Q modulator are used to generate the modulation (analog and digital) or to generate standard-compliant mobile radio signals.

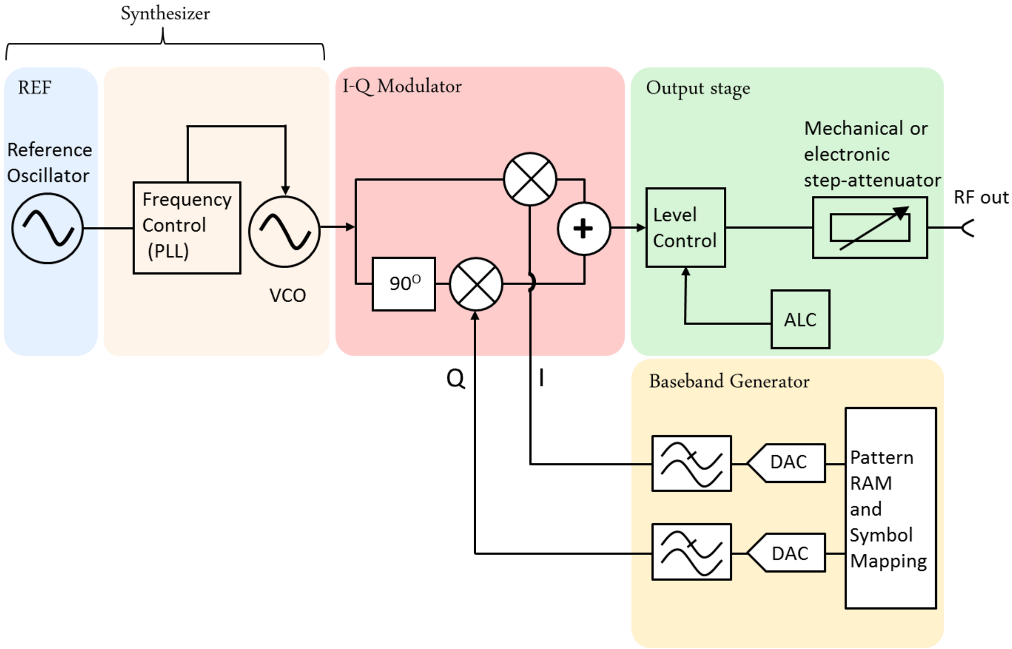


Fig. 1-24: Schematic of a vector signal generator.

Baseband generator

A vector signal generator produces modulated signals using I/Q modulation (see also section 2.3). Often, an integrated baseband generator supplies in realtime the modulation data that an I-Q modulator uses to modulate an RF carrier. Alternatively, a vector signal generator can be supplied with the complex modulation signals and a universal analog I and Q input via an external true baseband generator. Although digital supply is also common, only proprietary inputs are available because no industry-wide standard has been agreed upon for digital modulation using signal generators.

Fig. 1-25 shows how a baseband generator works. In the first step, a data generator (realtime coder) calculates and generates the digital modulation data for the selected signal standard from the applications of mobile radio, satellite navigation, avionics, broadcast, etc. The raw data from the data generator can undergo error protection coding (channel coding) as needed. Channel coding adds redundancy to the wanted data. Special methods can then be used to correct transmission errors caused by interference in the transmission channel. The coding method used depends on the selected signal standard and is defined as a requirement in the specification for that standard. The data stream generated in this manner is stored in memory. A certain number of bits, which varies depending on the modulation mode, are then combined into a symbol. In this example of a quadrature phase-shift keying (QPSK) modulation, two bits make up one symbol. With QPSK, therefore, two bits can be transmitted with each symbol. For higher data rates, higher-order modulation modes are used. In 256 QAM modulation, for example, 8 bits are transmitted per symbol. This provides 256 different possible I/Q values (signal states) ($2^8 = 256$). After the conversion to symbols, the mapping takes place. This process uses a mapping table to reproduce the symbols into points of the I/Q plane. In this example, the symbols of a QPSK modulation after mapping appear as points or vectors in all four quadrants of the I/Q plane, which is also known as a constellation diagram. Read counterclockwise, the four points represent bit sequences 00, 01, 10 and 11.

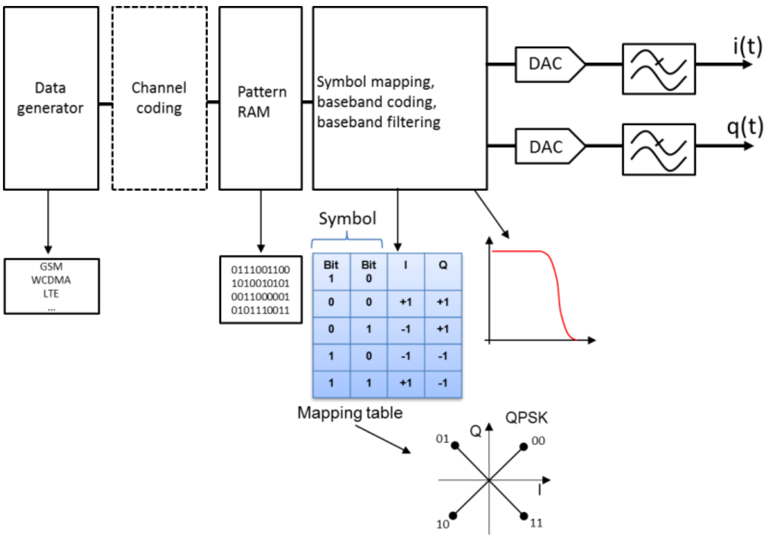


Fig. 1-25: Baseband generator functioning.

Fig. 1-26 shows the data streams i(t) and q(t) after the mapping of a QPSK-modulated signal and the derived vector position.

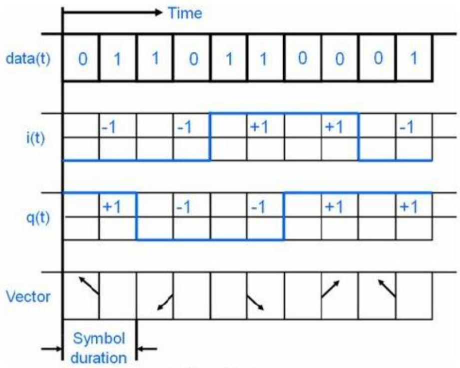


Fig. 1-26: Time characteristic of i(t) and q(t) for QPSK.

To increase the data rates, each symbol must transmit a greater number of bits. This can be done for example by a 32 amplitude and phase-shift keying (32-APSK) modulation, used for digital video broadcasting - satellite (DVB-S2). Instead of the two bits used for QPSK, each symbol in 32-ASPK transmits five bits. In order to display all possible bit combinations, $2^5 = 32$ points are needed in the constellation diagram. As seen in Fig. 1-27, ASPK also modulates the carrier in phase and amplitude. Because the individual constellation points are more densely grouped than for example with QPSK modulation, higher-order modulation methods are more susceptible to interference. This is the case because an imprecisely placed constellation point, e.g. due to noise effects, is easily confused with an adjacent constellation point. The transmitter or vector signal generator should therefore be capable of producing high-quality, digitally modulated signals. In an ideal modulation, there is no deviation of the generated symbols from the ideal constellation. See also section 1.3.5 Error Vector Magnitude (EVM).

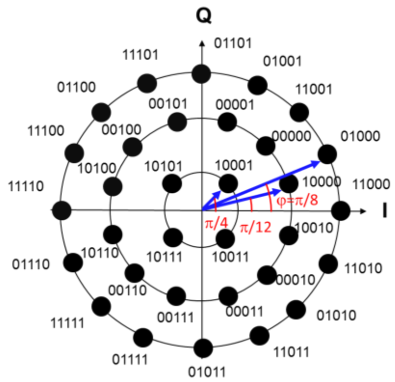


Fig. 1-27: Constellation diagram for 32-ASPK modulation.

In the case of a digital modulation, certain symbol sequences can result in extreme jumps between two points in the constellation diagram. For example, a change in the symbol from 11001 to 11111 (Fig. 1-27) will cause a transition from maximum amplitude – zero crossing – maximum amplitude; amplification of these types of signals requires a great deal of effort as a result of the high crest factor. To prevent these zero crossings, special modulation modes such as π/4-DQPSK are used (see section 2.4.3). To prevent high crest factors (ratio of peak value to RMS value, see section 2.4.6), a baseband coding also takes place in the baseband. With a suitable coding, this should prevent small amplitude jumps without zero crossings at a symbol transition (Fig. 1-28).

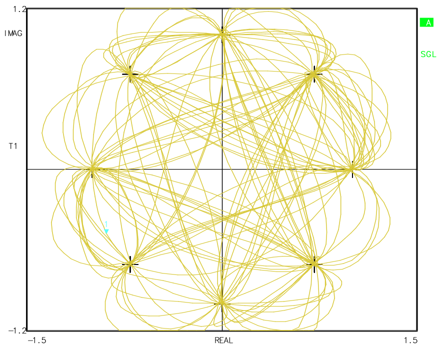


Fig. 1-28: Preventing large amplitude jumps through the use of a π/4-DQPSK modulation.

In the time domain, the data streams from Fig. 1-26 represent square-wave pulse sequences with pulse duration Δt and a pulse period duration T. In the frequency domain, this corresponds to a spectrum whose spectral lines occur at intervals of 1/T. The envelope of the spectral lines follows a sin(x)/x function with a theoretically infinite bandwidth (Fig. 1-29) and zero crossings with periodic frequency spacing 1/Δt.

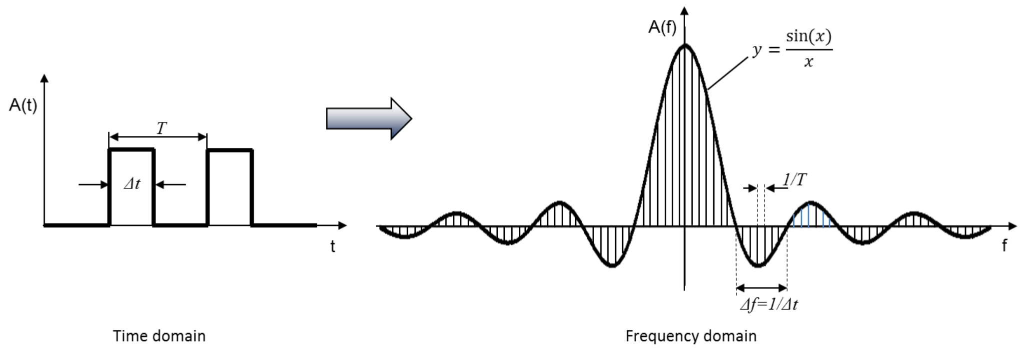


Fig. 1-29: Pulse train in time and frequency domain.

A high-frequency carrier modulated with the pulse sequence from the baseband likewise results in a spectrum with a progression of sin(x)/x. However, this spectrum is shifted from zero by the value of the carrier frequency f_c (Fig. 1-30).

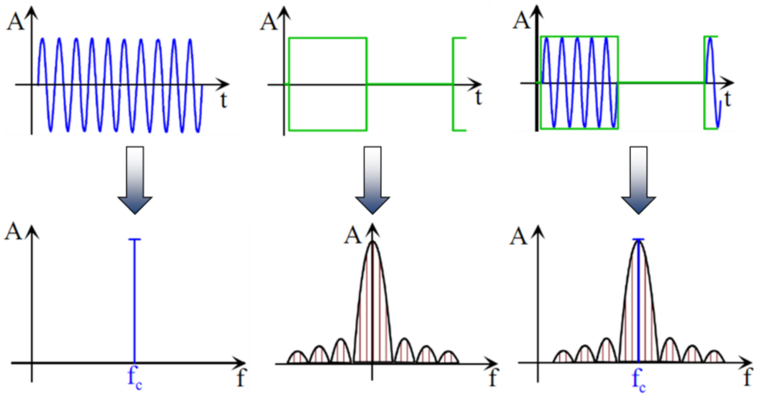


Fig. 1-30: Spectrum of a carrier modulated with square-wave pulses.

To better utilize the available frequency band and to ensure that adjacent frequency bands are not disrupted during the data transmission, it is necessary to limit the bandwidth by using a suitable filter. Because filters for the baseband range can be implemented with less effort than digital filters, the filtering already takes place in the baseband. This has several advantages over filtering the later RF signal:

- The filtering does not require any components, it is done mathematically in the signal generator.
- A lowpass filter is sufficient in the baseband for filtering, and no bandpass filter is required.
- The filter parameters remain constant. An adjustable filter is not needed (such as would be required for the various channel frequencies in the case of the RF signal).

For the demodulation of a digitally modulated signal, it is sufficient to transmit the fundamental of the sin(x)/x function (Fig. 1-31).

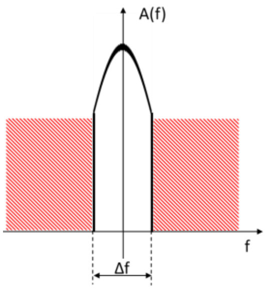


Fig. 1-31: Limited bandwidth of the sin(x)/x function.

In practice, however, the filtering is not performed using a square-wave filter as shown in Fig. 1-31. Square-wave filters (also called ideal lowpass filters) lead to strong, disruptive signal overshoots in the time domain (Fig. 1-30). This makes it impossible to correctly distinguish the individual bits in a bit sequence. To prevent this, filters with special filter characteristics or special edge shapes are used. Examples include raised cosine filters and Gaussian filters. The impulse responses in the time domain for the mentioned filter types are shown in Fig. 1-32.

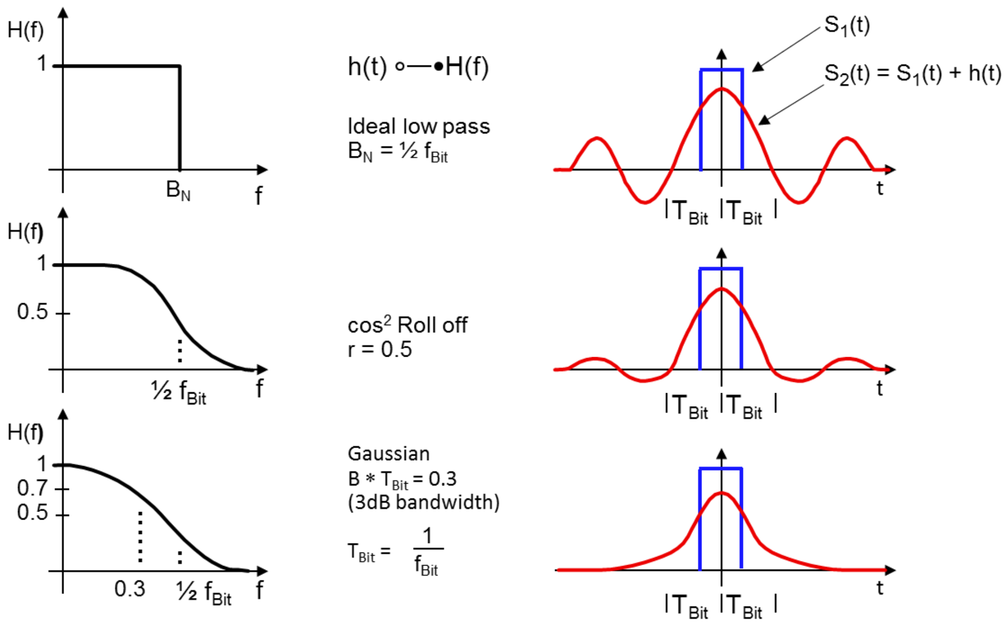


Fig. 1-32: Impulse response $S_2(t)$ for different filter types with transfer function H(f).

The raised cosine filter is a type of Nyquist filter and thus fulfills the first Nyquist criterion. This means that the sum of successive signal impulses after filtering equals zero in the time grid of the sample signals, and therefore they do not impair the preceding or following impulses at their respective sampling times. As a result, a time-discrete signal transmission is ensured without intersymbol interference (ISI). In other words, the individual impulses do not shift into one another, and thus remain distinguishable. To further minimize the necessary bandwidth, there are some instances where intersymbol interference is intentionally permitted in the transmission channel. In this case, the baseband signal is formed with a root-raised cosine filter (RRC filter). The RRC filter corresponds to the root of the raised cosine filter. It can be used to divide the characteristics of the raised cosine filter between the transmitter and the receiver. As such, it represents a matched filter, maximizing the signal-to-noise ratio at the receiver. However, the RRC filter in and of itself does not exhibit any intersymbol interference (ISI). It is only when the two RRC filters are combined at the transmitter and the receiver that a raised cosine filter results, ideally making an ISI-free transmission possible, and hence also permitting the individual impulses to be clearly distinguished over time. Next to the Gaussian filter, the RRC filter is one of the most frequently used filters for impulse formation in digital signal transmission.

The transfer function of the raised cosine filter and the RRC filter is dependent on both the symbol rate (1/T) and the rolloff factor r. Fig. 1-33 shows the transfer function of the above-mentioned filters and the derived rolloff factor. The rolloff factor can be a value between 0 and 1 and influences the steepness of the transmission curve.

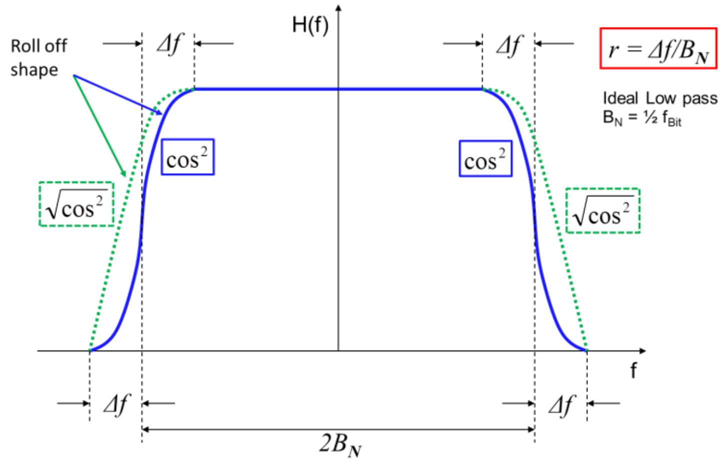


Fig. 1-33: Rolloff factor definition.

Fig. 1-34 shows the transfer function H(f) and time function h(t) of a raised cosine filter with four different rolloff factors. Where r = 0, the result is an ideal lowpass with a square-wave transfer function, while r = 1 delivers the flattest possible cosine edge. Where r is an intermediate value, the frequency response is close to constant over a certain range, then falls into a somewhat steeper cosine edge as the values for r decrease. The filter bandwidth increases as the rolloff factor increases. The maximum value of r = 1 is equal to the maximum bandwidth of 1/T. A decreasing rolloff factor produces a steeper filter edge and leads to greater undesirable overshoots in the time domain. In practice, however, a smaller rolloff factor in the range from 0.2 to 0.5 will save bandwidth. For example, the UMTS mobile cellular standard uses a rolloff factor of r = 0.22 for its impulse filter.

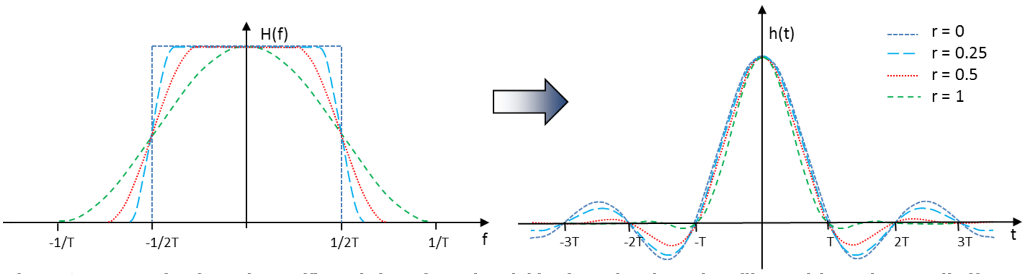


Fig. 1-34: Transfer function H(f) and time function h(t) of a raised cosine filter with various rolloff factors.

Gaussian filters do not exhibit any overshoots in the impulse response. They are used in the Gaussian minimum shift keying (GMSK) digital modulation method for GSM, for example. The square-wave transmit symbols are converted into impulses in the shape of a Gaussian bell-shaped curve with lower bandwidth requirements (Fig. 1-35).

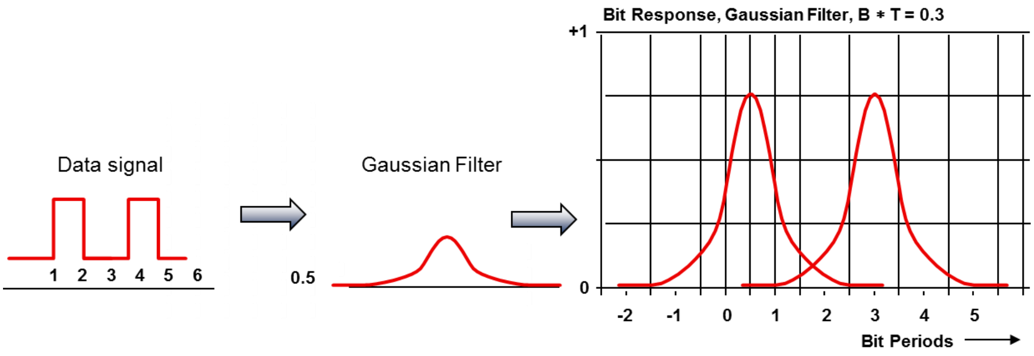


Fig. 1-35: Filtering of square-wave data symbols using a Gaussian filter.

The filter type and parameters (e.g. rolloff factor) being used depend on the mobile cellular standard and are defined in the respective specification. A vector signal generator must therefore be capable of performing a baseband filtering using a variety of filter types. Most signal generators also permit the filter parameters to be set manually for development tasks. The filtering is performed based on realtime calculations of digital filters. The filtered and coded baseband signals are then converted into analog I and Q signals, passed through a lowpass filter for signal shaping and then sent to the I-Q modulator (see the next section).

Each filter pass will always cause smoothing of the transitions. Fig. 1-36 shows filtered baseband signals i(t) and q(t) over time, based on the example of a 3GPP UMTS signal (dual BPSK signal).

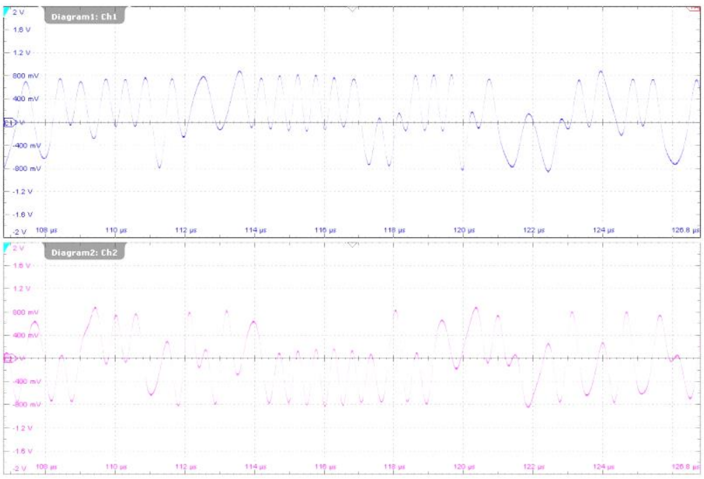


Fig. 1-36: Filtered I/Q signals.

These signals no longer give any hint of the erratic data change.

At the left of Fig. 1-37 is the position of an I/Q constellation relative to the symbol times and to the right is the I/Q vector over time. As a result of filtering, this no longer follows the shortest straight line between the symbols, but rather describes the filtering characteristics of the symbol transitions.

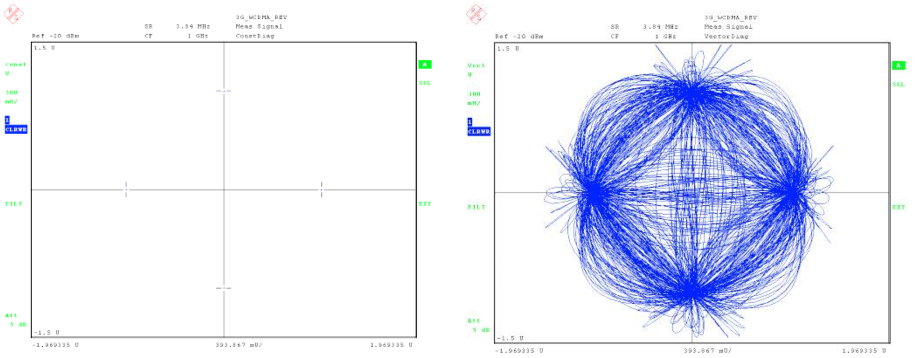


Fig. 1-37: Constellation (left) and continuous vector display.

I-Q modulator

Fig. 1-38 shows how an I-Q modulator works. I-Q modulators consist primarily of two mixers and a signal summer. The unmodulated output signal from the synthesizer functions as the local oscillator for the mixer. The modulated RF signal A(t) is generated by adding one carrier modulated with i(t) and one modulated with q(t) – but phase-shifted by 90°.

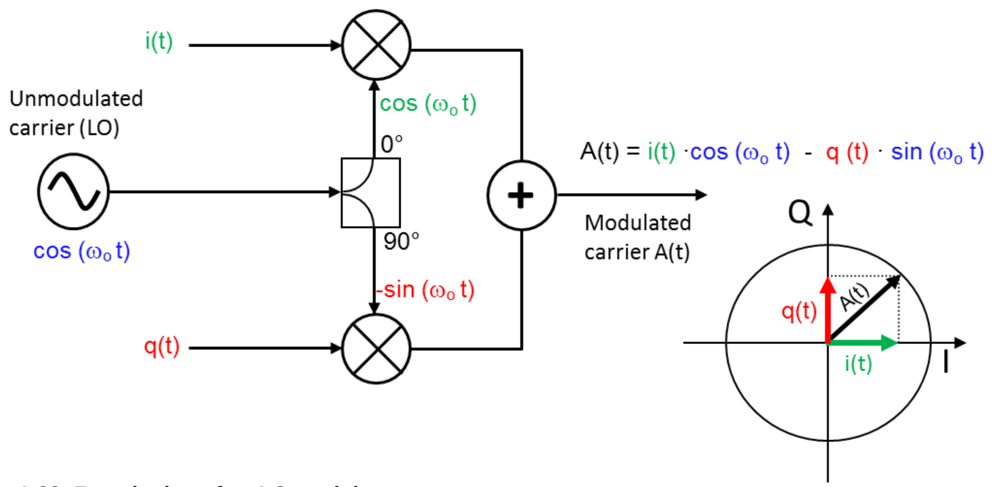


Fig. 1-38: Functioning of an I-Q modulator.

I-Q modulators can thus generate just about any vector signal waveform by changing the magnitude and phase of a vector (Fig. 1-39).

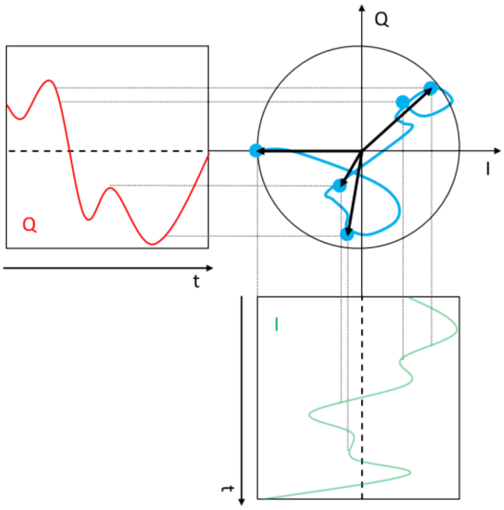


Fig. 1-39: Using an I-Q modulator to generate a signal from the I and Q signal.

For even more flexibility during signal generation, some signal generators can optionally also include an arbitrary waveform generator (ARB) as the baseband signal source (see also section 1.2.3). This generates signals of almost unlimited complexity. The disadvantage is that the signal generation cannot be performed in realtime and a precalculated signal waveform must be available, for example created using PC software or a mathematical software application.

1.2.3 Arbitrary Waveform Generators (ARBs)

Arbitrary waveform generators (ARBs) are specialized vector signal generators for the baseband. The modulation data for ARBs is calculated in advance (rather than in realtime) and stored in the instrument's RAM. The memory content is then output at the realtime symbol rate. Many vector signal generators have an ARB option; see selection in Fig. 1-19. ARBs differ from realtime vector generators in use and application as listed here:

- ARBs have zero restrictions for configuring the content of an I/Q data stream (hence the name "arbitrary").
- Only time-limited or cyclic signals are allowed (the memory depth is finite).

The memory depth and word width for the I/Q data sets are additional characteristics for ARBs. Some ARB generators allow sequential playback of different memory segments containing multiple precalculated waveforms, without any further preliminary calculation.

As with realtime generators, ARBs offer different triggering options as well as the ability to output marker signals for controlling connected hardware and measurement instruments in parallel to the waveform playback. Users can combine a variety of waveforms of varying durations into a single sequence for production tests (see Fig. 1-40). The sequence and duration is defined either in the ARB by means of a remote control program or by using an external trigger signal. For example, the output sequence can consist of data streams with varying bit rates that must be verified during production.

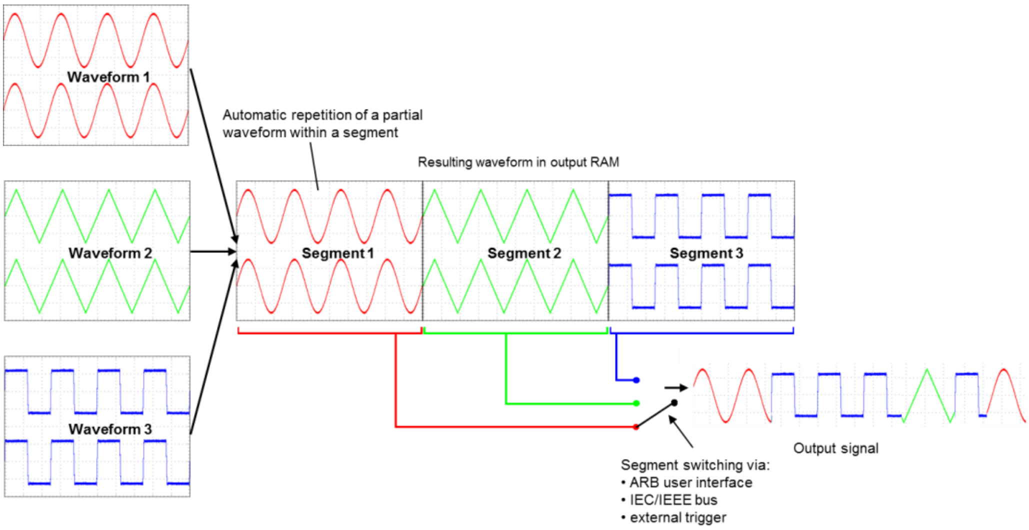


Fig. 1-40: Stringing together different waveforms into a sequence.

Many ARB generators can calculate additive Gaussian noise as a baseband source; some are also able to simulate multipath propagation (fading) or multi-antenna systems (MIMO) of the subsequent RF signal. In many cases, vendors that offer ARB generators also offer software for creating standard modulation sequences (I/Q data sets). As an example, Fig. 1-41 shows several screens from the R&S® WinIQSIM2 simulation software.

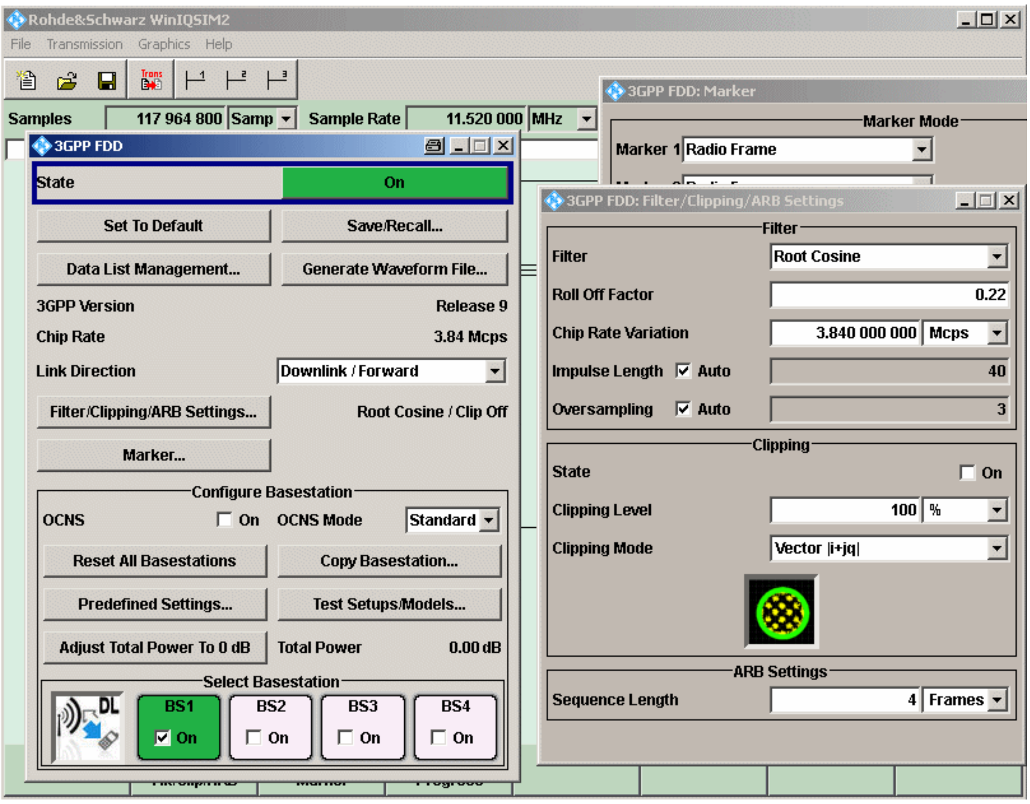


Fig. 1-41: PC program for calculating the I/Q data for standard signals.

This example uses the 3GPP FDD (UMTS) mobile cellular standard. It illustrates the creation of a downlink, i.e. 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. 1-41, only BS1 is active. The filtering complies with the UMTS standard. No *clipping* is performed. The *Marker1* device jack will subsequently deliver a signal with each new *radio frame*.

Once all the required entries have been made, the user initiates calculation of the I/Q data by clicking the *Generate Waveform File* button. The data is then transferred from the program to the ARB, and the output can be started immediately.

Fig. 1-42 shows the basic functioning of an ARB. The sequencer takes the waveforms saved in memory and combines them into a predefined sequence, then forwards them to the sample rate converter (resampler).

The resampler is used to adjust a variable memory symbol clock to a fixed clock rate from the D/A converter. This has two major advantages. First, the D/A converter always uses the optimal clock rate for the best possible resolution, and second, the variable memory clock ensures that the memory is optimally utilized. In other words, changing the clock ensures that the longest possible sequence can be generated.

This is illustrated in the following example:

An ARB with a memory depth of 1 Gsample can generate a GSM signal with 270 ksamples/s and triple oversampling with a duration of 1234 s.

$\left(t = \frac{1\ Gsample}{3 \cdot 270\ ksample}\right)$. If the memory clock rate is equal to that of the D/A converter, for example 300 MHz, the possible duration for the signal output would be reduced to 3.3 s $\left(t = \frac{1\ Gsample}{300\ Msample/s}\right)$.

Because the clock rate of the D/A converter is always within the optimal range, the output signal has a high, spurious-free dynamic range (SFDR) that is independent of the memory clock rate**.**

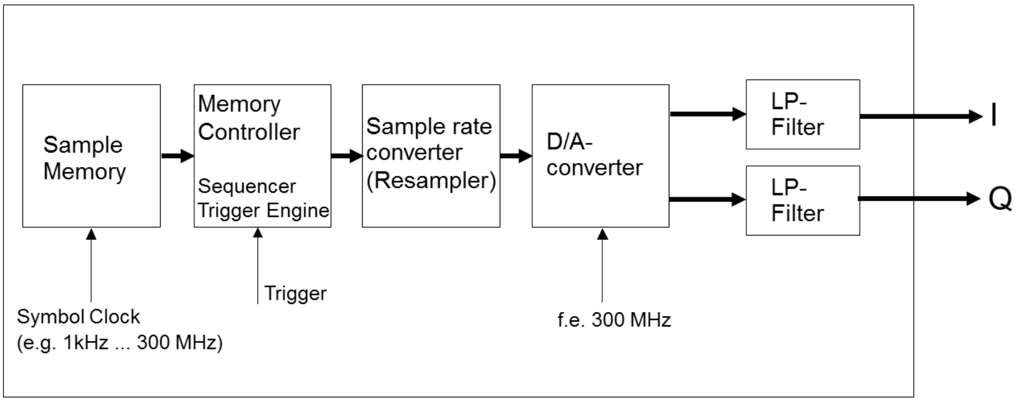


Fig. 1-42: ARB block diagram.

1.3 Key Signal Generator Characteristics

An ideal assumed signal generator delivers an undistorted sine-wave signal that exactly matches the defined frequency and the defined level at its output. Fig. 1-43 shows an ideal sine-wave signal in the time and frequency domains.

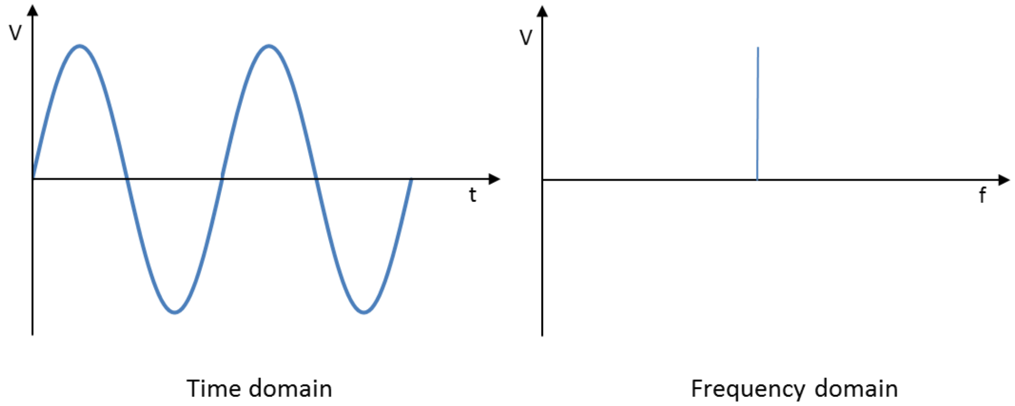


Fig. 1-43: Output signal of an ideal signal generator in the time and frequency domains.

Because a signal generator is constructed from real components with non-ideal characteristics, the wanted carrier signal with frequency f_c (c: carrier) arrives at the output along with other, unwanted signals. In reality, even the carrier signal will include a level error and a frequency error. In fact, as a result of the phase noise, the real carrier signal is not consistently displayed in the spectrum as a thin line, or Dirac pulse (see Fig. 1-44).

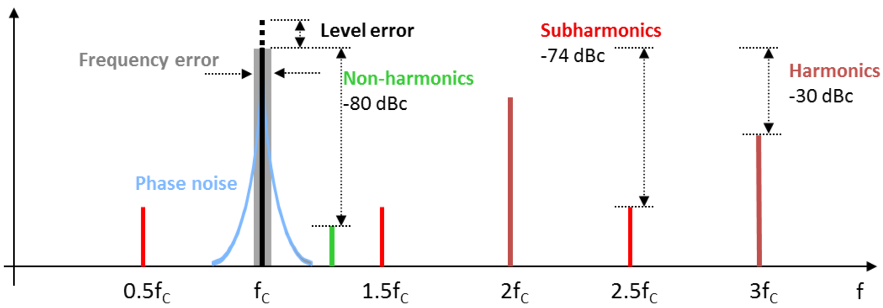


Fig. 1-44: Real spectrum for a non-ideal signal generator.

The following sections take a closer look at the key signal generator characteristics used to assess signal quality.

1.3.1 Phase Noise

Phase noise is a measure of the short-term stability of oscillators as used in a signal generator for producing signals of varying frequency and shape. Phase noise is caused by fluctuations in phase, frequency and amplitudes of an oscillator output signal, although the latter are usually not a consideration. These fluctuations have a modulating effect. Phase noise typically occurs symmetrically around the carrier, which is why it is sufficient to consider only the phase noise on one side of the carrier. Its amplitude is therefore specified as **s**ingle **s**ideband **p**hase noise (SSB) at a defined carrier offset referenced to the carrier amplitude. The specified values are usually relative, i.e. they are defined as a noise power level within a bandwidth of 1 Hz. Accordingly, the unit is noted as dBc (1 Hz) or as dBc/Hz, where "c" references the carrier. Because the phase noise power is lower than the carrier power, negative values are to be expected in the specifications for a signal generator. The effects of phase noise are shown in Fig. 1-45.

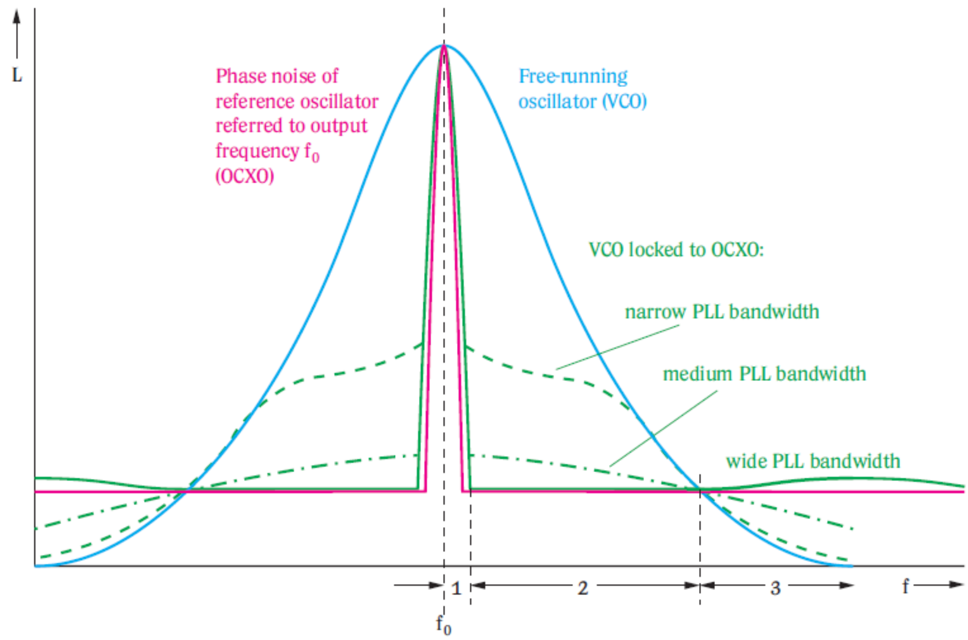


Fig. 1-45: Phase noise of an OCXO, a VCO and a VCO connected to the OCXO at varying PLL bandwidths.

Ideally, a pure sine-wave signal would result in a single spectral line in the frequency domain. In practice, however, the spectrum of a signal generated by a real oscillator is significantly broader. Signals from all oscillators exhibit phase noise to some degree. By making appropriate changes to the circuitry, this can be minimized to a certain extent, but never completely eliminated. In modern signal generators, oscillators are implemented as synthesizers, i.e. the actual oscillators are locked to a highly precise reference frequency, e.g. 10 MHz, by phase-locked loops as described in 1.2.1.1. The phase noise characteristics are affected by the PLL bandwidth of the frequency locking circuitry. The following subranges can therefore be identified (see also ranges 1, 2 and 3 in Fig. 1-45):

- Close to the carrier (frequency offset up to about 1 kHz)

  In this range, the phase noise caused by the multiplication in the locked loop is higher than that from the reference oscillator.

- Range extending to the upper boundary of the PLL bandwidth (starting at about 1 kHz)

Within the PLL bandwidth, the phase noise corresponds to the total noise from multiple components in the locked loop, including the attenuator, the phase detector and the multiplexed reference signal. The upper boundary of this range is dependent on the signal generator, or more accurately, on the type of oscillator used.

- Range outside of the control bandwidth

  Outside of the control bandwidth, the phase noise is determined almost entirely by the phase noise of the oscillator during unsynchronized operation. In this range, it falls 20 dB per decade.

Fig. 1-45 shows the phase noise for the various PLL bandwidths. Of interest is the comparison of the phase noise from the unsynchronized oscillator against the phase noise using the various PLL bandwidths as reference. The following different situations can occur:

- Large PLL bandwidth

  The loop gain in the locked loop is so high that the noise from the oscillator connected to the reference is reduced to the level of the noise from the reference. Far away from the carrier, however, the phase noise increases as a result of the phase shift caused by the filtering.

- Medium PLL bandwidth

  The loop gain is not sufficient to reach the reference noise close to the carrier. However, the increase in phase noise far away from the carrier is less than with a large control bandwidth.

- Narrow PLL bandwidth

  The phase noise far away from the carrier is not any worse than that from the unsynchronized oscillator. Close to the carrier however, it is significantly greater compared to the medium and large control bandwidths.

Mathematical description of phase noise:

The output signal v(t) of an ideal oscillator can be described by

Equation 1-3:

$v(t) = V_o \cdot \sin(2\pi f_o t)$ where

V_o	Signal amplitude
F_0	Signal frequency and
$2\pi f_o t$	Signal phase

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

Equation 1-4:

$v(t) = (V_o + \varepsilon(t)) \sin(2\pi f_o + \Delta\varphi(t))$ where

$\varepsilon(t)$ 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 fluctuation due to e.g. AC hum or to insufficient suppression of other frequencies during signal processing. These fluctuations appear as discrete lines of interference.
- Random phase fluctuations caused by thermal, shot or flicker noise in the active elements of oscillators.

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

Equation 1-5:

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

In practice, single sideband (SSB) phase noise L is usually used to describe an oscillator's phase-noise characteristics. It is defined as the ratio of the noise power measured in a single sideband P_{SSB} at 1 Hz of bandwidth) to the signal power $P_{Carrier}$ at a frequency offset f_{offset} from the carrier (see also Fig. 1-46).

Equation 1-6:

$$L(f_{offset}) = \frac{P_{\mathrm{SSB}}(1\mathrm{Hz})}{P_{\mathrm{Carrier}}}$$

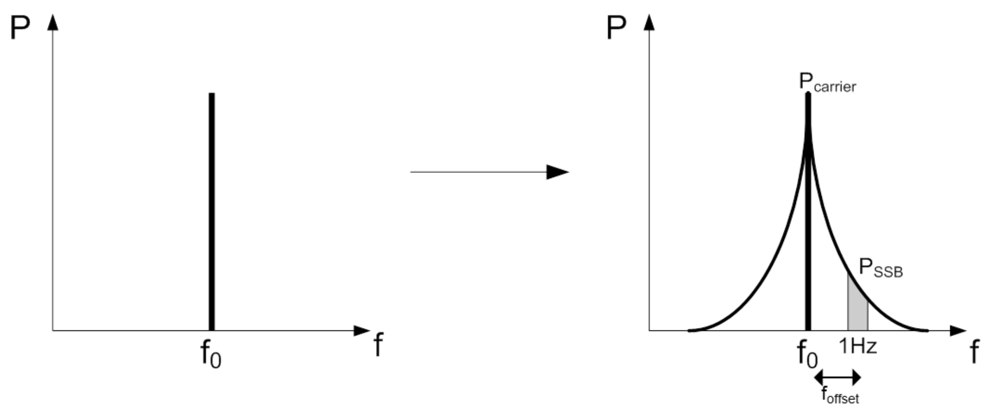


Fig. 1-46: Ideal signal and signal with phase noise.

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

Equation 1-7:

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

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

Equation 1-8:

$$L_c(f_{offset}) = 10\lg\,(L(f_{offset}))\ \mathrm{dBc}$$

The SSB phase noise from a signal generator based on the frequency offset to the carrier is shown in Fig. 1-47. Because the components used for signal generation possess frequency-dependent characteristics, and because the higher frequencies require additional signal paths for the signal generation, the phase noise is also dependent on the frequency currently defined for the signal generator. Fig. 1-47 shows this dependency for several carrier frequencies ranging from 10 MHz to 6 GHz. This illustrates that phase noise increases almost continuously in relationship to the frequency.

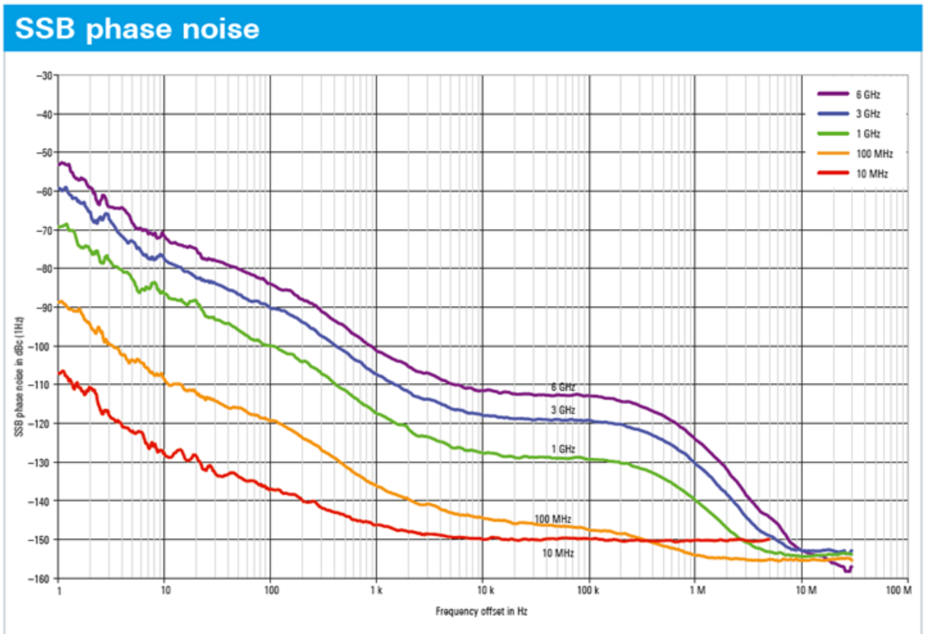


Fig. 1-47: SSB phase noise based on the frequency offset at various carrier frequencies.

Low phase noise values as seen in Fig. 1-47 are particularly significant when a signal generator will be used as a stable reference signal, for example for phase noise measurements, or as a calibration reference during servicing. Low-jitter microwave signals, i.e. signals with low SSB phase noise – in particular close to the carrier – are essential for characterizing ever-faster A/D converters.

1.3.2 Spurious Emissions

The spurious emissions shown in Fig. 1-44 (harmonics and subharmonics) are the result of non-linearities in the components (amplifiers, mixers, etc.) in the signal path of the signal generator. This is described in more detail below.

An ideal linear two-port device transfers signals from the input to the output without distorting them. The voltage transfer function of this type of two-port device is:

Equation 1-9:

$$V_{out}(t) = G_v \cdot V_{in}(t)$$

where $V_{out}(t)$: Voltage at the output of the two-port device

$V_{in}(t)$: Voltage at the input of the two-port device

G_V: Voltage gain in the two-port device

In practice, these types of ideal two-port devices can be achieved only by using passive components.

For example, resistive attenuators can be assumed to be ideal. On the other hand, two-port devices containing semiconductor components, e.g. amplifiers or mixers, do exhibit non-linearities. In this case, the transfer function can be approximated by using a power series. The following applies:

Equation 1-10:

$$\mathrm{v}_{out}(\mathrm{t}) = \sum_{n=1}^{\infty} a_n \cdot \mathrm{v}_{in}^n(t) = a_1 \cdot \mathrm{v}_{in}(t) + a_2 \cdot \mathrm{v}_{in}^2(t) + a_3 \cdot \mathrm{v}_{in}^3(t) + \ldots$$

where $v_{out}(t)$: Voltage at the output of the two-port device

$v_{in}(t)$: Voltage at the input of the two-port device

a_n: Coefficient of the respective non-linear element for the voltage gain

In most cases, it is sufficient to consider the square and cubic members, so that the power series need only be developed up to n = 3 as shown in Equation 1-10. Signal generators should deliver the most distortion-free signals possible. Linearity is therefore a key criterion for assessing a signal generator. The effects of a two-port device's non-linearities on its output spectrum are dependent on the input signal:

Single-tone modulation

If a single sine-wave signal ue(t) is present at the input of the two-port device

where

Equation 1-11:

$$v_{in}(t) = \hat{V}_{in} \cdot \sin(2\pi f_{in,1} \cdot t)$$

where $\hat{V}_{in}$ Peak value of $v_{in}(t)$

$f_{in,1}$ Frequency of $v_{in}(t)$,

this is considered to be single-tone modulation. By using Equation 1-11 in Equation 1-10, it can be seen that the non-linearities cause the presence of harmonics of the input signal at the frequencies $f_{n.H} = n \cdot f_1$ (see Fig. 1-48).

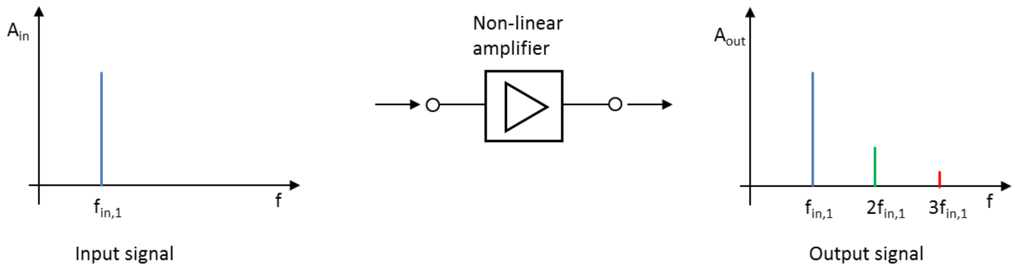


Fig. 1-48: Spectrum before and after a non-linear two-port device.

The level of these harmonics depends on the coefficients a_n in Equation 1-10 . However, there also exists other dependencies on the order n of the respective harmonics and on the input level. When the input level increases, the harmonics levels increase out of proportion with their order, i.e. an input level change of Δ dB leads to a change in the harmonics level of n • Δ dB along with a change in the input level.

Data sheets list the harmonics relative to the carrier signal in dBc (c: carrier). Fig. 1-49 provides an example of the data sheet specifications. The harmonics are specified by the frequency as well as the output power up to which the specifications for the harmonics will apply. If the level offset of the harmonics to the carrier signal is not adequate for the application, optional filters can be used to further reduce the harmonics, as shown in this example. Fig. 1-50 provides a graphical representation of the measurement results of the second and third harmonics over the frequency.

Harmonics		
R&S®SMB-B101/-B102/-B103/-B106 -B112/-B112L(*	1 MHz < f ≤ 6 GHz; level ≤ 13 dBm [6] f > 6 GHz; level ≤ 10 dBm [6]	< –30 dBc
R&S®SMB-B120/-B120L/-B140/-B140L(**	standard; level ≤ 8 dBm [6]	
	f > 1 MHz	< –30 dBc
	with R&S®SMB-B25, R&S®SMB-B26 option low harmonic, low harmonic filter on, level ≤ 10 dBm [6]	
	1 MHz < f ≤ 150 MHz	< –30 dBc
	150 MHz < f ≤ 3 GHz	< –58 dBc
	3 GHz < f ≤ 20 GHz	< –50 dBc
	f > 20 GHz	< –60 dBc (meas)

*) ohne / mit elektronischer Eichleitung; **) ohne / mit mechanischer Eichleitung

Fig. 1-49: Example specifications for harmonics.

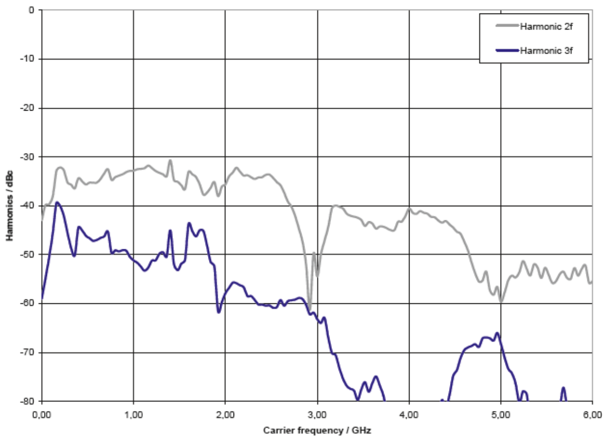


Fig. 1-50: Graphical display of the second and third harmonics up to 6 GHz, output power +15 dBm (measured).

Two-tone modulation

In addition to the harmonics described above, the signal generation also results in non-harmonic signal components. Because the isolation of the signal paths is limited, these components can be caused by crosstalk from internal power supplies or from other clock frequencies present in the signal generator, for example. This includes intermodulation products such as those that occur during two-tone modulation of non-linear components. In the case of two-tone modulation, a signal $V_{in}(t)$ consisting of two sine-wave signals of the same amplitude is applied at the input of the two-port device.

The following applies for the input signal:

Equation 1-12:

$$v_{in}(t) = \hat{V}_{in} \cdot \sin\left(2\pi f_{in,1} \cdot t\right) + \hat{V}_{in} \cdot \sin(2\pi f_{in,2} \cdot t)$$

where $\hat{V}_{in}$ Peak value of the two sine-wave signals

$f_{in,1}$, $f_{in,2}$ Signal frequencies

By using Equation 1-12 in the non-linear transfer function per Equation 1-10, the intermodulation products listed in Table 1-1 are present at the output of the two-port device. The angular frequency ω is always specified with $\omega_1 = 2\pi \cdot f_{in,1}$ and $\omega_2 = 2\pi \cdot f_{in,2}$.

DC component	$a_2 \cdot 0.5\,(\hat{U}_{in,1}^2 + \hat{U}_{in,2}^2)$
Fundamentals	$a_1 \cdot \hat{U}_{in,1} \cdot \sin(\omega_1 t)$ $a_1 \cdot \hat{U}_{in,2} \cdot \sin(\omega_2 t)$
2nd harmonics	$a_2 \cdot 0.5 \cdot \hat{U}_{in,1}^2 \cdot \cos(2 \cdot \omega_1 t)$ $a_2 \cdot 0.5 \cdot \hat{U}_{in,2}^2 \cdot \cos(2 \cdot \omega_2 t)$
Intermodulation products of 2nd order	$a_2 \cdot \hat{U}_{e,1} \cdot \hat{U}_{in,2} \cdot \cos(\omega_1 - \omega_2)t$ $a_2 \cdot \hat{U}_{e,1} \cdot \hat{U}_{in,2} \cdot \cos(\omega_1 + \omega_2)t$
3rd harmonics	$a_3 \cdot 0.25 \cdot \hat{U}_{in,1}^3 \cdot \sin(3 \cdot \omega_1 t)$ $a_3 \cdot 0.25 \cdot \hat{U}_{in,2}^3 \cdot \cos(3 \cdot \omega_2 t)$
Intermodulation products of 3rd order	$a_3 \cdot \hat{U}_{in,1}^2 \cdot \hat{U}_{in,2} \cdot 0.75 \cdot \sin(2\omega_1 + \omega_2)t$ $a_3 \cdot \hat{U}_{in,1}^2 \cdot \hat{U}_{in,2} \cdot 0.75 \cdot \sin(2\omega_2 + \omega_1)t$ $a_3 \cdot \hat{U}_{in,1}^2 \cdot \hat{U}_{in,2} \cdot 0.75 \cdot \sin(2\omega_1 - \omega_2)t$ $a_3 \cdot \hat{U}_{in,1}^2 \cdot \hat{U}_{in,2} \cdot 0.75 \cdot \sin(2\omega_2 - \omega_1)t$

Table 1-1: Intermodulation products from two-tone modulation.

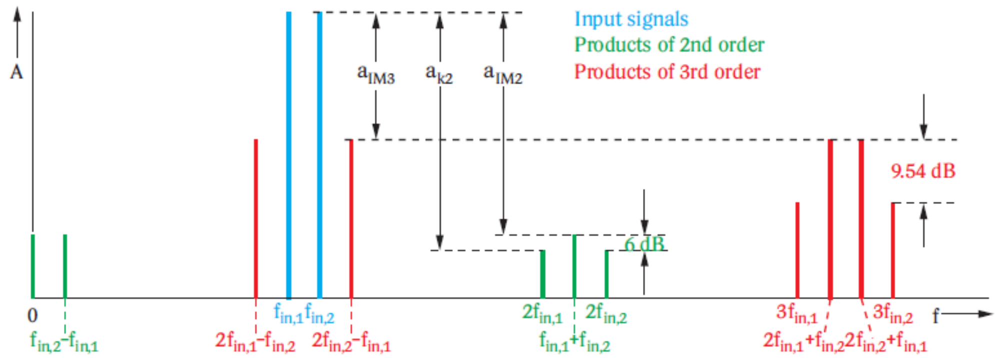


Fig. 1-51: Output spectrum of a non-linear, two-port device during two-tone modulation.

In addition to the harmonics, intermodulation products are also generated (Fig. 1-51). The order of the intermodulation products corresponds to the sum of the orders of the involved components. For the product $2 \cdot f_{in,1} + 1 \cdot f_{in,2}$, for example, the result order is 2 + 1 = 3. Table 1-1 lists only products up to the third order. While even intermodulation products (e.g. second order) in the frequency domain always lie far away from the two input signals, odd intermodulation products of a lower order (e.g., third or fifth order) are always in the immediate vicinity of the input signals.

Depending on the application, products of both even and odd orders can cause interference. Just as for the higher-order harmonics, a level change of Δ dB on both sine-wave carriers at the input leads to a level change of n • Δ dB in the intermodulation products. For example, if the modulation of an amplifier is increased by 3 dB, the third-order intermodulation product increases by 9 dB. Fig. 1-52 shows this interrelationship based on the example of second- and third-order intermodulation products:

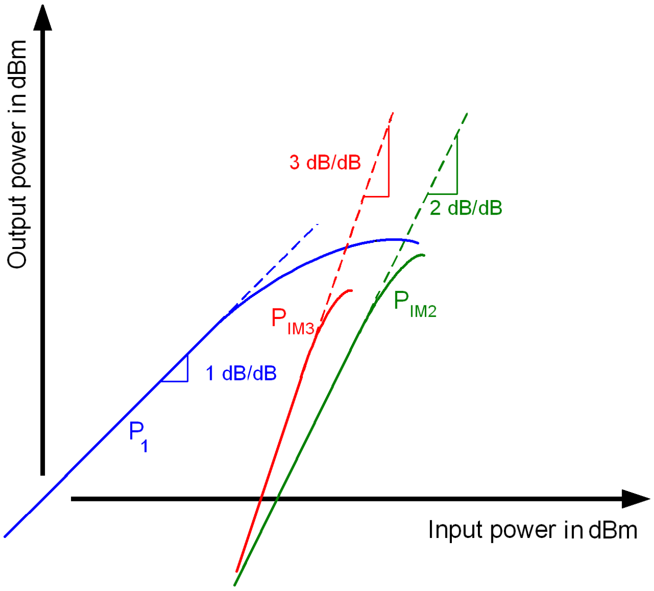


Fig. 1-52: Characteristic curves for the fundamental (blue) and for the second-order (green) and third-order (red) intermodulation products.

To ensure that the specification of non-harmonics for different signal generators can be compared, the level and the frequency offset to the carrier must be included in the specifications (Fig. 1-53).

Nonharmonics	CW, level > –10 dBm (level > 0 dBm for instruments without step attenuator), offset > 10 kHz from carrier	
	f ≤ 23.4375 MHz	< –70 dBc
	23.4375 MHz < f ≤ 1500 MHz	< –70 dBc, < –84 dBc (typ.)
	1500 MHz < f ≤ 3 GHz	< –64 dBc, < –78 dBc (typ.)
	3 GHz < f ≤ 6.375 GHz	< –58 dBc, < –72 dBc (typ.)
	6.375 GHz < f ≤ 12.75 GHz	< –52 dBc, < –66 dBc (typ.)
	12.75 GHz < f ≤ 25.5 GHz	< –46 dBc, < –60 dBc (typ.)
	25.5 GHz < f ≤ 40 GHz	< –40 dBc, < –54 dBc (typ.)

Fig. 1-53: Example data sheet specifications for non-harmonics.

To assess the spectral purity of a generator signal, not only the phase noise, harmonics and non-harmonics must be specified, but also the subharmonics. The frequencies of the subharmonics have an integer attenuation factor with respect to the carrier frequency:

- First subharmonic $1/2 \cdot f_c$
- Second subharmonic $1/3 \cdot f_c$
- Nth subharmonic $f_c/(n+1)$

Subharmonic frequency components arise in the signal generator exclusively through the use of frequency multipliers. The figure below shows an example of the data sheet specifications for subharmonics. The carrier output power reference is required as well as the comparison value in dBc. The signal generator in the example below does not use a frequency multiplier below 6.375 GHz, and no subharmonics occur in this range.

Subharmonics	level > –10 dBm (level > 0 dBm for instruments without step attenuator)	
	f < 6.375 GHz	none
	6.375 GHz < f ≤ 20 GHz	< –55 dBc
	20 GHz < f ≤ 40 GHz	< –50 dBc

Fig. 1-54: Example data sheet with subharmonics specifications.

1.3.3 Frequency Stability

When characterizing frequency stability, a distinction must be made between long-term stability and short-term stability.

Short-term stability is defined by statistical or deterministic fluctuations in frequency around an average value. The spectral line of a single frequency signal is expanded under its influence. A distinction must be made here between two types of frequency fluctuations, random and deterministic fluctuations. Deterministic (or systematic, periodic) fluctuations can be traced back to discrete modulation frequencies. In other words, they appear as discrete but adjacent lines close to the average frequency and can be tracked back to known causes, such as AC hum or intermodulation products. Random (or statistical) frequency or phase fluctuations are known as phase noise (see 1.3.1 Phase Noise).

Long-term stability refers to a change in the average frequency, caused by slow changes in the components that determine the frequency (oscillators). For example, the expansion of a resonant structure due to temperature can cause a change in frequency. Long-term stability is often described by a relative frequency change per time unit. The specification of long-term stability is composed of several individual parameters (see Fig. 1-55).

Reference frequency

Frequency error	at time of calibration in production	$< 1 \times 10^{-7}$
	with R&S®SMB-B1/R&S®SMB-B1H option	$< 1 \times 10^{-8}$
Aging (after 10 days of uninterrupted operation)	standard	$< 1 \times 10^{-6}$/year
	with R&S®SMB-B1 option	$< 1 \times 10^{-9}$/day, $< 1 \times 10^{-7}$/year
	with R&S®SMB-B1H option	$< 5 \times 10^{-10}$/day, $< 3 \times 10^{-8}$/year
Temperature effect (0 °C to +50 °C)	standard	$< 2 \times 10^{-6}$
	with R&S®SMB-B1 option	$< 1 \times 10^{-7}$
	with R&S®SMB-B1H option	$< 1 \times 10^{-8}$
Warm-up time	to nominal thermostat temperature with R&S®SMB-B1/R&S®SMB-B1H option	≤ 10 min

Fig. 1-55: Frequency stability specification.

- The term "frequency error" defines the accuracy of the reference oscillator at the time of production or after calibration.
- Because the reference oscillator is subject to aging, specifications list the resulting frequency drift per day or per year. From this specification, it can be calculated when the oscillator must be recalibrated from a service-department. The time interval for the recalibration is determined based on the permissible frequency drift for a given application.
- In addition to the deviations discussed above, the total frequency inaccuracy calculation must also must include the deviation due to temperature.

As described in section 1.2.1.1 under "Synthesizer", the frequency accuracy of a signal generator is dependent on the frequency oscillator being used. A TCXO is typically used as the reference oscillator for most signal generators. For applications demanding greater frequency accuracy, signal generators can usually also be equipped with an optional OCXO. This is why the specification for the frequency stability also includes the values for the OCXO (Fig. 1-55, SMB®-B1: OCXO reference oscillator; SMB®-B1H: high-performance reference oscillator). Beyond the advantage of higher frequency accuracy, an OCXO also provides a shorter warm-up time. The specifications provided in the data sheet apply only after this time period. In this example, the warm-up time with the standard frequency oscillator is 30 min. With the OCXO, the warm-up time is reduced to 10 min. (Fig. 1-55).

1.3.4 Level Accuracy and Level Setting Time

As described in section 1.2.1.1, level accuracy and level setting time depend on the synthesizer model being used, the type of step attenuator and whether ALC is used. Fig. 1-1 provides example level error specifications from a data sheet for various frequency domains. Due to the high degree of precision for a mechanical step attenuator and the similarly improved matching of the RF output jack, the level error is less than that for a signal generator without a step attenuator.

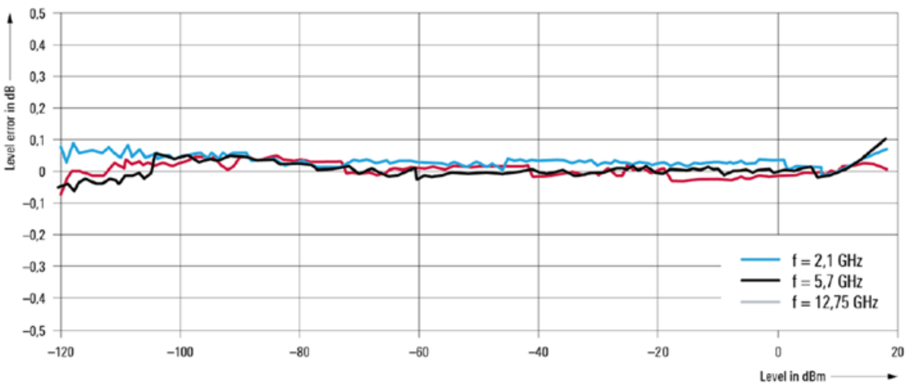
No step attenuator

With mechanical step attenuator

Level error	ALC state on, temperature range +18 °C to +33 °C		
R&S®SMB-B101/-B102/-B103/-B106/-B112	9 kHz ≤ f ≤ 200 kHz [4]	< 1.0 dB	
	200 kHz < f ≤ 3 GHz	< 0.5 dB	
	f > 3 GHz	< 0.9 dB	
R&S®SMB-B112L	200 kHz < f ≤ 3 GHz	< 0.7 dB	
	f > 3 GHz	< 1.1 dB	
R&S®SMB-B120L/-B140L	200 kHz < f ≤ 3 GHz	< 0.7 dB	
	3 GHz < f ≤ 20 GHz	< 1.1 dB	
	20 GHz < f ≤ 40 GHz	< 1.2 dB	
R&S®SMB-B120/-B140		level > –90 dBm	level ≤ –90 dBm
	200 kHz < f ≤ 3 GHz	< 0.5 dB	< 0.5 dB
	3 GHz < f ≤ 20 GHz	< 0.9 dB	< 1.2 dB
	20 GHz < f ≤ 40 GHz	< 1.0 dB	< 1.5 dB

Fig. 1-56: Specification of level error with mechanical step attenuator and without step attenuator.

As described in section 1.2.1.1 under "Level attenuation using step attenuators", when an electronic step attenuator is used, a power amplifier is connected downstream in order to compensate the attenuation. This leads to a drop in the spectral purity and to an increase in the level error. The diagrams in Fig. 1-57 and Fig. 1-58 show the level error for various frequencies based on the output level. Fig. 1-58 shows the negative effect on the level accuracy when a power amplifier is switched on.



Level linearity with R&S®SMB-B112 option, ALC on (meas.).

Fig. 1-57: Level error with electronic step attenuator at different frequencies.

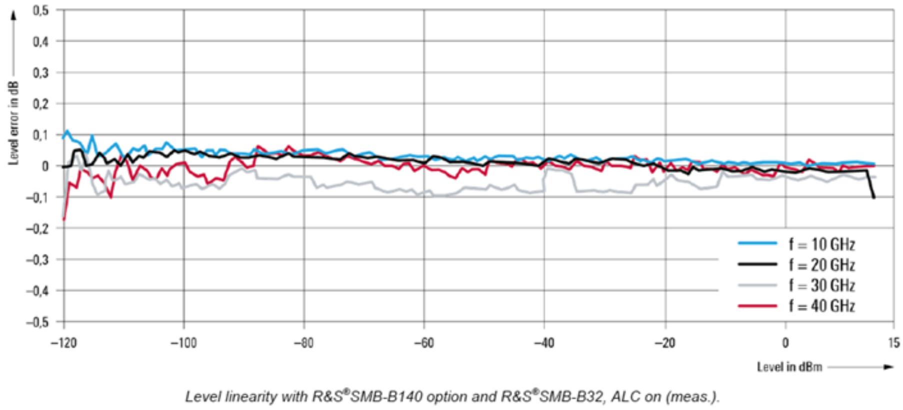


Fig. 1-58: Level error with electronic step attenuator and power amplifier at different frequencies.

Fig. 1-59 shows an example specification for level setting time. If it becomes necessary to add a mechanical step attenuator, the setting time increases considerably.

Level setting times

Setting time	level deviation < 0.1 dB[5] from final value, with GUI update stopped, temperature range +18 °C to +33 °C, without switching of the mechanical step attenuator	
	after IEC/IEEE bus delimiter	
	ALC state on	< 2.5 ms
	with switching of the mechanical step attenuator	
	ALC state on	< 25 ms

Fig. 1-59: Specification of the level setting time.

1.3.5 Error Vector Magnitude (EVM)

An important characteristic for the modulation quality of a vector signal generator is the error vector magnitude (EVM). The EVM is a measure of how much the generated symbols deviate from the ideal constellation (see Fig. 1-60) and corresponds to the relationship between the error vector V_{error} and the ideal reference vector V_{ref}.

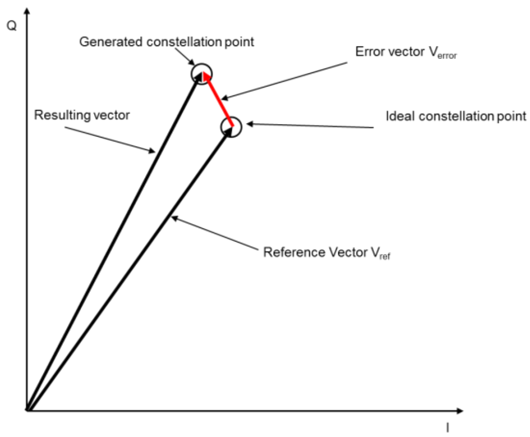


Fig. 1-60: Modulation errors due to deviation of the symbols from the ideal constellation.

The error vector V_{error} in the complex I/Q plane extends from the ideal constellation point to the (real) constellation point generated by the signal generator. The reference vector V_{ref} extends from zero to the ideal constellation point.

Typically, the EVM value is stated in decibels (dB) or as a percentage (Equation 1-13 and Equation 1-14).

Equation 1-13:

$$EVM(dB) = 10 \log \left(\frac{V_{error}}{V_{ref}} \right)$$

Equation 1-14:

$$EVM(\%) = 100\ \% \cdot \sqrt{\frac{V_{error}}{V_{ref}}}$$

Higher-order modulation modes such as 16QAM (see section 2.4.4) with narrower symbol spacing react with more sensitivity to vector errors than simpler modulation modes with wider symbol spacing. At the left of Fig. 1-61 is the constellation diagram of a QPSK signal shifted by 45° in a signal-to-noise radio (carrier / noise, or C / N) of 20 dB. In this example, the EVM value is 26 %. In spite of the high EVM, the large symbol spacing makes symbol recognition possible without problem. In contrast, even with the same C / N, only a partially correct symbol allocation is possible for the 16QAM signal at the right of Fig. 1-61, leading to a higher bit error probability.

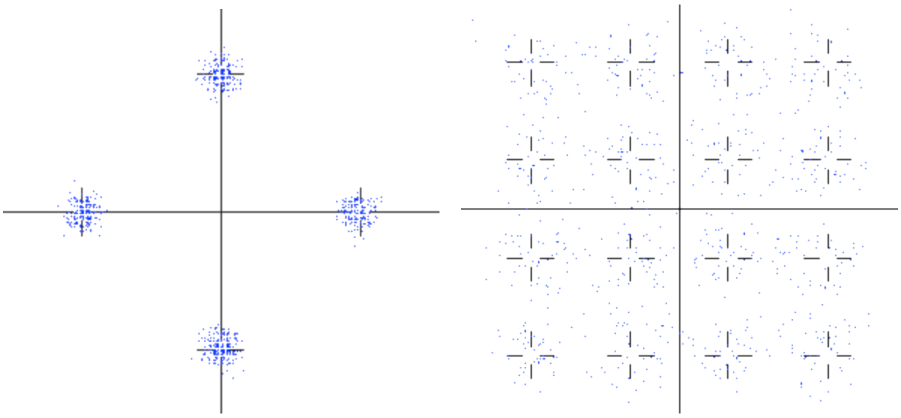


Fig. 1-61: Constellation diagram for QPSK and 16QAM with equivalent C/N.

Although the signal-to-noise ratio is the same for the two signals in the left and right figures, the 47.9 % EVM for 16QAM is significantly higher than that for QPSK. This is because the average vector length of the reference vector referenced by the error vector during the calculation is significantly smaller for 16QAM than for QPSK.

In practice, the receiver is assessed by measuring the demodulation quality, and thus also the EVM. In order to keep the influence of the generated signal on the measurement result to a minimum, it is therefore important that the signal generator produce a signal with a high modulation quality, i.e. a low EVM. The EVM of a vector signal generator depends on the baseband filtering, modulation mode, symbol rate and, for mobile radio signals, also on the selected mobile cellular standard (see also excerpt from a data sheet in Fig. 1-62).

Modulation performance for main digital standards

Measured values except otherwise stated.

Standard	GSM	EDGE	WCDMA 3GPP		CDMA2000®	IEEE 802.11a/g	IEEE 802.11ac	WiMAX™	LTE
			1DPCH	TM1-64				BW = 10 MHz	
Frequency	400 MHz to 2000 MHz	400 MHz to 2000 MHz	1800 MHz to 2200 MHz	1800 MHz to 2200 MHz	800 MHz	2400 MHz to 2485 MHz; 5150 MHz to 5825 MHz	2400 MHz to 2485 MHz; 5150 MHz to 5825 MHz	5000 MHz	1800 MHz to 2200 MHz
EVM	–	0.25 % (typ.)	0.4 % (typ.)	0.4 %	0.4 %	0.6 %	0.44 %	0.4 %	0.4 %

Modulation performance for custom digital modulation

Deviation error with 2FSK, 4FSK	deviation 0.2 to 0.7 × symbol rate Gaussian filter with B × T = 0.2 to 0.7	
	symbol rate up to 2 MHz	0.4 % (meas.)
	symbol rate up to 10 MHz	1.2 % (meas.)
Phase error with MSK	Gaussian filter with B × T = 0.2 to 0.7	
	bit rate up to 10 MHz	0.3° (meas.)
EVM with QPSK, OQPSK, π/4-DQPSK, 8PSK, 16QAM, 32QAM, 64QAM	cosine, root cosine filter with α = 0.2 to 0.7	
	symbol rate up to 5 MHz	0.5 % (meas.)
	symbol rate up to 20 MHz	2.0 % (meas.)

Fig. 1-62: Example of an EVM specification.

1.3.6 Adjacent Channel Power (ACP)

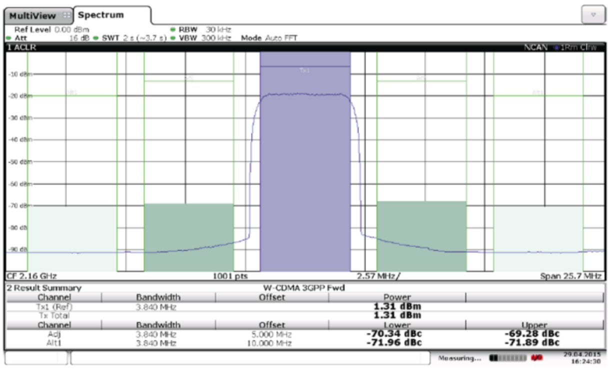
In order to ensure that transmitted information is received without interference by as many subscribers as possible, the adjacent channels in the frequency domain must always remain free of interference. One measure for this interference is the adjacent channel power, which should remain as low as possible and which is stated either as an absolute value (in dBm) or as a relative value referenced to the channel power in the transmission channel. The adjacent channel power (ACP) is also known as adjacent channel leakage power ratio (ACLR) or the adjacent channel power ratio (ACPR). The unwanted spurious emissions are primarily caused by amplifiers and their third-order intermodulation products in the first adjacent channel as well as by fifth-order intermodulation products in the second adjacent channel (the alternate channel). In order to test the amplifiers for generated disturbance power in the adjacent channel, it is important that the test signal produced by the vector signal generator itself have a very high adjacent channel power suppression. The ACPR value is therefore another important quality characteristic for a vector signal generator. As seen in the data sheet excerpt in the figure below, this value is also dependent on the generated signal and on the selected standard.

Modulation performance for main digital standards

Measured values except otherwise stated.

Standard	GSM	EDGE	WCDMA 3GPP		CDMA2000®	IEEE 802.11a/g	IEEE 802.11ac
			1DPCH	TM1-64			
Frequency	400 MHz to 2000 MHz	400 MHz to 2000 MHz	1800 MHz to 2200 MHz	1800 MHz to 2200 MHz	800 MHz	2400 MHz to 2485 MHz; 5150 MHz to 5825 MHz	2400 MHz to 2485 MHz; 5150 MHz to 5825 MHz
Adjacent channel power ratio (ACPR) in dB							
Channel spacing	200 kHz	200 kHz	5 MHz	5 MHz	30 kHz	20 MHz	160 MHz
In adjacent channel	–38	–38	–69	–67 (typ.)	–79 at 0.75 MHz	–42	–50
In alternate channel	–70	–70	–74	–71 (typ.)	–91 at 1.98 MHz	–55	–56
In 2nd alternate channel	–78	–78	–	–	–	–56	–56

Fig. 1-63: ACPR specification based on the standard.



Digital standard 3GPP FDD test model 1, 64 DPCHs ACLR (meas.).

Fig. 1-64: ACPR specification based on the standard.

1.3.7 Impedance Matching (VSWR)

Signal generators for the most part are designed for 50 Ω systems. To ensure that as much of the generated power can be used by the connected load or by the DUT as possible, the signal generator output and the load input must be as close to 50 ohms as possible. A good impedance matching is ensured only if no signal reflection occurs and if the mismatch uncertainty is kept to a minimum.

The mismatch uncertainty ε is calculated as follows:

Equation 1-15:

$$\varepsilon = \pm 20\,\lg(1 + r_G \cdot r_L)\ \mathrm{dB}$$

where

r_g : Generator reflection coefficient

r_L: Load reflection coefficient

Definition of the reflection coefficient:

Any kind of RF power transmission takes place between a source and a load. The source and load are connected via a standard transmission line, for example a section of coaxial cable with the characteristic impedance Z_0 (Fig. 1-65).

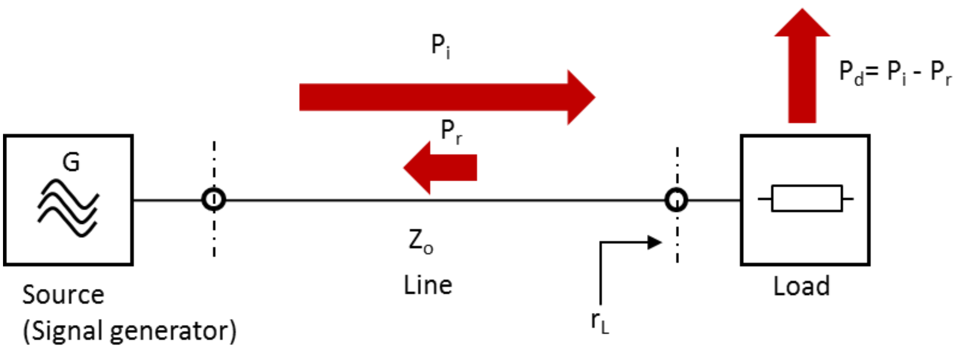


Fig. 1-65: Power flow between source and load.

The source supplies a sine-wave signal with a constant amplitude. The line is assumed to be lossless. Once all transients have dropped off, two stationary waves remain. One flows from the source to the load (forward direction) and the other flows in the opposite direction (reverse direction). The two waves carry the incident power P_i and the typically smaller reflected power P_r. The ratio of P_r/P_i depends solely on how well the load is matched to the line.

The measure for the load matching is the reflection coefficient r_L :

Equation 1-16:

$$r_L = \sqrt{\frac{P_r}{P_i}}$$

In an ideal impedance matching, no power is reflected from the load ($P_r = 0$). It follows that $r_L = 0$. If all of the power is reflected ($P_r = P_i$), then $r_L = 1$. In other words, the values for the reflection coefficient always lie between 0 and 1. For most calculations, it is sufficient to use only the magnitude of the reflection coefficient. At other times, the phase angle Φ_L must additionally be specified. The magnitude r_L and phase Φ_L can then be combined into a complex reflection coefficient Γ_L.

The reflection coefficient can also be specified as return loss a_r (Equation 1-17, Fig. 1-66). This indicates, stated in dB, by how much the reflected power is attenuated as compared to the forward power.

Equation 1-17:

$$a_r = 20 \lg \frac{1}{r_L} dB$$

As seen in Fig. Fig. 1-65, the load receives only the net difference between the forward power and the reflected power. It is therefore also known as the net power or the absorbed power P_d:

Equation 1-18:

$$P_d = P_i + P_r$$

The specifications for a signal generator output typically include the impedance matching but not the reflection coefficient. Another typical measure for the impedance matching is the voltage standing wave ratio (**VSWR**).

Definition of VSWR:

The interference pattern resulting from the superposition of forward and reflected waves is called a standing wave. It is periodic with one half a line wavelength. At the positions where the voltage and current of both waves are either in phase or out of phase, maxima and minima values are produced. The ratio of maximum to minimum values is a measure of the magnitude of the load-side reflection coefficient. This ripple s and the extreme values can be very precisely determined on transmission lines. More commonly used than the term "ripple" is the term voltage standing wave ratio (VSWR), or simply SWR. Unlike the reflection coefficient, the VSWR is calculated not from the power ratio, but rather from the maximum and minimum voltage values of the standing wave. The following definition applies for s, or VSWR:

Equation 1-19:

$$s = VSWR = SWR = \frac{V_{max}}{V_{min}}$$

The VSWR and the reflection coefficient have the following relationship:

Equation 1-20:

$$VSWR = \frac{1 + r_L}{1 - r_L}$$

For impedance matching where $r_L = 0$, the result is VSWR = 1. For total reflection where $r_L = 1$, the result is VSWR = ∞. The value range for VSWR therefore lies between 0 and ∞ (see also Fig. 1-66).

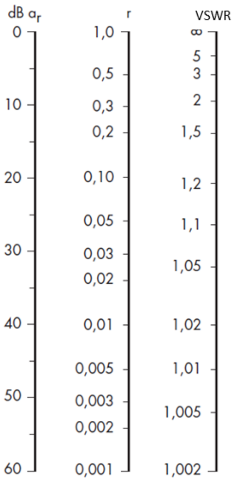


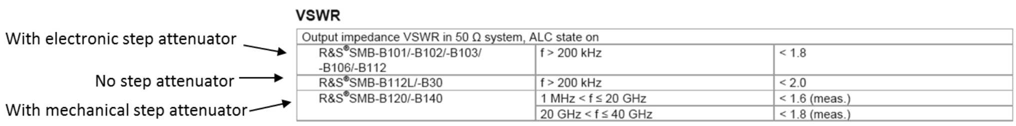
Fig. 1-66: Conversion table for reflection coefficient r, return loss a_r and standing wave ratio VSWR.

The abbreviation VSWR is so widely used that it has also come to describe the source-side impedance matching. However, that is not to say that the concept of replacing r_L with r_G in Equation 1-20 is merely semantics, with no foundation in physical science. The following equation can also be used for the VSWR of a signal generator:

Equation 1-21:

$$VSWR = \frac{1 + r_G}{1 - r_G}$$

Because the type of step attenuators used in the signal generator can have an effect on the generator matching, the VSWR value is typically specified for several different instrument configurations. Fig. 1-67 provides an example from a data sheet for a signal generator.



VSWR

Output impedance VSWR in 50 Ω system, ALC state on		
R&S®SMB-B101/-B102/-B103/ -B106/-B112	f > 200 kHz	< 1.8
R&S®SMB-B112L/-B30	f > 200 kHz	< 2.0
R&S®SMB-B120/-B140	1 MHz < f ≤ 20 GHz	< 1.6 (meas.)
	20 GHz < f ≤ 40 GHz	< 1.8 (meas.)

Fig. 1-67: Excerpt from the R&S® SMB100A signal generator data sheet.

In the following example, the mismatch uncertainty of a signal generator is calculated with a connected load, or DUT.

Impedance matching for the generator : $VSWR_G$= 1.6 (R&S® SMB100A up to 20 GHz)

Impedance matching for the load: $VSWR_L$= 1.7

To make it possible to calculate the mismatch uncertainty using Equation 1-15, the reflection coefficients r_G and r_L are calculated from the known VSWR as follows after solving Equation 1-20 and

Equation 1-21 using r_L and r_G:

$$r_G = \frac{VSWR_G - 1}{VSWR_G + 1} = \frac{1.6 - 1}{1.6 + 1} = 0.23$$

$$r_G = \frac{VSWR_L - 1}{VSWR_L + 1} = \frac{1.7 - 1}{1.7 + 1} = 0.26$$

The mismatch uncertainty is calculated as follows:

$$\varepsilon = \pm 20\,\lg(\,1 + \mathrm{r_G} \cdot \mathrm{r_L})\ \mathrm{dB} = \pm 20\,\lg(1 + 0.23 \cdot 0.26)\ \mathrm{dB} = \pm 0.5\ \mathrm{dB}$$

For example, for a generator power of +10 dBm (10 mW), the power at the load lies in the range of 9.5 dBm (8.91 mW) and 10.5 dBm (11.22 mW) as a result of the mismatch uncertainty. If the mismatch uncertainty for a certain application is still too high, it can be reduced by inserting an attenuator (Fig. 1-68). Of course, this makes sense only if sufficient power remains to meet the requirements at the load.

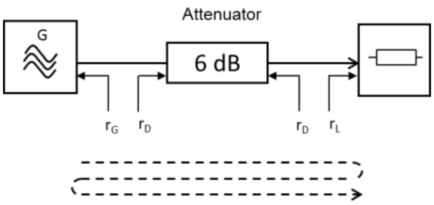


Fig. 1-68: Improving the generator matching with an attenuator.

In the example above, a 6 dB attenuator is inserted for the mismatch uncertainty:

$$\varepsilon = 20\,\lg(1 + \mathrm{r_G} \cdot \mathrm{r_L} \cdot r_D^2)\mathrm{dB} = 20\,\lg(1 + 0.23 \cdot 0.26 \cdot 0.5^2)\ \mathrm{dB} = 0.13\ \mathrm{dB}$$

where

$$\mathrm{r_D} = 10^{\frac{\mathrm{a_D}}{20\mathrm{dB}}} = 10^{\frac{-6\ \mathrm{dB}}{20\ \mathrm{dB}}} = 0.5$$ (from Equation 1-17)

The power at the load lies in the range of 9.87 dBm (9.71 mW) and 10.13 dBm (10.30 mW) as a result of the mismatch uncertainty.

The diagram below (Fig. 1-69) can be used to quickly determine the expected mismatch uncertainty as a %.

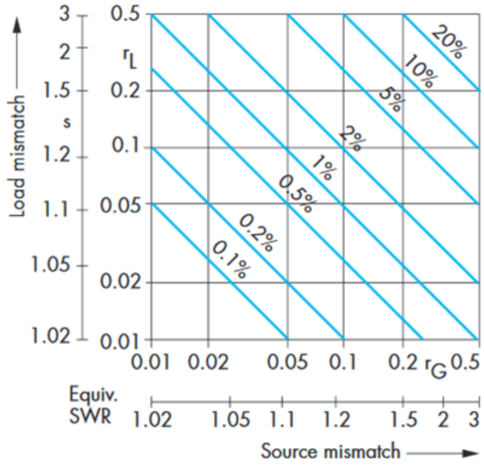


Fig. 1-69: Mismatch uncertainty for the forward power and for the power available at Z_0.

2 Modulation Methods

In order to transmit electrical information signals efficiently and with as little distortion as possible, the signals must be adapted to the physical characteristics of the transmission channel. To do this, the amplitude, frequency or phase angle of a carrier signal is modified by the wanted signal (voice, music, data) to permit efficient utilization of the transmission channel. This process is known as modulation. Within the carrier frequency range, the transmit signal occupies a bandwidth that is dependent on the wanted signal. The carrier signal itself has no significance with respect to the transmitted message and can be suppressed for certain modulation modes. Because only high-frequency signals can be radiated as electromagnetic waves over an antenna, the wireless transmission of signals requires a modulated, high-frequency carrier. On the receiver end, the wanted signal is restored by means of a demodulator (Fig. 2-1). The transmission channel can involve a conducted transmission over electrical or fiber-optic cables, or it can involve a radio link via antennas.

Fig. 2-1: Signal path during the transmission of information using modulated signals.

Continuous-time modulation methods use a continuous signal, such as a sine-wave oscillation, as the carrier. The two varieties of continuous-time methods are continuous value modulation and discrete value modulation. The continuous-value influence of a time-continuous carrier is better known as analog modulation, while a discrete-value influence of the carrier is known as digital modulation.

To better utilize the transmission channel, wanted signals are transmitted in parallel over a common channel by means of multiplexing. The practical implementation of the various multiplexing methods, including time-division multiplexing, frequency-division multiplexing and code-division multiplexing, is likewise achieved through the use of the appropriate modulation methods.

2.1 Analog Modulation Methods

Analog modulations continuously process analog wanted signals such as voice, music or video signals without digitizing the transmission signal values. There are two main types of analog modulation, amplitude modulation and angle modulation. All other analog modulation methods are derived from these two modulation techniques. Table 2-1 shows the most important types of analog modulation.

Type of modulation	Abbreviation
Amplitude modulation	AM
Single-sideband modulation	SSB-AM
Vestigial-sideband modulation	VSB-AM
Double-sideband suppressed carrier	DSBSC-AM
Frequency modulation	FM
Phase modulation	PM

Table 2-1: Summary of the most important types of analog modulation

2.1.1 Amplitude Modulation

Amplitude modulation (AM) continuously applies the information from the wanted signal to the amplitude oscillations of the transmit signal. In the process, the low-frequency wanted signal from the baseband is converted into a high frequency range, which causes the appearance of new frequency components. These result from the product of the modulated wanted signal and the carrier oscillation. In the case of a linear amplitude modulation, the amplitude of the carrier oscillation changes linearly with the value of the wanted signal, so only first-order components appear.

Important applications for amplitude modulation include:

- Radio broadcasting across various frequency bands (long-wave, medium-wave, short-wave)
- Television broadcasting, depending on the broadcast standard
- Amateur radio (mostly in a modified format as single-sideband modulation)
- Flight navigation (ADF and VOR)
- Flight communications

2.1.2 Time Function and Frequency Spectrum of Amplitude Modulation

Fig. 2-2 shows the concept behind AM modulation. The wanted signal $v_s(t)$ with superimposed direct voltage V_0 as well as the carrier signal $v_T(t)$ are applied to an amplitude modulator. The modulation product $v_{AM}(t)$ appears at the output of the AM modulator. After modulation, the DC component (V_0) then represents the carrier signal component in the modulated signal.

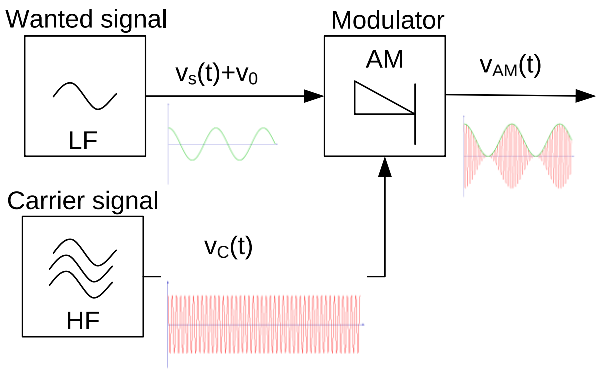


Fig. 2-2: Amplitude modulator.

The modulator uses the following transfer function:

Equation 2-1:

$$v_M(t) = v_{AM}(t) = (v_s(t) + V_0) \cdot \frac{v_c(t)}{\hat{v}_c(t)}$$

The influence of the carrier amplitude changes the constant amplitude $\hat{v}_c$ of the carrier oscillation into a time-dependent amplitude $\hat{v}_{AM}(t)$. The zero reference value $\hat{v}_c$ is defined by the DC component V_o. The maximum amplitude deviation $\Delta\hat{v}_c$ is proportionate to the amplitude $\hat{v}_s$ of the modulating signal. The influence exerted by the amplitude of the carrier oscillation can be stated as follows:

Equation 2-2:

$$\hat{v}_{AM}(t) = \hat{v}_c + \Delta\hat{v}_c \cdot cos\omega_s t$$

where:

Angular frequency: $\omega_s = 2\pi f_s$

The modulation depth m specifies how strongly the modulating wanted signal affects the amplitude of the modulated carrier signal and is defined as a relative change in the carrier amplitude (Equation 2-3).

Equation 2-3:

$$m = \frac{\Delta\hat{v}_c}{\hat{v}_c}$$

Equation 2-2 can also be written as:

Equation 2-4:

$$\hat{v}_{AM}(t) = \hat{v}_c + (1 + m \cdot cos\omega_s t)$$

The following equation states the time function of the amplitude-modulated oscillation:

Equation 2-5:

$$v_{AM}(t) = \hat{v}_{AM}(t) \cdot cos\omega_c t = \hat{v}_c \cdot (1 + m \cdot cos\omega_s t) \cdot \ cos\omega_s t$$

Equation 2-5 results in the time characteristic of a carrier signal that is modulated by a wanted sine-wave signal with a modulation depth of 60 % as shown in Fig. 2-3.

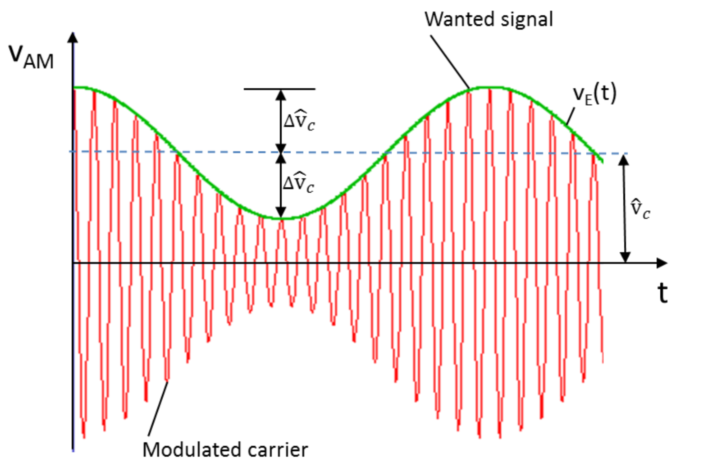


Fig. 2-3: Time characteristic of an amplitude-modulated signal.

The modulation depth typically lies between 0 and 1. At m = 0, the carrier oscillation is unmodulated and only the carrier is transmitted. At m = 1, there is a 100 % amplitude modulation. The limiting envelopes of the wanted signals are based on their trough values. An overmodulation is indicated if m > 1. This happens if a higher signal amplitude is present at the modulator than required for m = 1. In this case, phase shifts occur that produce a distorted wanted signal. To prevent this, the amplitude of the modulating signal is limited before the modulation.

By accepting $\cos\alpha \cdot \cos\beta = \frac{1}{2}\cos(\alpha+\beta) + \frac{1}{2}\cos(\alpha-\beta)$, it becomes possible to mathematically transform the above equation as follows:

Equation 2-6:

$$v_{AM}(t) = \underbrace{\hat{v}_c \cdot cos\omega_c t}_{\text{Carrier oscillation}} + \underbrace{\hat{v}_c \frac{m}{2} \cdot \cos(\omega_c + \omega_s)\, t}_{\text{Upper sideband oscillation}} + \underbrace{\hat{v}_c \frac{m}{2} \cdot \cos(\omega_c - \omega_s)\, t}_{\text{Lower sideband oscillation}}$$

From Equation 2-6, it is clear that the modulation product includes not only the carrier oscillation with frequency f_c, but also other oscillations at the frequencies f_c+f_s and $f_c–f_s$. The amplitude of the two sideband modulations is dependent on the modulation depth m; at 100 % AM modulation (m = 1), it is one-half the value of the carrier amplitude $\hat{v}_c$. For an amplitude-modulated oscillation, this results in the following figure in the frequency domain:

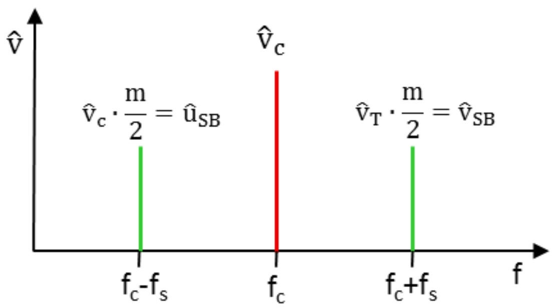


Fig. 2-4: Frequency spectrum of an amplitude modulation.

As seen in Fig. 2-4, the lower and upper sideband oscillations lie in symmetry to the carrier frequency. Because in practice the wanted signal – e.g. music or voice – consists of a frequency band ($f_{s\,min}$ to $f_{s\,max}$) instead of a single frequency, the carrier signal has an upper sideband (USB) and a lower sideband (LSB) placed symmetrically to either side of it (see Fig. 2-5). The bandwidth required to transmit the AM signal is:

Equation 2-7:

$$B_{AM} = 2 \cdot f_{s\,max}$$

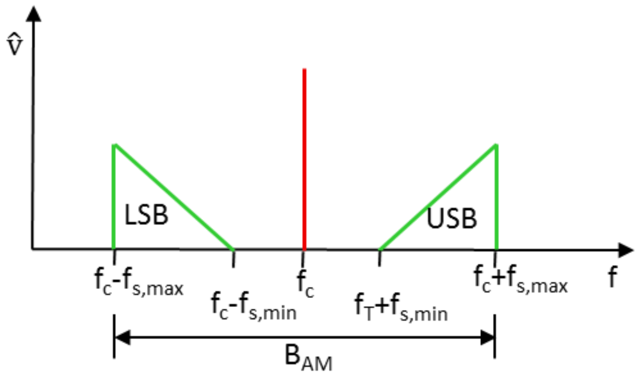


Fig. 2-5: Sidebands of an amplitude-modulated signal.

In practice, the modulation depth can be determined from the time characteristic or the frequency spectrum. To do this, an oscilloscope is used to determine the peak and trough values of the envelope in the time domain (Fig. 2-6), and from these the modulation depth is calculated using the following equation:

Equation 2-8:

$$m = \frac{A_{max} - A_{min}}{A_{max} + A_{min}}$$

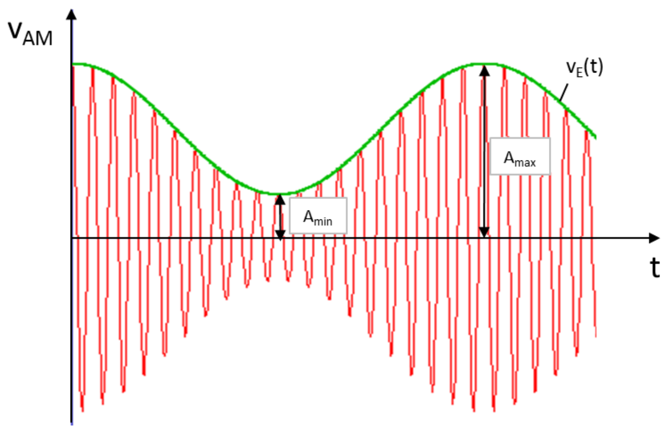


Fig. 2-6: Maximum and minimum values for an amplitude-modulated signal.

In the frequency domain, the modulation depth can be calculated as follows using a spectrum analyzer and the ratio of sideband oscillations to carrier amplitude (see also Fig. 2-4):

Equation 2-9:

$$m = 2 \cdot \frac{\Delta \hat{v}_{SB}}{\hat{v}_c}$$

Usually a spectrum analyzer displays the signal amplitudes on a logarithmic scale (Fig. 2-7). From the delta ΔL between the carrier level L_c and the sideband oscillation level L_{SB}, the modulation depth can be calculated using Equation 2-10 without converting to linear values.

Equation 2-10:

$$m = 10^{\left(\frac{6\,dB - \Delta L}{20\,dB}\right)}$$

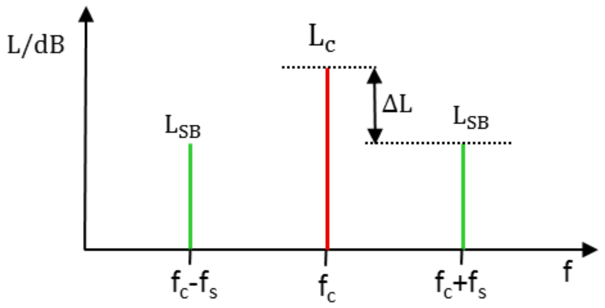


Fig. 2-7: Logarithmic frequency spectrum of an amplitude-modulated signal.

Another way to determine the modulation depth is to use the trapezoid method. This is done by operating the oscilloscope in XY mode. The horizontal deflection (X-deflection) of the modulating signal $v_s(t)$ and the vertical deflection (Y-deflection) of the amplitude-modulated signal $v_{AM}(t)$ are applied. Displayed as an oscillogram, for m < 1 a trapezoid results with sides $2 \cdot A_{max}$ and $2 \cdot A_{min}$. (Modulation depth calculated using Equation 2-8). For m = 1, the trapezoid changes to a triangle ($A_{min} = 0$). With overmodulation (m > 1), an extension of the triangle can be seen (Fig. 2-8).

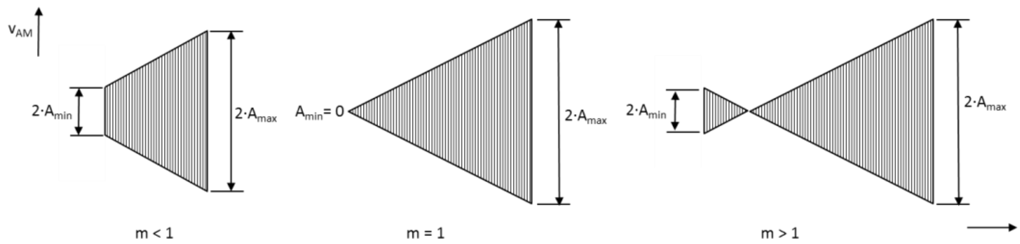


Fig. 2-8: Modulation trapezoid at various modulation depths.

If a phase shift lies between the modulating signal and the envelope, the upper and lower boundaries of the trapezoid are represented by an ellipse (Fig. 2-9).

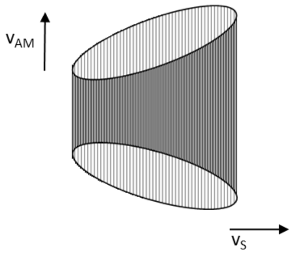


Fig. 2-9: Modulation trapezoid resulting from a phase shift between the envelope and the modulating wanted signal.

A modulation trapezoid can also provide qualitative information about the linearity of the amplitude modulation. To do this, the modulating signal of a second y-deflection is fed to the oscilloscope. When the amplitude is set appropriately, a sloping line results that in the case of a linear amplitude modulation coincides with the upper boundary of the modulation trapezoid. In the case of non-linear amplitude modulation, the deviation of the line from the upper boundary is clearly visible, especially with a large modulation depth (Fig. 2-10).

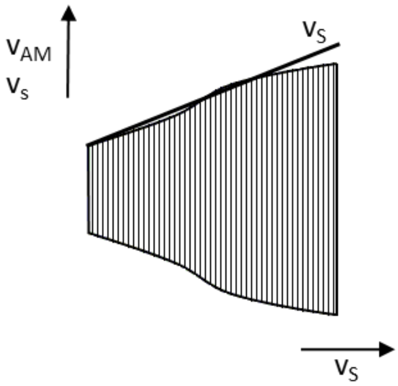


Fig. 2-10: Assessing linearity using a modulation trapezoid.

Complex display and vector diagram

The modulation product of an amplitude-modulated oscillation can also be displayed using a vector diagram. In complex notation, a modulation product can be described as follows:

Equation 2-11:

$$v_{AM}(t) = \underbrace{\hat{v}_c \cdot cos\omega_c t}_{\text{Carrier oscillation}} + \underbrace{\hat{v}_c \frac{m}{2} \cdot \cos(\omega_c + \omega_s)\,t}_{\text{Upper sideband oscillation}} + \underbrace{\hat{v}_c \frac{m}{2} \cdot \cos(\omega_c - \omega_s)\,t}_{\text{Lower sideband oscillation}}$$

As seen in Equation 2-11, the modulation product is the result of three vectors added together. If the three vectors are transferred to a diagram (Fig. 2-11), the vectors for the upper and lower sideband oscillations will lie in symmetry to the carrier vector. Like

for a parallelogram of forces, the resulting vector is composed of the two vectors from the sideband oscillation.
If the carrier vector is assumed to be fixed, then the two vectors for the upper and lower sideband oscillation will rotate in opposing directions at a frequency ω_S (Fig. 2-12). The resulting instantaneous amplitude of the modulated signal always lies in the same direction as the carrier vector, and is therefore in phase with the carrier signal. This means that the length of the resulting vector is identical to the instantaneous value of the envelope $v_E(t)$. The vector diagram also shows how the amplitude of the high-frequency sum signal, consisting of the carrier frequency and sidebands, changes with the rhythm of the modulation. The amplitude of the carrier $\hat{v}_c$ remains constant. This can be proven either with a spectrum analyzer in which the resolution bandwidth is set to be narrower than the modulation bandwidth of the AM signal, or with a narrow bandpass that blocks the modulation.

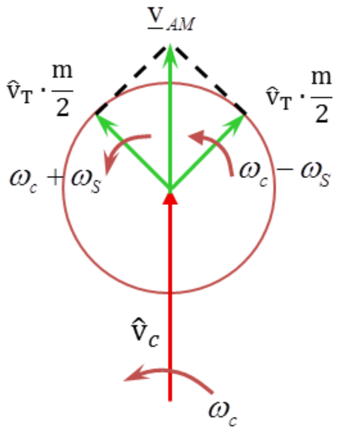


Fig. 2-11: Vector diagram of an amplitude-modulated oscillation with rotating carrier vector.

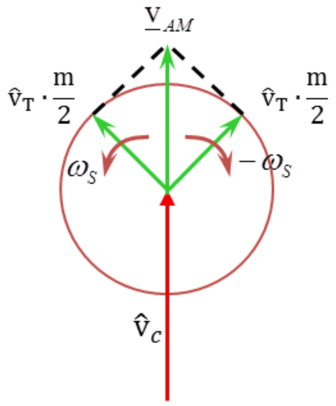


Fig. 2-12: Vector diagram of an amplitude-modulated oscillation with fixed carrier vector.

2.1.3 Types of Amplitude Modulation

As described in section 2.1.2, an amplitude-modulated oscillation is made up of the carrier oscillation and two sidebands for the wanted signal. The disadvantage of this arrangement is that the carrier itself does not contain any information; instead, the information, or wanted signal, is divided equally onto two sidebands. This has clear disadvantages, in that the AM modulation requires a high output power (carrier + two sidebands) as well as a large bandwidth for a redundant sideband. The division of the power to the two sidebands alone reduces the power of each sideband by about 50 %. These disadvantages are compensated by special variations in the amplitude modulation, such as amplitude modulation with suppressed carrier, single sideband modulation (SSB) or vestigial sideband modulation. Applications for amplitude modulation include analog radio broadcasting via medium wave and analog TV broadcasting. Single sideband modulation requires more technical effort, but it uses the frequency band more efficiently; example uses include short-wave applications (maritime communications, flight communciations, amateur radio).

2.1.3.1 Dual Sideband Modulation with Carrier Suppression

The power for an amplitude-modulated oscillation is derived from the frequency spectrum (Fig. 2-5) and is equal to:

Equation 2-12:

$$P_{AM} = P_c + P_{LSB} + P_{USB} = P_c \cdot (1 + \frac{m^2}{2})$$

In the case of an unmodulated signal (m = 0), the power of the amplitude-modulated oscillation P_{AM} is equal to the power of the carrier P_c. At 100 % AM-modulation (m = 1), it follows that the carrier power is at least 66 % of the total power. A reduction in the transmitter power by up to two-thirds would therefore require a complete or partial suppression of the carrier signal. The carrier signal must be applied again as a reference value for demodulation in the receiver.

To suppress the carrier, the modulator transfer function (Equation 2-1) uses a value of zero for the DC component V_0. In the modulation product, this causes the carrier components to disappear, leaving a dual sideband amplitude modulation without carrier, also known as dual sideband modulation (DSB-AM).

The following equation therefore results for the time function of a dual sideband modulation:

Equation 2-13:

$$v_{DSB}(t) = \hat{v}_{SB}(t) \cdot \cos(\omega_c + \omega_s)t + \cos(\omega_c - \omega_s)t$$

or, in complex notation:

Equation 2-14:

$$\underline{v}_{DSB}(t) = \hat{v}_{SB} \cdot e^{j(\omega_c+\omega_s)t} + \hat{v}_{SB} \cdot e^{j(\omega_c-\omega_s)t}$$

By suppressing the constant carrier component $\hat{v}_c$, the envelope $u_E(t)$ experiences a phase shift of 180° at the zero crossing. The envelope now consists of half-sine waves with a repetition frequency that is double the signal frequency.

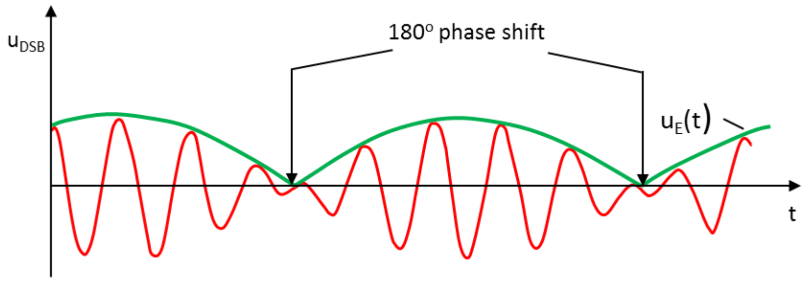


Fig. 2-13: Time characteristic of a dual sideband modulation with carrier suppression.

The occupied bandwidth of a DSBSC amplitude modulation is equal to that of an amplitude modulation without carrier suppression, and therefore is of little practical value. The modulation product of the DSB amplitude modulation is however used as an output quantity for single sideband modulation.

2.1.3.2 Single Sideband Modulation

In order to reduce the required bandwidth, only one sideband is used for single sideband amplitude modulation (SSB-AM). This is done by suppressing the second sideband. This is possible because both sidebands contain the same information. In addition to the bandwidth advantage, a SSB modulation can apply all of the transmit energy to the information contained in the signal. The result is an increased coverage and a greater signal-to-noise ratio at the same transmit power. In many cases, the transmission includes not only the sidebands, but also a residual carrier as a reference quantity for the demodulation. The highly accurate transmit and receive oscillators available today make this type of modulation of little significance, however.

2.1.3.3 Vestigial Sideband Modulation

When vestigial sideband modulation is used to transmit information signals with very small frequency components, a filter with a very steep edge slope must be used to suppress the unwanted sidebands. Apart from the high technical effort to implement, steep-edged filters have the disadvantage that they exhibit a strong group delay at the boundary to the passband. Pulsed signals in particular are strongly distorted during transmission. A vestigial sideband modulation (VSB-AM) can help. This is accomplished by using filters whose transfer function has a relatively flat falling edge. The ensures that a portion of the sideband that is to be suppressed (vestigial sideband) is included in the transmission (Fig. 2-14).

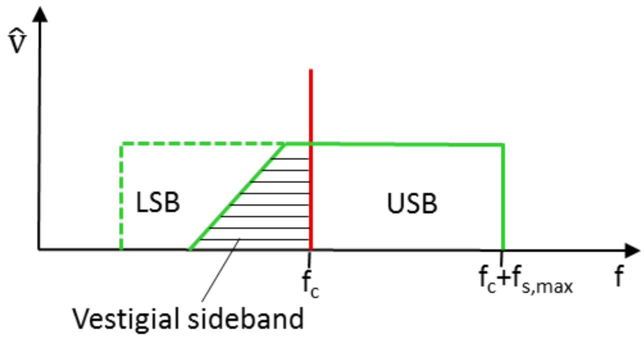


Fig. 2-14: Partial suppression of the lower sideband during vestigial sideband modulation.

Because VSB-AM causes the entire upper sideband and a portion of the lower sideband to be transmitted, the signal amplitudes of the low and high frequencies are significantly different after demodulation. To compensate, a filter with a Nyquist edge is used before the demodulation. The carrier frequency lies about halfway on the sloping filter edge (Fig. 2-15). At high signal frequencies, this filter provides less amplification within the lower frequency range on the upper sideband than it does at the upper frequencies. This dissipation is eliminated by the portion of the lower sideband that lies below the Nyquist edge. As a result, a signal with a compensated amplitude frequency response can be supplied to the demodulator. The restored information is for the most part free of linear distortion. One use for VSB-AM is transmitting vision carriers. The edge width of the Nyquist filter in a television receiver is 1.5 MHz.

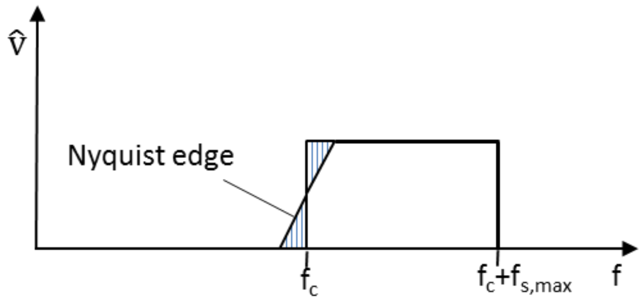


Fig. 2-15: Frequency response compensation using a filter with a Nyquist edge on the receive end.

2.1.4 Angle Modulation

Unlike amplitude modulation where the amplitude is modified, angle modulation modifies the phase angle of a carrier oscillation dependent on the wanted signal. Because this causes a simultaneous change in the carrier frequency, the angle modulation can be considered a type of phase modulation (PM) or frequency modulation. The distinctions are described in more detail in section 2.1.6. With both modulations, the amplitude of the carrier signal remains constant.

Angle modulations, of which the primary types are frequency modulation (FM) and phase modulation (PM), apply the wanted signal to the phase angle of the carrier signal. This causes a change in the carrier frequency or the phase angle of the carrier signal. Examples of applications for these technologies are found in analog VHF FM radio broadcasting.

2.1.5 Time Function and Frequency Spectrum

Fig. 2-16 shows the principles behind angle modulation. The wanted signal $v_s(t)$ with superimposed DC voltage V_0 and the carrier signal $v_c(t)$ are fed to an angle modulator.

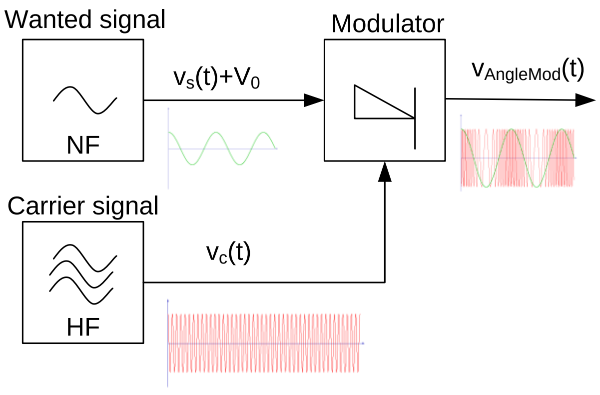


Fig. 2-16: Angle modulator.

The starting point for angle modulation is a sine-wave carrier oscillation:

Equation 2-15:

$$v_c(t) = \hat{v}_c \cdot cos\omega_c t = \hat{v}_c \cdot cos\varphi_c(t)$$

where $\varphi_c(t) = \omega_c \cdot t = 2\pi \cdot f_c \cdot t$

At constant carrier frequency f_c, the phase angle $\varphi_c(t)$ increases proportionately over time (Fig. 2-17a). An angle modulator influences the carrier phase angle using the wanted signal. In modulation product $v_{AngleMod}(t)$, the phase angle $\varphi_c(t)$ is converted to phase angle $\varphi_{AngleMod}(t)$, which is dependent on the wanted signal. This consists of a time-dependent component $\varphi_c(t)$ and the alternate component that is dependent on the wanted signal $\varphi_{c_a}(t)$ (Fig. 2-17b).

Equation 2-16:

$$\varphi_{AngleMod}(t) = \varphi_c(t) + \varphi_{c_a}(t) = \omega_c \cdot t + \Delta\varphi_c \cdot cos\omega_s t$$

$\Delta\varphi_c$ is the phase deviation. It is proportionate to the amplitude $\hat{v}_s$ of the wanted signal and represents the maximum change in the phase angle $\varphi_{AngleMod}(t)$ referenced to the carrier phase angle $\varphi_c(t)$.

The time function of an angle-modulated oscillation can be stated as follows:

Equation 2-17:

$u_{AngleMod}(t) = \hat{v}_c \cdot \cos(\omega_c \cdot t + \Delta\varphi_c \cdot \cos\omega_S \cdot t)$ (see also Fig. 2-17d).

Given the generally accepted statement $\omega = \frac{d\varphi_{AngleMod}}{dt}$,

the instantaneous frequency $f_{AngleMod}(t)$ of an angle-modulated oscillation calculates out to:

Equation 2-18:

$$f_{AngleMod}(t) = \frac{1}{2\pi}\frac{d\varphi_{WM}}{dt} = f_c - \Delta\varphi_c \cdot f_s \cdot \sin\omega_s \cdot t = f_c - \Delta f_c \cdot \sin\omega_s \cdot t$$

(see also Fig. 2-17c).

From Equation 2-18 it can be seen that the instantaneous frequency $f_{AngleMod}(t)$ is made up of the constant frequency component f_c and the alternate component $f_{c_a} = -\Delta f_c \cdot \sin\omega_s t$. The frequency deviation is calculated out to:

Equation 2-19:

$$\Delta f_c = \Delta\varphi_c \cdot f_s$$

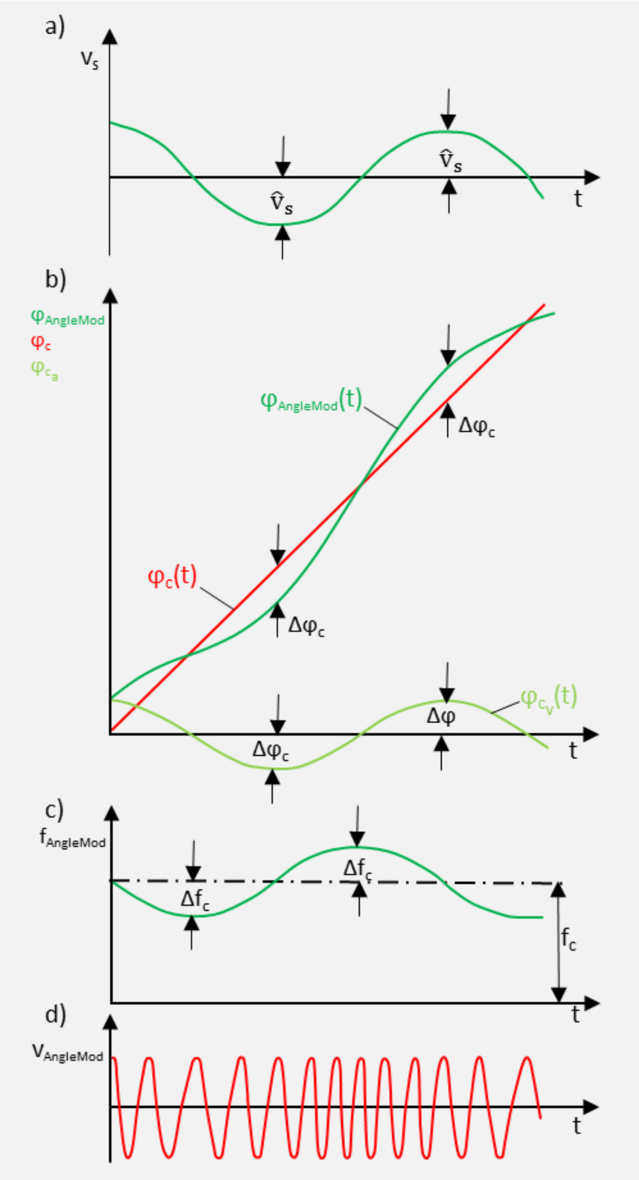


Fig. 2-17: The phase angle and frequency of an angle-modulated signal based on the wanted signal. a) Wanted signal, b) Phase angle, c) Frequency, d) Time function of the angle-modulated oscillation.

Like for amplitude modulation (see Equation 2-11), the equation for angle modulation $v_{AngleMod}(t)$ can be used to derive a representation of the frequency spectrum. Because a simple trigonometric transformation of the expression $\cos(\omega_c \cdot t + \Delta\,\varphi_c \cdot \cos\omega_c \cdot t)$ from Equation 2-17: is not possible, the derivation requires the use of complex notation, followed by an application of the power series expansion of the e-function. The full derivation will not be explained in more detail here. The result of this method is:

Equation 2-20:

$$v_{AngleMod}(t) = \hat{v}_c \cdot [J_0(\Delta\varphi_C) \cdot \cos\omega_c t - J_1(\Delta\varphi_c) \cdot [\sin(\omega_c + \omega_s)\,t +$$
$$+ \sin(\omega_c - \omega_s)\,t] - J_2(\Delta\varphi_c) \cdot [\cos(\omega_c + 2\omega_s)\,t + \cos(\omega_c - 2\omega_s)\,t] +$$
$$+ J_3(\Delta\varphi_c) \cdot [\sin(\omega_c + 3\omega_s)\,t + \sin(\omega_c - 3\omega_s)\,t] + \cdots]$$

From the above equation, it can be seen that the spectrum of an angle-modulated oscillation contains not only the carrier signal with frequency f_c, but theoretically also an infinite number of discrete sideband oscillations at the frequencies $f_c \pm n \cdot f_s$ in symmetry to the carrier frequency. The amplitudes of the individual frequency components are calculated from the Bessel function values of the first kind $J_n(\Delta\varphi_c)$ and the nth kind (Fig. 2-18). Even where the Bessel functions have negative values, the amplitude values of the two side lines are displayed as positive numbers in the spectrum. In the phase diagram, this is indicated by a phase shift of 180°.

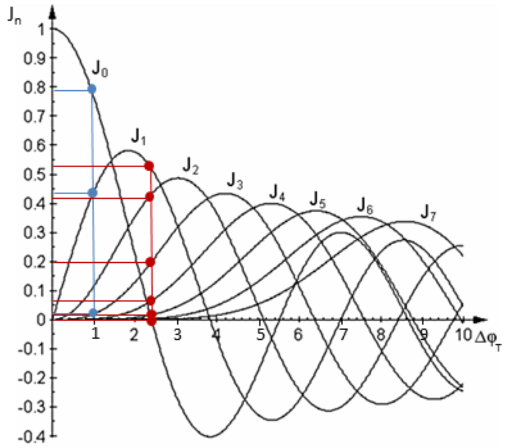


Fig. 2-18: Bessel function $J_n(\Delta\varphi_c)$ for n=0 to 7 over a range from $\Delta\varphi_c = 0$ *to* $\Delta\varphi_c = 10$

The example values in Fig. 2-18 for $\Delta\varphi_c$ ($\Delta\varphi_c = 1$, blue; $\Delta\varphi_c = 2.4$ red) result in the spectra shown in Fig. 2-19 for a sine-wave, angle-modulated oscillation.

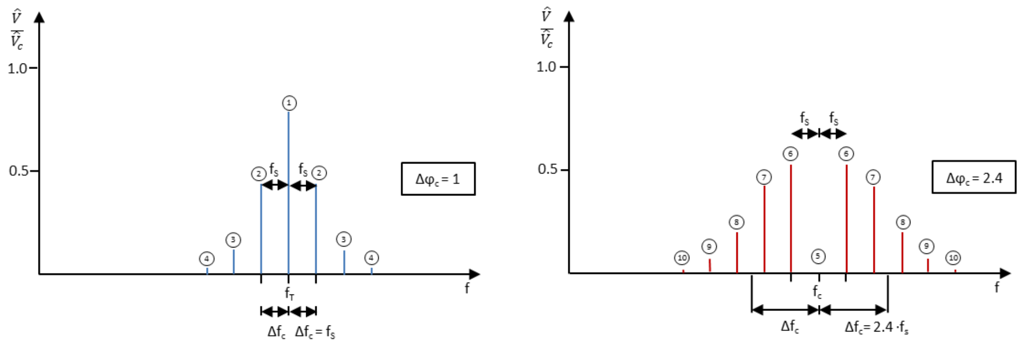


Fig. 2-19: Frequency spectra for an angle-modulated oscillation at various phase deviations. $\Delta\varphi_T$

As seen from the Bessel functions, the carrier oscillation or individual sideband oscillations can range from $\Delta\varphi_c$ to zero at certain values. For example, at $\Delta\varphi_c = 2.4$, the carrier component disappears completely from the spectrum. It can also be seen that the carrier amplitude depends on the phase deviation and therefore takes on the larger value when unmodulated ($\Delta\varphi_c = 0$).

Per Equation 2-20, the bandwidth of an angle-modulated signal can be very large. In particular, an increase in the phase deviation will cause ever more sideband oscillations to appear. In practice, therefore, the bandwidth is considered to be finite. This includes all spectral lines up to around 10 % or 1 % of the carrier amplitude. As a result, 90 % or 99 % of the spectral lines lie within the bandwidths calculated with Equation 2-21 or Equation 2-22.

Equation 2-21:

$B_{AngleMod_{10\%}} = 2 \cdot f_{s_{max}} \cdot (\Delta\varphi_c + 1)$ for average transmission quality

Equation 2-22:

$B_{AngleMod_{1\%}} = 2 \cdot f_{s_{max}} \cdot (\Delta\varphi_{Tc} + 2)$ for high transmission quality

This means that the bandwidth of an angle-modulated signal ($\Delta\varphi_c > 0$) is always greater than the bandwidth of an amplitude-modulated signal.

2.1.6 Differentiation of Frequency versus Phase Modulation in Angle Modulation

In angle modulation, frequency modulation (FM) and phase modulation (PM) occur simultaneously. The instantaneous phase angle $\varphi_{AngleMod}(t)$ is calculated as shown in Equation 2-16 and the instantaneous frequency $f_{AngleMod}(t)$ is calculated as shown in Equation 2-18. There is a fundamental link between the frequency deviation and the phase deviation

as described in Equation 2-19: $\Delta f_c = \Delta\varphi_c \cdot f_s$

In practice, however, a distinction is made between FM and PM during angle modulation. The difference lies in how the modulation product for an angle-modulated oscillation is generated. It is not possible to tell from the modulation product or the representation as a time function $v_{AngleMod}(t)$ whether an FM or a PM is involved. However, it is possible to distinguish between the two by studying the spectrum at different frequencies of a wanted sine-wave signal. As seen in Fig. 2-20, a change in the wanted signal frequency causes a change in the spectral line spacing for both FM and for PM. In phase modulation, the amplitudes of the frequency components are not influenced by the wanted signal frequency. In contrast, the amplitude of the carrier and the sideband oscillations undergo a change in frequency modulation.

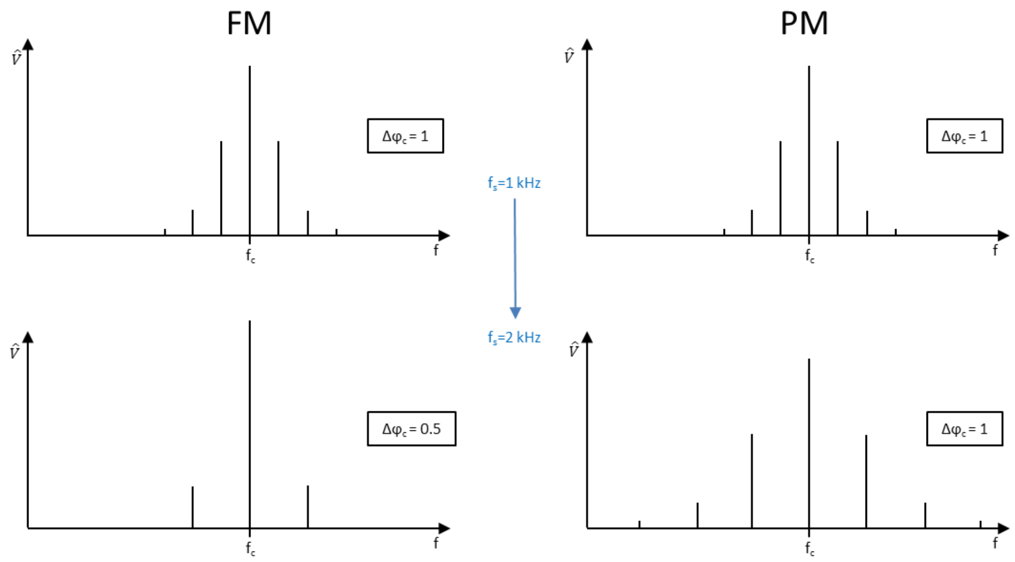


Fig. 2-20: Frequency spectra for FM and PM at different frequencies and constant amplitude $\hat{u}_s$ of the wanted signal.

The following applies for frequency modulation: After a change in the frequency of the wanted signal with constant amplitude, the frequency deviation remains constant and the phase deviation changes proportionately with the wanted signal frequency.

The following applies for phase modulation: After a change in the frequency of the wanted signal with constant amplitude, the phase deviation remains constant and the frequency deviation changes directly proportionately with the wanted signal frequency.

In practice, a true phase modulation makes little sense for analog applications. It is very important however for digital applications and is quite frequently used (2.4.3).

2.1.7 Frequency Modulation with Predistortion

For FM at high frequencies for the wanted signal, the amplitudes of the two sideband oscillations are smaller, which results in a reduction of the signal-to-noise ratio during radio transmission. The reason is that the phase deviation behaves in directly proportionate opposition to the wanted signal frequency ($\Delta\varphi_c \sim \frac{1}{f_s}$).

In FM radio broadcasting, this disadvantage of a decreasing phase deviation at higher frequencies can be compensated by a transmitter-side predistortion of the signal using an RC highpass before the modulator, called preemphasis. This ensures that the amplitudes of high frequencies are boosted nonlinearly. Caused by the predistortion the FM transmitter signal contains both frequency-modulated and phase-modulated characteristics. To restore a linear amplitude frequency response on the receiver end, the boost in amplitude for the high frequency signal components after the FM demodulator must be reversed by means of a counteracting deemphasis using a lowpass filter.

The corresponding standards define the appropriate time constants for the precorrection and the distortion. These can be used to calculate the associated limit frequencies for the highpass and lowpass filters.

2.1.8 Pros and Cons of Frequency Modulation versus Amplitude Modulation

Because almost all spurious signals affect the signal amplitude during radio transmission, but do not change the frequency, a significant advantage of FM is a greater resistance to interference. In FM demodulation, the wanted signal is recovered from the sequence of signal zero crossings, which means that amplitude changes of the FM signal will have no effect as long as the receiver is not overdriven or operated at its sensitivity limit. In addition, the dynamic range for FM is greater than that for AM and is limited only by the distance to an adjacent transmitter. The relatively large spacing in transmitter carrier frequencies for VHF FM permits a good quality low frequency transmission, since almost the entire bandwidth can be used. The bandwidth for FM is 15 kHz while it is 4.5 kHz for AM. The maximum wanted signal frequency for AM is dependent on the offset to the adjacent carrier. At 4.5 kHz offset, the upper limit frequency is therefore similar to that for telephony.

The disadvantage of FM is that the technical effort to implement the required signal predistortion (see section 2.1.7) is greater than for AM.

2.2 Digital Modulation Methods

It is possible to reduce the influence of interference signals on transmitted information signals by expanding the information signal spectrum for the selected modulation method. Examples are the analog frequency modulation and phase modulation described in section 2.1.4.

Another, more efficient option of transmitting a signal as free of interference as possible is to use digital modulation methods. Only digital modulation provides methods such as coding that make it possible to regenerate a faulty information signal by means of an error correction during decoding at the receiver. The interference from some digital modulation methods can be analyzed using an eye diagram or by a constellation diagram, which displays transmit symbols in the complex plane (constellation diagram).

The digital information signal is transmitted either in its original frequency domain via wire (baseband transmission) or via radio link. The radio link transmission is accomplished by means of modulation (keying) of a high-frequency carrier.

Methods for baseband transmission are for example pulse code modulation and delta modulation. These types of modulation are not discussed in this educational note. Radio transmission involving the keying of a high-frequency sine-wave carrier, as used in wireless communications, is described in more detail in section 2.4.

Digital modulation methods transmit symbols that are uniquely defined for the transmitter and receiver. A symbol represents the smallest information element. Symbols combine – depending on modulation mode – n individual bits, where n is taken from the set of natural numbers. These symbols form the set of numbers used by a modulation mode. The throughput of the air interface is constrained by the symbol rate and the number of bits per symbol. The shape of the symbols must be changed through filtering (baseband filtering) so that the spectrum required for transmission remains within the prescribed bandwidth for the transmission channel. Analog signals (voice, music) must be digitized for transmission by means of a digital modulation method. The digital data is then mapped to the symbols to be transmitted. The digital modulation delivers valid values only at defined sample times and is therefore time discrete. The spacing in time between the sampling points is called the symbol rate. Digital modulation transmits only a finite number of different values, and is therefore value discrete. The transmission can hold 2^n numbers with n bits. Referenced to the transmitted information signal, a digital modulation therefore is considered to be a discrete-time and discrete-value modulation method. The time characteristic of the high-frequency modulation signal on the other hand is continuous time and continuous value.

Some of the digital modulation methods are derived directly from the analog modulation methods. By studying a digitally modulated signal in the vector diagram, it becomes clear that digital modulation is nothing more than analog modulation with a finite number of discrete states.

Phase modulation plays an important role here, in particular for wireless communications (see 2.4.3). However, there are also a number of digital modulations that do not originate directly from analog methods; these are not discussed in any detail here. For example, pulse width modulation represents a special digital angle modulation that can also be used for discrete time sampling of an analog signal.

Digital modulation methods can also be used to divide the wanted data stream over multiple carriers (multi-carrier modulation). This leads to an additional option of optimally matching the characteristics of a transmission channel. If narrowband interference occurs within the wanted signal spectrum, this method can be used to exclude the carriers affected by the interference from the data transmission. Although this reduces the overall data throughput somewhat, the data transmission remains possible in spite of the interference. Because destructive interference as a result of multipath reception (fading) affects only individual carriers, multi-carrier transmission offers a clear advantage over a single-carrier method.

A typical multi-carrier method is orthogonal frequency division multiplex (OFDM), which is used for the long-term evolution (LTE) mobile cellular standard. This also includes the coded orthogonal frequency division multiplex (COFDM) method, which is used for terrestrial digital television DVB-T, for example.

2.3 Principles of I/Q Modulation

Conventional analog modulation methods vary an RF carrier signal in only one dimension. Amplitude modulation changes only the amplitude A(t), while frequency and phase modulation affect only the frequency or the phase φ(t) of the carrier, respectively. Amplitude modulation takes place primarily during the RF output stage, and the frequency and phase modulation take place during the oscillator stages. The modulation signal has the same waveform as the wanted signal.

State-of-the-art I-Q modulators vary the RF carrier signal in two dimensions. They synthesize the modulation signal from the sum of two baseband signals, i(t) and q(t), where i(t) represents the inphase component and q(t) the quadrature component. Orthogonally applied in the I/Q plane, both extend over a vector with length A(t) and phase φ(t); see Fig. 2-21. i(t) and q(t) are normalized at a constant ≤ 1.

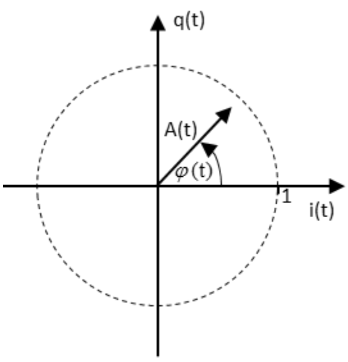


Fig. 2-21: Baseband vector in the I/Q plane.

The modulated RF signal is generated by adding one carrier modulated with i(t) and one modulated with q(t) – but phase-shifted by 90°; see Fig. 2-22.

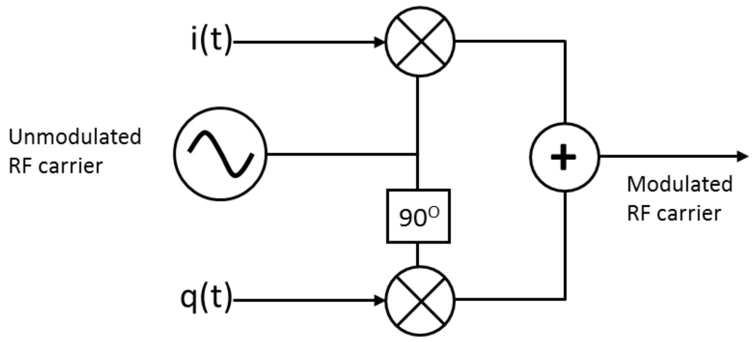


Fig. 2-22: Upconversion of the baseband signals.

By using two baseband signals, it is possible to determine the magnitude as well as the phase of the modulated signal at any given point in time. This also makes it fairly easy to implement analog modulation.

Amplitude modulation is performed by varying the length of the vector φ(t) versus time without changing the phase. Frequency modulation and phase modulation involve varying the phase φ(t) of the sum vector without changing the amplitude (see also I-Q Modulator, section 1.2.2.1).

Constellation diagram

The constellation diagram provides a graphical representation of the in-phase and quadrature-phase components of the digitally modulated signal. This displays a vector in a Cartesian system of coordinates, along with the amplitude and phase of a given carrier at the sampling point in time. Because the symbols to be transmitted represent complex numbers, they can be displayed as points in the constellation diagram, which corresponds to the complex plane. These points in turn correspond to the terminals of the respective I/Q vectors.

Fig. 2-23 shows the constellation diagram for binary phase shift keying (BPSK) and for 4-PSK (see section 2.4.3). As can also be seen here, the included I/Q states are in a steady state only at the symbol points in time. Of course, transitions occur between those points in time.

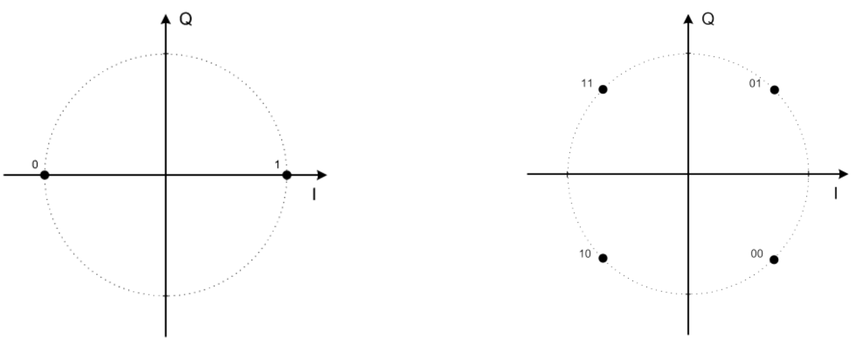


Fig. 2-23: Constellation diagram for BPSK (left) and 4-PSK (right).

2.4 Digital Modulation of a Sine-Wave Carrier

Like for analog modulation, the wireless transmission of a digital baseband signal also requires transition to a higher frequency range by modulating a sine-wave carrier. One or more parameters of the carrier oscillation, i.e. amplitude, frequency or phase, can still be influenced by the digital baseband signal. Because the digital signal oscillates only between two fixed signal values, this is not considered modulation but rather a keying of the carrier oscillation. As such, there are three keying options:

- Amplitude shift keying (ASK)
- Frequency shift keying (FSK)
- Phase shift keying (PSK)

With a carrier that can assume M = 2^n states for amplitude, frequency or phase, a group of n bits can be transmitted simultaneously in one clock step. This is included in the method names, e.g. 2-ASK, 4-PSK or 8-FSK.

2.4.1 Amplitude Shift Keying (ASK)

If a digital bit sequence changes only the carrier amplitude, this is considered to be amplitude shift keying. For a digital high bit "1", the maximum carrier amplitude is transmitted. For a low bit "0", the carrier switches off (Fig. 2-24). ASK can be performed with multiple increments of the carrier amplitude. For example, with four amplitude increments, it is possible to transmit 2 bits (00, 01, 10, 11) simultaneously at double the rate. Because interference in the transmission channel is almost always exhibited as amplitude interference, the multi-value ASK is more prone to interference.

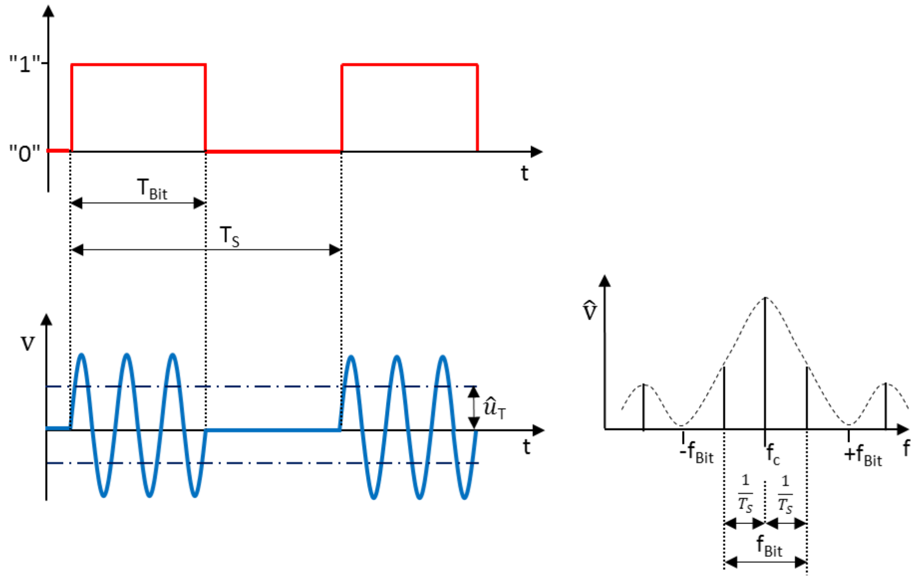


Fig. 2-24: Amplitude shift keying (ASK) in the time and frequency domain for a periodic 1-0 bit sequence.

The amplitude shift keying of the binary code signal in Fig. 2-24 with step duration T_{Bit} and period duration T_S is performed using the values $2 \cdot \hat{v}_c$ and 0 for the carrier amplitudes. As such, a 2-ASK signal can be stated as follows:

Equation 2-23:

$$\mathrm{v}(t) = \begin{cases} 0 & \text{for logical "0"} \\ \hat{\mathrm{v}}_T \cdot cos\omega_c t & \text{for logical "1"} \end{cases}$$

The spectrum of the modulation product is obtained by convolution of the code signal spectrum with carrier frequency f_c or through multiplication of the Fourier series

Equation 2-24:

$$c(t) = \frac{1}{2} + \frac{2}{\pi}\left(cos\ 2\pi\frac{t}{T_S} - \frac{1}{3}cos 2\pi\frac{3t}{T_S} + \cdots\right)$$

with the carrier oscillation

Equation 2-25:

$$\mathrm{v}_c(t) = \hat{\mathrm{v}} \cdot cos\omega_c t$$

The resulting spectrum of the keyed carrier oscillation then contains the information for the code signal in the upper and lower sideband, in symmetry to the carrier (Fig. 2-24, bottom right). The ASK bandwidth is greater than that for analog AM. As a result, fewer transmitters can transmit next to one another within a transmit frequency domain without interference. To detect the signal state, it is sufficient to use appropriate filters to ensure that the transmission includes only the spectrum for the first sideband oscillations.

The minimum required transmission bandwidth is:

Equation 2-26:

$$B_{HF} = 2 \cdot \frac{1}{T_S} = f_{Bit}$$

To ensure that the step frequency f_{bit} is transmitted correctly and the receiver can still unambiguously decode the signal, in practice a bandwidth around a factor of 1.4 wider is selected, i.e.

Equation 2-27:

$$B_{HF,pr} = 1.4 \cdot f_{Bit}$$

2.4.2 Frequency Shift Keying (FSK)

Frequency shift keying takes place when a digital bit sequence changes only the carrier frequency. In this case, the frequency of a carrier oscillation changes erratically during digital data transmission. For example, with a 2-FSK method, the binary states "0" and "1" are transmitted with frequencies f_1 and f_2 (see Fig. 2-26). These two identification frequencies are placed in symmetry to the carrier frequency f_c. In the case of 4-FSK, each symbol frequency transmits two bits simultaneously, whereby the each of the symbols 00, 01, 10 or 11 is assigned one frequency.

The switchover between the individual frequencies can be done in any of several different ways. The simplest method is to switch the digital information directly between two separate oscillators with frequency f_1 and f_2. This method is known as "hard FSK keying". Because the individual oscillators can have a random phase angle to one another, typically a discontinuous transition takes place at the individual switching times, resulting in a discontinuous phase characteristic (Fig. 2-25). The Fourier analysis of this type of FSK signal shows an undesirably high bandwidth requirement. A 2-FSK transmit signal with a square-wave, "hard" keying can be described as follows:

Equation 2-28:

$$\mathrm{v}(t) = \begin{cases} \hat{\mathrm{v}}_c \cdot cos(2\pi f_1 t + \varphi_1) \text{ for logical "0"} \\ \hat{\mathrm{v}}_c \cdot cos(2\pi f_2 t + \varphi_2) \text{ for logical "1"} \end{cases}$$

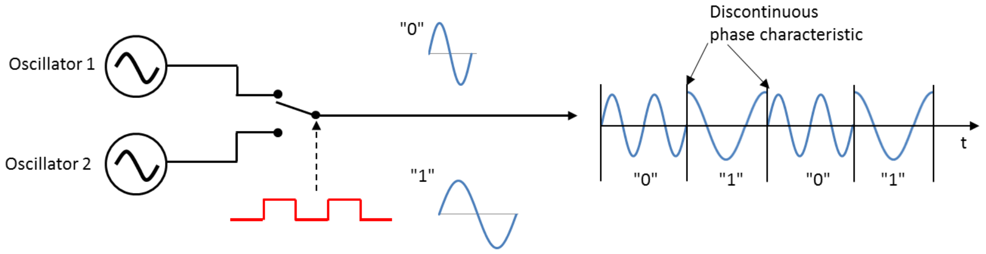


Fig. 2-25: "Hard" FSK keying.

Due to the high bandwidth requirement, an FSK signal with continuous phase characteristic as shown in Fig. 2-26 is typically used. This is also known as continuous phase FSK (CPFSK). The signal can be generated through the use of a VCO or using an I-Q modulator, for example. The bandwidth requirement is less because the individual oscillators are synchronous unto themselves and do not cause any phase discontinuity. The "hard" keying is thus replaced with a continuous characteristic. The result is a smaller bandwidth requirement, and is known as a "soft FSK keying". However, because the switchover between transmit frequencies is not abrupt, intersymbol interference arises, which can make the demodulation more difficult.

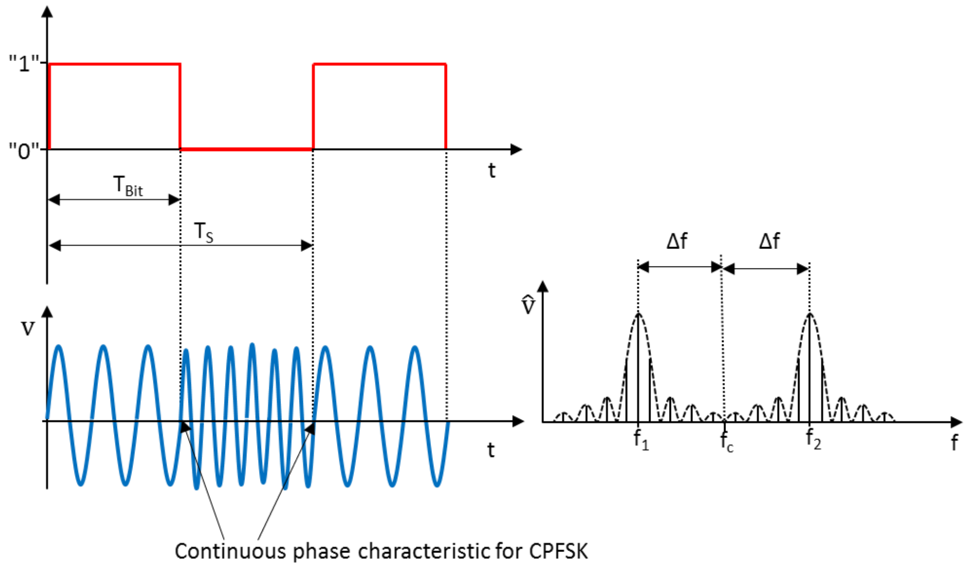


Fig. 2-26: 2-FSK frequency shift keying in the time and frequency domain for a periodic 1-0 bit sequence.

A 2-FSK signal can also be seen as the superposition of two amplitude-keyed oscillations with the frequencies f_1 and f_2. This translates into the spectrum based on a superposition of the spectra of the two ASK signals. The required bandwidth for a 2-FSK signal depends on the frequency deviation Δf or on the frequency difference $|f_1-f_2| = 2\cdot\Delta f$. For $2\cdot\Delta f \gg \frac{1}{T_{Bit}}$, with a periodic 1-0 bit sequence, the result is a symmetrical spectrum as shown in Fig. 2-26, bottom right. If the distribution is unequal, the spectrum will be asymmetrical. For $2\cdot\Delta f > \frac{1}{T_{Bit}}$, the following bandwidth can be approximated:

Equation 2-29:

$$B_{HF} = 2\cdot(\Delta f + B)$$

where: B = Bandwidth of the baseband signal

The modulation index m for FSK is the product of the overall frequency deviation $(2\cdot\Delta f)$ and the duration T_{Bit} of a symbol:

Equation 2-30:

$$\mathrm{m} = 2\cdot\Delta f\cdot T_{Bit}$$

Like for analog FM, the frequency deviation Δf is defined as the maximum deviation of the instantaneous frequency from the carrier frequency f_c. As seen in Equation 2-30, the modulation index decreases as the offset for the two frequencies f_1 and f_2 is reduced at a constant symbol rate. The smallest possible value of m that still permits an orthogonal FSK is 0.5. It is only with an orthogonal FSK and an integer multiplier of m = 0.5 that the intersymbol interference between two individual symbols remains at a minimum, with the result that the mutual influence is also minimal. The method that uses m = 0.5 is known as minimum shift keying (MSK), or fast frequency shift keying (FFSK). For MSK, Δf between binary "1" and "0" is exactly one half the data rate.

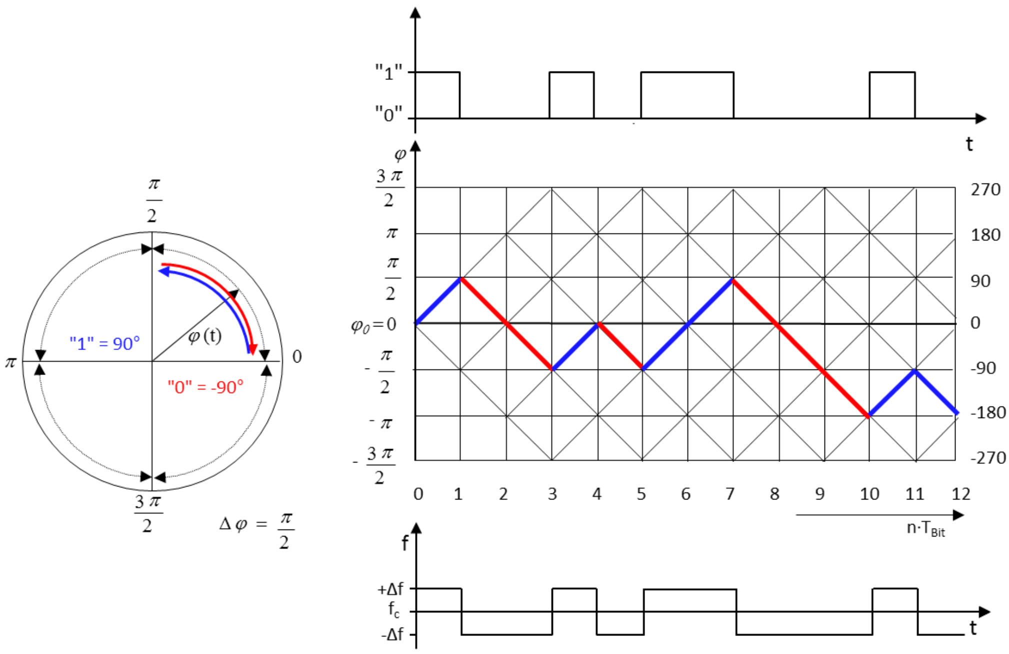


Fig. 2-27: Relationship between data signal, frequency, and phase for MSK.

Fig. 2-27 shows the frequency and phase characteristics of an MSK. Like CPFSK, MSK exhibits a constant phase characteristic. A phase shift of +90 or –90 occurs for logical "1" or "0" at the respective bit boundaries, resulting in breakpoints in the phase at every bit change. Around the carrier frequency f_c, the frequency hops $+\Delta f$ or $-\Delta f$ occur at the breakpoints. To save bandwidth, the digital data stream is filtered using a Gaussian filter before the modulation, and the MSK is transformed into a Gaussian-filtered minimum shift keying (GMSK). This filter flattens the steep edges of the square-wave digital signals and eliminates high frequency signal components. Like 2-FSK, GMSK transmits only one bit per symbol. Because the modulation is performed by means of a phase shift of ± π/2, exactly four possible states result. This causes a frequency shift in the spectrum corresponding to one quarter of the symbol frequency. The level of the modulated signal, i.e. the length of the I/Q vector, remains constant. Dependent on the previous history, the actual phase shift might be somewhat more or less than π/2 as a result of the GMSK filtering. Three possible signal states arise in each quadrant as shown in Fig. 2-28.

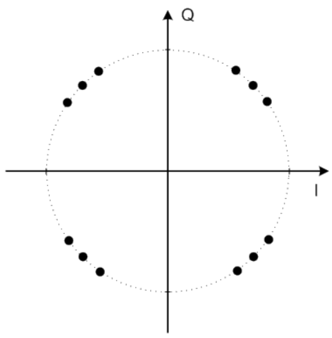


Fig. 2-28: Constellation diagram for GMSK.

GMSK significantly reduces the sidelobes in the spectrum, which can interfere with adjacent channels. An example application for GMSK is the global system for mobile communications (GSM) mobile cellular standard for data transmission. The filtering of the binary signal transforms the 3.7 µs square-wave into 18.5 µs Gaussian impulses. Fig. 2-29 shows a comparison of the MSK and GMSK spectrums, which in addition to GSM are also used for digital enhanced cordless telecommunications (DECT). The term $B \cdot T$ where $T=T_{Bit}$ describes the efficiency of the filtering. This term normalizes the filter bandwidth B to the bit duration T_{Bit}. $B \cdot T_{Bit}=\infty$ refers to MSK, i.e. without filtering.

The sidelobes become less pronounced as the value of the bandwidth time product decreases, and the intersymbol interference increases as the required bandwidth decreases. These superpositions and the resulting misinterpretation of adjacent bits in the receiver can be compensated by means of an error correction after the demodulation.

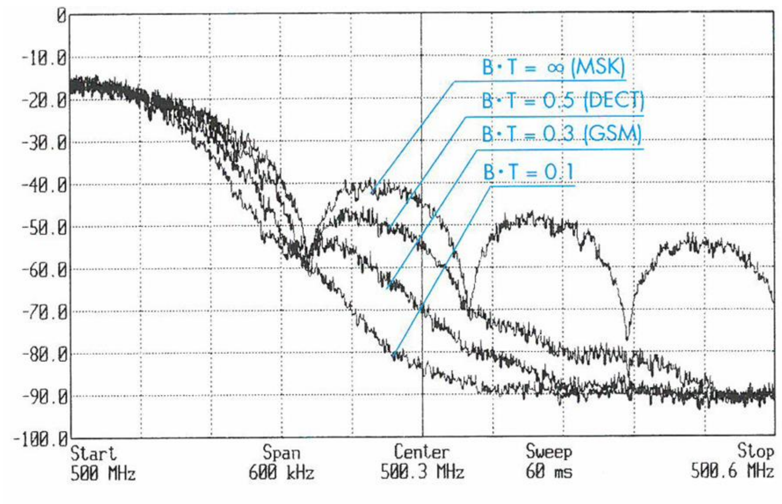


Fig. 2-29: GMSK spectra at different values for B·T.

One reason that GMSK is used for GSM is that the amplitude does not fluctuate. In the vector diagram, the constellation points lie in a circle around zero. The advantage of this is that a GMSK signal without distortion can be processed with a simple, class C amplifier in switched mode. Switching amplifiers have a very high efficiency, typically over 80 %, and thus permit a long battery runtime for cell phones. However, the net data rate for GSM in a timeslot is only 33.85 kbit/s, which is adequate for voice transmission but not for the demands placed on mobile data transmission today. High data rates require higher-order modulations (that also modulate the amplitude). Signals with a higher crest factor place greater demands on the amplifier, which then no longer functions in switched operation and the power consumption is therefore greater.

2.4.3 Phase Shift Keying (PSK)

In phase shift keying, the phase of a carrier oscillation is shifted in accordance with the digital bit sequence. Each symbol to be transmitted is assigned an absolute phase angle. For better differentiation, the phase states are typically equally divided over 360^0. All phase shift keying is based on binary phase shift keying (2-PSK), which is also known as binary shift keying (BSK). Two phase states are available in 2-PSK for displaying the binary "0" or "1". BPSK thus transmits only one bit per symbol. There are therefore $2^1 = 2$ different symbols or states (see Fig. 2-30). The symbol rate is equal to the bit rate.

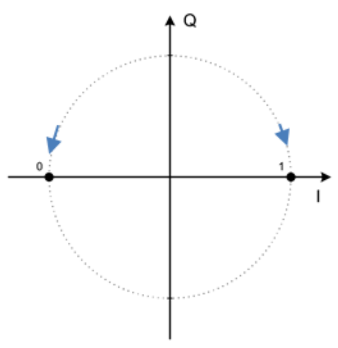


Fig. 2-30: Constellation diagram for BPSK.

As a result, 2-PSK is represented as follows for square-wave keying:

Equation 2-31:

$$\mathrm{v}(t) = \begin{cases} \hat{\mathrm{v}}_c \cdot cos\omega_c t \text{ for logical "1"} \\ -\hat{\mathrm{v}}_c \cdot cos\omega_c t \text{ for logical "0"} \end{cases}$$

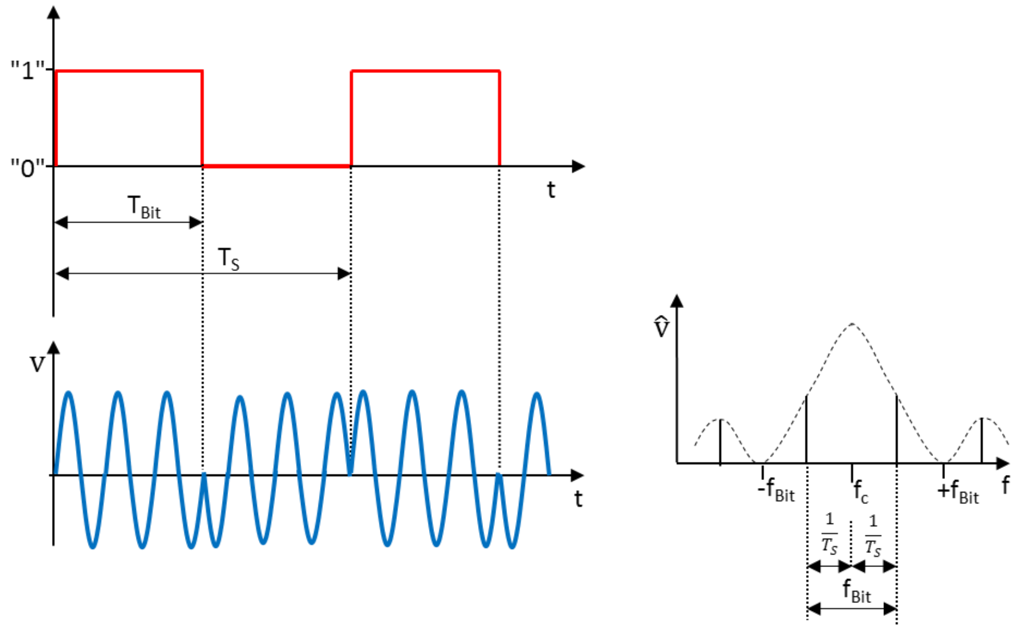


Fig. 2-31: 2-PSK phase shift keying in the time and frequency domain for a periodic 1-0 bit sequence.

Fig. 2-31 shows the data signal and the modulation product in the time domain along with its spectrum. If the signal state of the digital signal changes, the carrier oscillation changes it's phase by 180⁰. The spectrum of the digital modulation signal appears at both sides of the suppressed carrier. The spacing for the first two sideband oscillations is f_{Bit}. The phase shifts occur at the envelope zero crossings. A bandwidth of $B_{HF} = f_{Bit}$ is theoretically adequate for the signal transmission or for the phase selection in the receiver. Like for 2-ASK, in practice 2-PSK is calculated with a value from Equation 2-32.

Equation 2-32:

$$B_{HF,pr} = 1.4 \cdot f_{Bit}$$

Binary keying methods such as 2-FSK have a high immunity to interference (see also the section "Bit error probability" below); however, because only one bit per symbol can be transmitted, the bandwidth utilization is low. Higher transmission rates can be achieved with 4-PSK, for example. 4-PSK is more commonly called quadrature phase shift keying (QPSK). QPSK transmits two bits per symbol. This translates into $2^2 = 4$ different symbol states (Fig. 2-32).

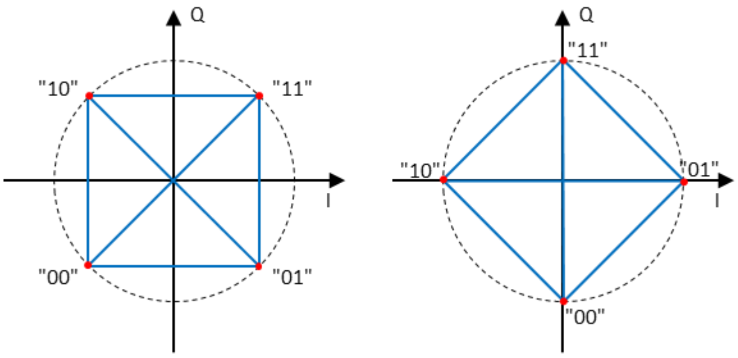


Fig. 2-32: Constellation diagram of possible phase states and shift paths (blue) for QPSK.

As seen here, it is possible to "hop" to any state from any other state. In contrast to GMSK, the entered positions correspond to defined symbols. The symbols used in the example to the left of Fig. 2-32 can be defined in the following mapping table.

Symbol (Dibit)	Δφ
1 1	+45°
0 1	+135°
0 0	–135°
1 0	–45°

Table 2-2: OPSK mapping table

Fig. 2-33 shows the QPSK signal for a symbol sequence 10 11 01 10 00. The following is assumed for the carrier: $v_c(t) = \hat{v}_c \cdot \cos(\omega_c t + \frac{3\pi}{4} + \Delta\varphi)$. The keying is performed in accordance with the mapping table from Table 2-2.

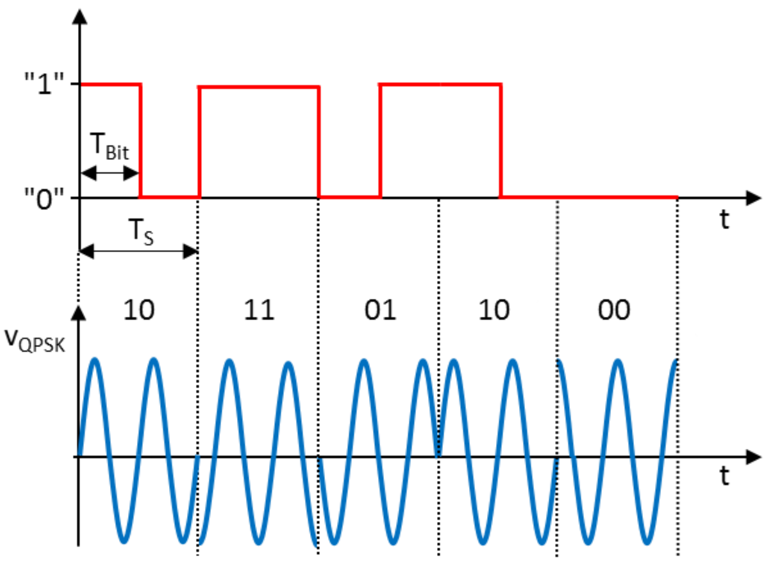


Fig. 2-33: Time characteristic of a binary signal and the QPSK carrier oscillation.

For QPSK, the duration of a signal step T_s is: $T_s = 2T_{bit}$. The width of the QPSK spectrum is reduced by one-half compared to 2-PSK (see Fig. 2-31). In other words, in the 2-PSK spectrum from Fig. 2-31 ,f_{bit} is replaced by $f_{bit}/2$, resulting in the following spectrum for QPSK:

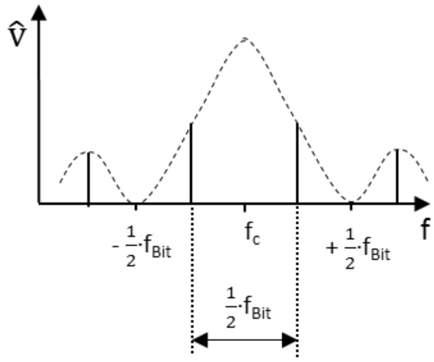


Fig. 2-34: QPSK spectrum.

With the restriction to the first two sideband oscillations, the minimum required transmission bandwidth for QPSK is:

Equation 2-33:

$$B_{HF} = \frac{1}{2} f_{Bit}$$

In practice, the bandwidth for QPSK is also selected at a factor of 1.4 higher to ensure that the step frequency f_{bit} is transmitted correctly and the receiver can correctly decode the signal:

Equation 2-34:

$$B_{HF} = 0.7 \cdot f_{Bit}$$

As a result, QPSK theoretically has twice the bandwidth efficiency (bitrate per Hz of bandwidth) than 2-PSK. The explanation behind this statement is that QPSK can be understood as a quadrature method made up of two 2-PSK methods with carriers shifted by 90º. This orthogonality means that there is no mutual influence or interference of the two 2-PSK signals. As a result, the data rate can be doubled without increasing the transmission bandwidth. However, to achieve the same bit error rates, the required signal-to-noise ratio for QPSK is 3 dB higher than for 2-PSK. More on this topic in the section "Bit error probability".

Important QPSK variants:

As seen in Fig. 2-32, the transition between two diagonally placed transmit symbol points (10 to 01 and 11 to 00) in the complex plane passes through the zero crossing. This means that the carrier power fluctuates strongly and returns to zero at the zero crossing. The result is an undesirably strong AM modulation of the carrier. It is undesirable because modulation methods without a constant envelope, such as GMSK, cannot use non-linear class C amplifiers with a high degree of efficiency and low energy consumption. One of the following would have to be used instead in order to prevent strong signal distortion:

- Linear A amplifier with poor efficiency, or

- CPU-intensive signal predistortion of the amplifier characteristics

As seen in Fig. 2-34, the spectral line for carrier frequency f_c is missing from the spectrum in QPSK modulation, or in any phase-modulated RF carrier for that matter. As a result, the current phase angle is not referenced to the phase of the unmodulated carrier. The zero crossings in QPSK make it more difficult to restore the carrier for a coherent demodulation, by which is meant the exact reconstruction of the carrier frequency with the associated phase angle.

This problem can be alleviated with π/4-DQPSK modulation. Independent of the user data, an additional phase shift of π/4 (45°) is performed after each transmit symbol. This ensures that the transition between two symbols never transverses the origin and always only one carrier amplitude is transmitted (see Fig. 2-35).

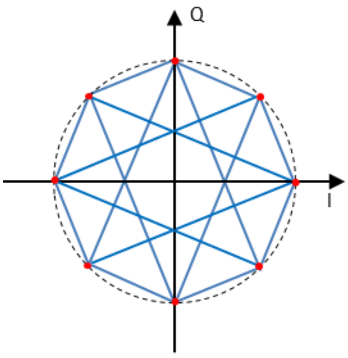


Fig. 2-35: $\frac{\pi}{4}$ DQPSK for preventing zero crossings.

In addition, information is not transmitted regarding the current phase angle, but rather regarding the difference to the previous phase angle. Hence the name **d**ifferential quadrature phase shift keying (π/4-**D**QPSK). The biggest advantage of this method is that no coherent demodulation is needed, eliminating the complex reconstruction of the carrier. In addition, the clock synchronization on the receiver end is made easier because regular phase shifts are present in the receive signal, independent of the user data and its coding. Table 2-3 lists the permissible phase states for π/4-DQPSK. This shows that π/4-DQPSK allows only 45° and 135° phase rotations.

Symbol (Dibit)	Phase transition Δφ
0 0	$0 \cdot \pi/2 + \pi/4 = 45^\circ$
0 1	$1 \cdot \pi/2 + \pi/4 = 3\pi/4 = 135^\circ$
1 1	$2 \cdot \pi/2 + \pi/4 = 5\pi/4 = 225^\circ = -135^\circ$
1 0	$3 \cdot \pi/2 + \pi/4 = 7\pi/4 = 315^\circ = -45^\circ$

Table 2-3: Allowed phase states for π/4-DQPSK

Another option for preventing zero crossings and the associated reduction in amplitude is offered by **o**ffset QPSK (**O**QPSK). To achieve this, the maximum phase change for OQPSK is only 90° instead of the 180° for conventional QPSK. With QPSK, the two bit streams used to generate the I and Q signals are synchronous, i.e. their sign changes occur simultaneously. If both binary signals change state at the same time in QPSK, a phase shift of 180º occurs. In the case of OQPSK, the two data streams are shifted toward one another at half a step length of T_S or at a bit time of T_{Bit} and then applied to the modulators for I and Q (Fig. 2-36), ensuring that they no longer change simultaneously.

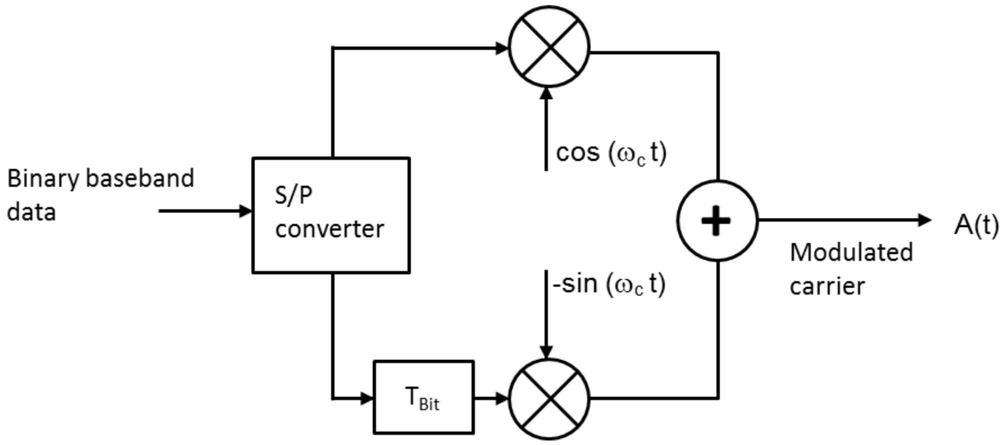


Fig. 2-36: Generating an offset QPSK.

As a result, only one of the two modulators will undergo a phase change at the same time as a bit change. 180º phase shifts are therefore eliminated, and instead are split into two shifts of 90º each. An example is provided in Fig. 2-37.

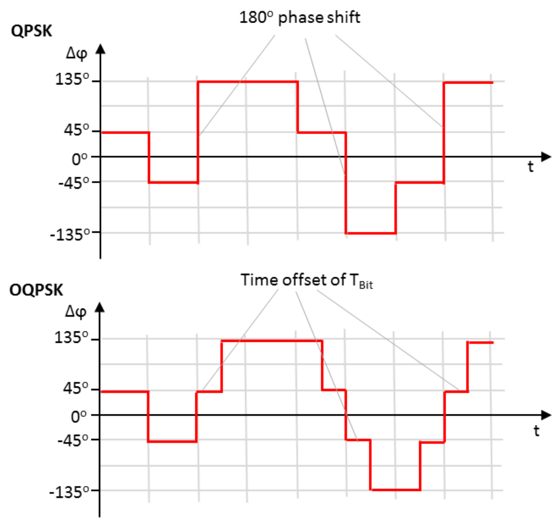


Fig. 2-37: Phase characteristic of QPSK and OQPSK over time.

In contrast to true QPSK, state transitions for OQPSK form a square defined by the four states. This prevents diagonal paths with a zero crossing (Fig. 2-38).

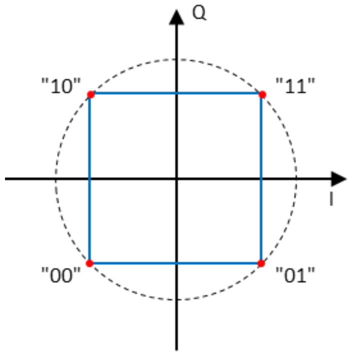


Fig. 2-38: Possible phase states for an OQPSK without zero crossing.

Bit error probability:

At a constant signal-to-noise ratio, the 2-PSK method provides the lowest bit error probability. The susceptibility to noise increases steadily as the PSK order increases. However, because more bits are transmitted per symbol, the data rate also increases. For example, with an 8-PSK with eight possible carrier states, 3 bits are transmitted per step or symbol. Fig. 2-39 shows the phase states of a 2-PSK and 8-PSK in a vector diagram. The decision limit and/or the decision space required for unambiguous detection of the signal state are shown. It is clear that for an error-free demodulation of signal state "1", the amplitude of a spurious signal can come close to the value of the wanted carrier. In the case of 8-PSK, the decision space is significantly smaller, which means that a considerably smaller amplitude of the spurious signal can lead to a transmission error.

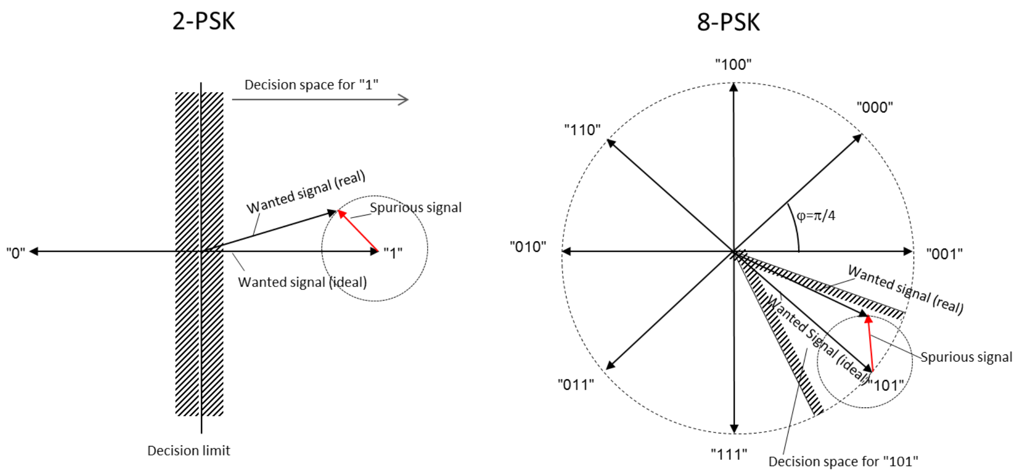


Fig. 2-39: Signal states and decision space for a 2-PSK and a 4-PSK signal in a vector diagram.

2.4.4 Quadrature Amplitude Modulation (QAM)

The preceding sections introduced digital modulation methods for data transmission in which only one parameter (amplitude, frequency or phase) of a carrier oscillation was changed by a modulating binary signal. An increase in the transmission rate is achieved by modulating the amplitude and phase of a high-frequency carrier. The method that combines amplitude and phase modulation is known as quadrature amplitude modulation (QAM). While the signal states lie in a circle for PSK, this is no longer the case when the amplitude is changed. The only exception is 4-QAM, which is identical to 4-PSK/QPSK (see Fig. 2-40, left). Fig. 2-40 (right) shows the signal states for 16-QAM as an example. Four bits are transmitted per symbol. The result is $2^4 = 16$ symbols. The symbol rate is ¼ the bit rate. This symbol placement in the constellation diagram includes three different amplitudes, one with eight different phase values and two with four phase values.

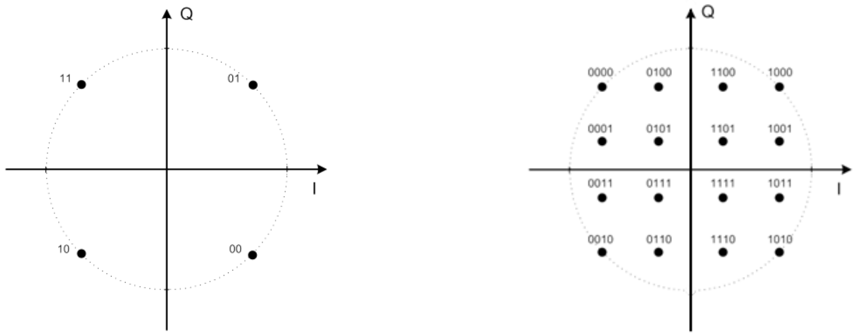


Fig. 2-40: Constellation diagram for 4-QAM or 4-PSK (QPSK) on the left and for 16-QAM on the right.

Fig. 2-41 shows another example of the symbol placement for 16-QAM. This diagram includes four different values for amplitude, each with four values for phase.

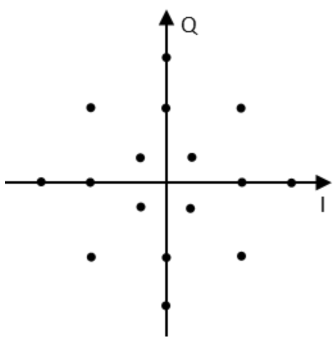


Fig. 2-41: 16-QAM with four amplitudes, each with four different phases.

A QAM requires less power than PSK at the same weighting M and the same bit error probability. The following general statement applies for a QAM signal:

Equation 2-35:

$$\mathrm{v}(t) = I(t) \cdot cos\omega_c t - Q(t) \cdot \sin \omega_c t$$

where $\omega_c = 2\pi f_c$

The following applies for the complex envelope of the QAM signal:

Equation 2-36:

$$\underline{\mathrm{v}}(t) = I(t) + jQ(t) = \hat{\mathrm{v}}_c(t) \cdot e^{j\varphi(t)}$$

Equation 2-36 shows that both the amplitude $\hat{u}_T$ and the phase φ(t) of the modulated carrier are dependent on time.

Fig. 2-40 to the right illustrates in the I/Q plane how the unit circle and enclosed area for a high-order QAM, e.g. 16-QAM, can be used for efficient distribution of the symbols. However, as the number of symbols increases, the spacing of adjacent states is of necessity reduced (Fig. 2-42). As a result, the symbols are more difficult to isolate on the receiver end, especially when they are accompanied by noise from a poor radio link.

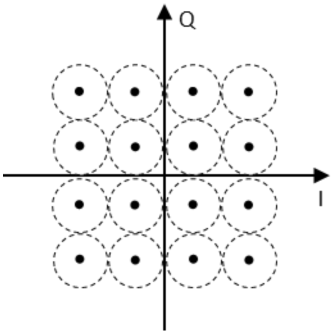


Fig. 2-42: 16-QAM with areas in which an unambiguous symbol detection is possible.

This is why the 256-quadrature amplitude modulation (256-QAM) as shown in Fig. 1-43 is not used for wireless communications standards. However, 256-QAM and other high-order modulations are used in low-noise broadband cable networks with reliable levels, such as DVB-C. The symbol rate is equal to the bit rate / 8; hence 256-QAM transmits eight bits per symbol, or 2^8 = 256 symbols.

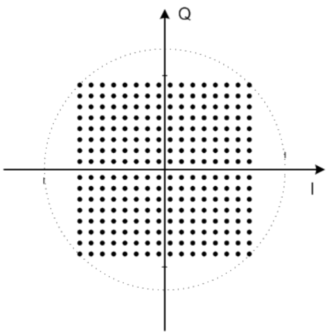


Fig. 2-43: Constellation diagram for 256-QAM.

The higher the modulation order – i.e. the narrower the symbol spacing – the closer the discrete modulation comes to a continuous modulation with respect to the use of amplitude and phase. However, it must be noted that the number of I/Q vector changes per unit of time is relatively low in analog modulation, while it is typically fairly high for digital modulation. The I/Q vector must "hop" back and forth between random symbols.

The following table provides an overview of the types of modulation for important standards from wireless communications as well as from audio and video broadcasting:

Standard	Transmission Method	Modulation
GSM	Single carrier	GMSK
EGPRS	Single carrier	8-PSK (3Pi/8 rotation)
TETRA	Single carrier	Pi/4-DQPSK, 16-QAM, 64-QAM
UMTS, HSPA, HSPA+	CDMA	QPSK, 16-QAM, 32-QAM
LTE	COFDM	QPSK – 64-QAM
DVB-S	Single carrier	QPSK
DVB-S2	Single carrier	QPSK, 8-PSK, 16-APSK, 32-APSK
DVB-T	COFDM	QPSK, 16-QAM, 64-QAM
DVB-C	Single carrier	64-QAM, 256-QAM
WLAN	COFDM	DBPSK – 64-QAM
DAB	COFDM	DQPSK

Table 2-4: Applications for digital modulation methods

2.4.5 Comparing Different Phase Shift Keying Methods

The selection of the correct digital modulation method depends on the requirements for bandwidth utilization and on the required immunity to noise.

The RF bandwidth required for all phase shift keying methods from the 1st Nyquist criterion up to the first harmonic of the fastest possible switch is:

Equation 2-37:

$$B_{HF} = 2 \cdot \frac{1}{2 \cdot T_S} = \frac{1}{T_S}$$

where $T_S = n \cdot T_{Bit}$, it follows:

Equation 2-38:

$$B_{HF} = 2 \cdot \frac{1}{2 \cdot n \cdot T_{Bit}} = \frac{1}{n} \cdot f_{Bit} = \frac{1}{n} \cdot \frac{r_{Bit}}{Bit}$$

As seen in Equation 2-37 and Equation 2-38, the bandwidth depends on the step size T_S or the symbol rate $1/T_S$. The symbol rate in turn can be referenced to the weighting 2^n, i.e. to the number of symbols, the bit rate r_{Bit} or bit sequence frequency f_{Bit} of the binary signal. From 2^n weighted signal elements can be derived a theoretical maximum bandwidth utilization (Equation2-39:).

Equation2-39:

$$\frac{r_{Bit}}{Bit} = n \frac{bit/s}{Hz}$$

The bandwidth utilization that can be achieved in practice lies under that number, as a larger bandwidth must be assumed per Equation 2-32. Practically speaking, the bandwidth utilization is reduced by a factor of 0.7 (Equation 2-40).

Equation 2-40:

$$\frac{r_{Bit}}{Bit} = 0.7 \cdot n \frac{bit/s}{Hz}$$

The comparison of the variously weighted PSK and QAM methods in Table 2-5 shows that the bandwidth utilization increases with the weighting of the method, and the bandwidth can be reduced at the specified bit rate. The table also shows that with an increasing bandwidth utilization, the signal-to-noise ratio (C/N) must steadily increase in order not to stay below a defined bit error rate (BER).

Modulation	2-PSK	4-PSK	8-PSK	16-PSK	16-QAM	32-QAM	64-QAM
Amplitude states	1	1	1	1	3	5	9
Phase states	2	4	8	16	12	28	52
Carrier states	2	4	8	16	16	32	64
Theoretical maximum bandwidth utilization in (bit/s)/Hz	1	2	3	4	4	5	6
Practical bandwidth utilization in (bit/s)/Hz	0.7	1.4	2.1	2.8	2.8	3.6	4.2
C/N in dB for BER=10^{-6}, in practice	10.7	13.7	18.8	24	20.5	24	27

Table 2-5: Comparison of different phase shift keying methods

2.4.6 Crest Factor / Linearity Requirements

The amplitude of the modulated RF transmit signal is always proportionate to the corresponding vector length in the baseband – including between the symbol points in time. When considered in this light, it is clear that the unit circle is frequently violated by the transitions between the symbols. The ratio of peak value to average value is called the crest factor. It determines the power margin that must be reserved at the TX output stage.

Similarly, Fig. 2-44 shows signal breaks that pass through the origin and then return to zero. For an error-free demodulation on the receiver end, it is important that the signal characteristic be transmitted totally without distortion. The transmit output stage must therefore not only maintain some headroom for high power levels, but also must also exhibit very high linearity over the entire power range from 0 watts to the maximum power.

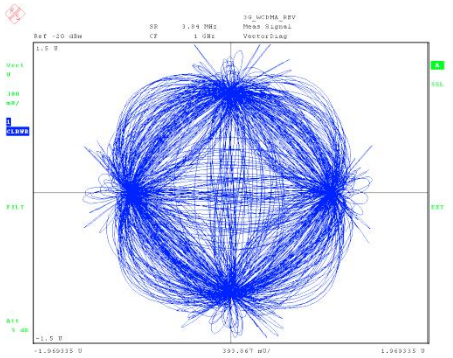


Fig. 2-44: Constellation and continuous vector diagram.

GSM primarily uses the two modulation methods GMSK and 8-PSK. In GMSK, the ideal I/Q vector remains exclusively on the unit circle. Theoretically there are no power peaks or amplitude notches, and no crest factors of 1.

With 8-PSK, the zero crossings required during hops between opposing symbols in the I/Q diagram are avoided by rotating the constellation diagram by 3π/8 between the symbol clocks. The vector diagram in Fig. 2-45 shows the desired effect, which has no zero crossings. The transmit power in Fig. 2-46 shows level breaks of about 12 dB during data transmission, but not down to zero. This somewhat moderates the linearity requirements for the TX output stage.

As a result of the rotation, about 16 symbol states will be recalculated at the receiver end during demodulation.

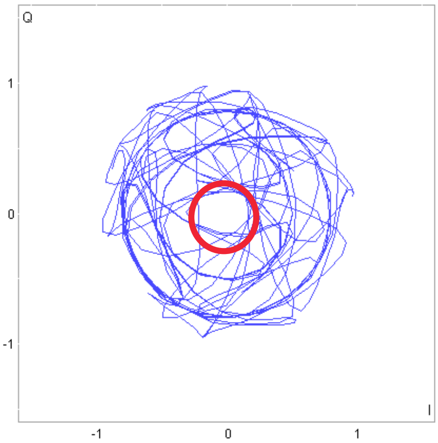


Fig. 2-45: Preventing zero crossings by rotating the constellation.

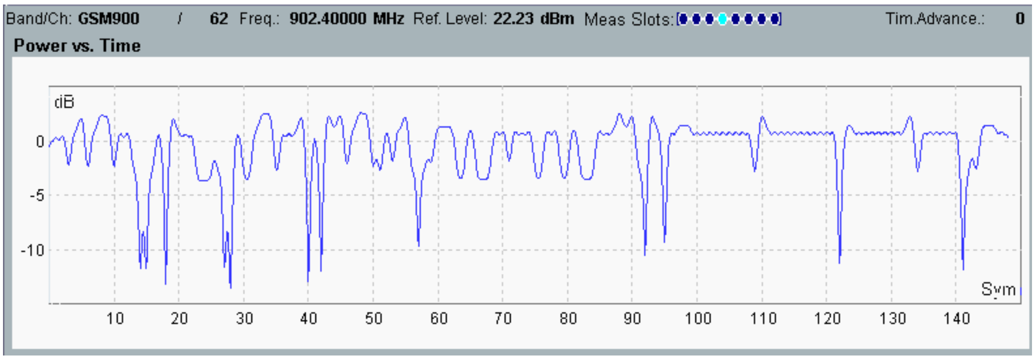


Fig. 2-46: Progression of the envelope power for a 3π/8 rotated, 8-PSK-modulated signal.

2.5 Access Methods

Digital wireless communications networks are organized into cells. Because radio waves have a limited range, it is possible to define an area – called a cell – inside of which service to wireless subscribers is provided by a single base station (BS). At the reception boundaries, a mobile phone (also called a mobile station, MS or user equipment, UE) is automatically switched to the adjacent cell that is most suitable (handover). This typically also involves a change in frequency. The original frequency might then be available again in the next cell over (frequency reuse).

Because every cell is limited to a specific area, it can handle a finite number of wireless subscribers. Or, stated conversely: The size of a wireless communications cell is determined primarily by its required capacity. Because the available resources of frequency, time and – as we will see later – code are limited, locations with a large potential number of subscribers must have many base stations with correspondingly low power. Regions with fewer subscribers can have large cells with powerful base stations. The point is to ensure that multiple subscribers have access to a fixed percentage of the resources frequency, time and code – independent of the uplink / downlink duplex.

Three different methods are used most frequently:

- Frequency-division multiple access (FDMA)
- Time-division multiple access (TDMA)
- Code-division multiple access (CDMA)

Modern vector signal generators are able to offer these methods for standard-compliant generation of signals.

Fig. 2-47 illustrates the primary differences. This simplification assumes only five subscribers (each represented by a different color), all with equal power. This is seldom the case in practice, but it is sufficient for understanding the differences.

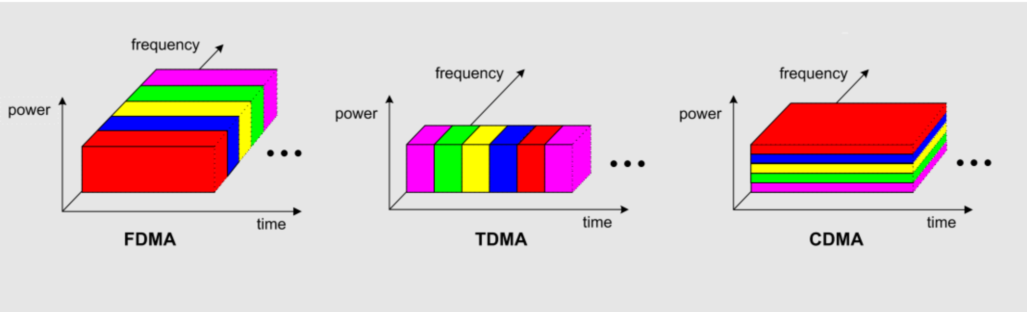


Fig. 2-47: Three important access methods.

FDMA divides the wireless subscribers in a cell among different frequencies. Fig. 2-47 shows five subscribers allocated to five frequencies. For FDD, this means five frequency pairs. FDMA is very easy to implement. This method is particularly useful in small cells where the allocated frequencies can be reused a short distance away. FDMA is frequently used in combination with TDD, e.g. for DECT[1] and WLAN[2] (both standards specialize in connections over short distances).

[1] Digital enhanced cordless telecommunications

[2] Wireless local area network

3 Exercises

3.1 Amplitude Modulation

Amplitude modulation is one of the oldest modulation methods. Even if this type of modulation plays a lesser role in today's world, amplitude-modulated signals represent a very good way of understanding modulation fundamentals.

The objective of this lab exercise is to generate amplitude-modulated signals and to observe and measure variable modulation parameters. It will also provide a feel for the various signal waveforms in the time and spectral domain.

3.1.1 Test Setup

In order to generate and measure modulation signals, you will need a generator, a signal analyzer and an oscilloscope as shown in Fig. 3-1. The oscilloscope is used to measure the modulation signal at the same times as it measures the unmodulated wanted signal.

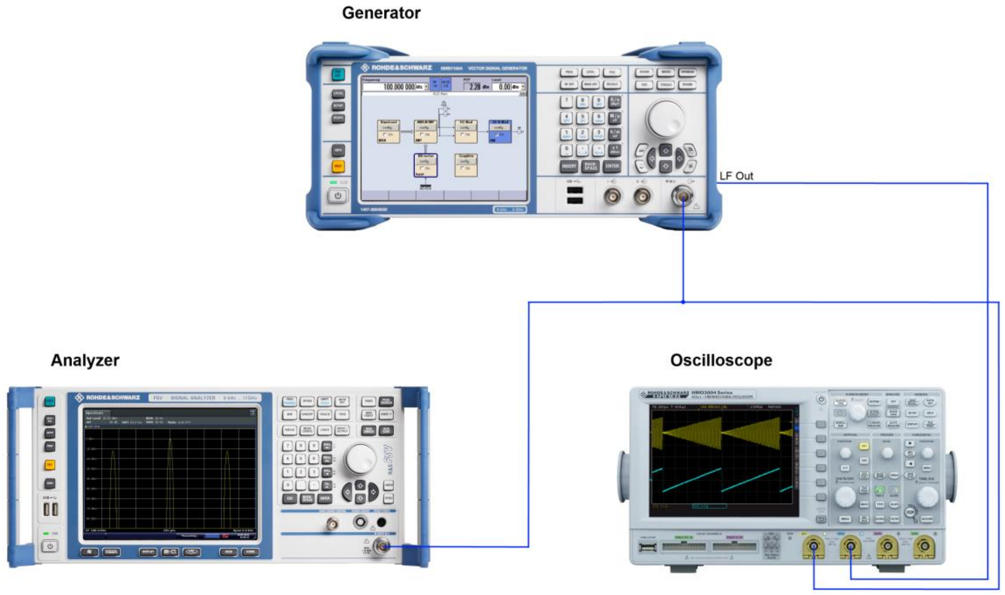


Fig. 3-1: Hardware setup for measuring AM signals.

3.1.2 Measurement

► Set up the signal generator as follows:

Carrier frequency	*100 kHz*
Signal frequency	*1 kHz*
Amplitude	*0 dBm*
Modulation depth	*any*
LF Gen Shape	*Sine*

These settings are made in "RF/A Mod" > "Amplitude Modulation"; see Fig. 3-2.

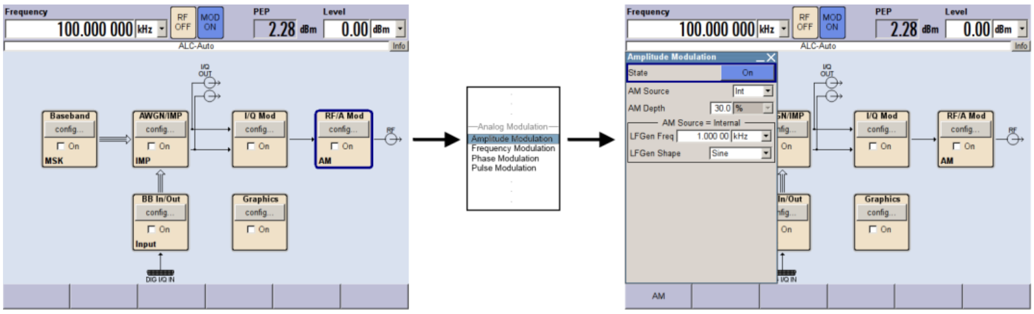


Fig. 3-2: Configuring the generator.

To display the wanted signal on the oscilloscope, switch on the LF generator output on the SMBV.

► Under "RF/A Mod" > "LF Generator / Output", set the "LF Output State" option to ON and set the level to 500 mV; see Fig. 3-3.

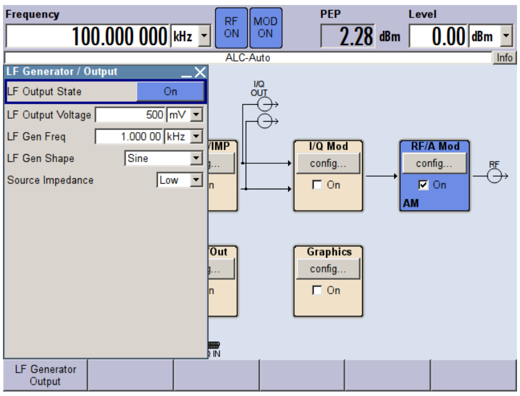


Fig. 3-3: Configuring the LF generator.

► Make the appropriate settings on the oscilloscope.

On the screen, at least two periods of the wanted signal should be clearly visible along with the modulated and the original signal. Verify both signals, taking special note of the phase angle.

Change the modulation depth and other parameters and observe the effects on the time domain signal.

Measurements in the time domain

In the time domain, the modulation depth can be calculated using the formula $m = \frac{\hat{v}_{c,max} - \hat{v}_{c,min}}{\hat{v}_{c,max} + \hat{v}_{c,min}} = \frac{\hat{u}_s}{\hat{u}_c}$; see Fig. 3-4.

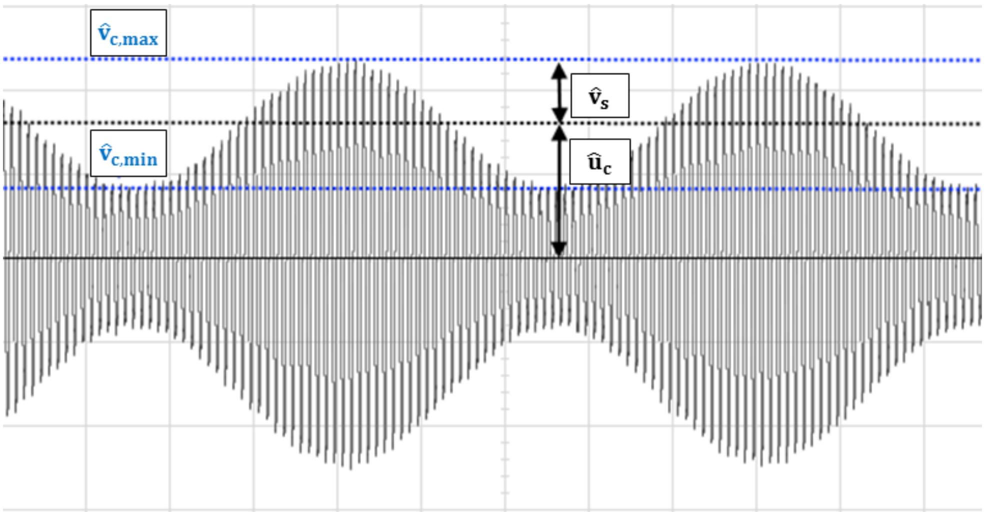


Fig. 3-4: Determining the modulation index in the time domain display.

► Use the oscilloscope to calculate the modulation depth for your setup.

You can also use the modulation trapezoid to quickly determine the modulation depth or to detect non-linear distortions.

► Switch the oscilloscope into XY mode.

► Vary the modulation depth and observe the effects.

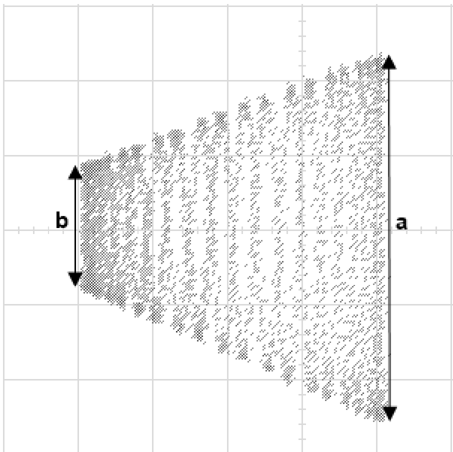


Fig. 3-5: Modulation trapezoid.

The modulation depth can be determined using the following formula for the modulation trapezoid (see Fig. 3-5)

$$m = \frac{a-b}{a+b}.$$

► Use the modulation trapezoid to calculate the modulation depth for your setup.

► What is the shape of the modulation trapezoid for m < 1, m = 1 and m→∞? What special amplitude modulation situation corresponds to m→∞?

Measurements in the frequency domain

► Set the following values on the signal analyzer:

Frequency	*100 kHz*
Range	*10 kHz*

► Set a variety of signal waveforms on the generator and observe the spectrum.

LF Gen Shape	*Square, Triangle, …*

► What bandwidth requirements are placed on the channel in order to make distortion-free transmission possible?

The formula $m = \frac{\hat{u}_{sb}}{\hat{u}_c}$ can likewise be used in the spectrum to determine the modulation depth. Remember that the power of the wanted signal is split into the two sidebands and is displayed in dBm as shown in Fig. 3-6.

► How can the power balance of the transmission be improved? What considerations must be taken into account?

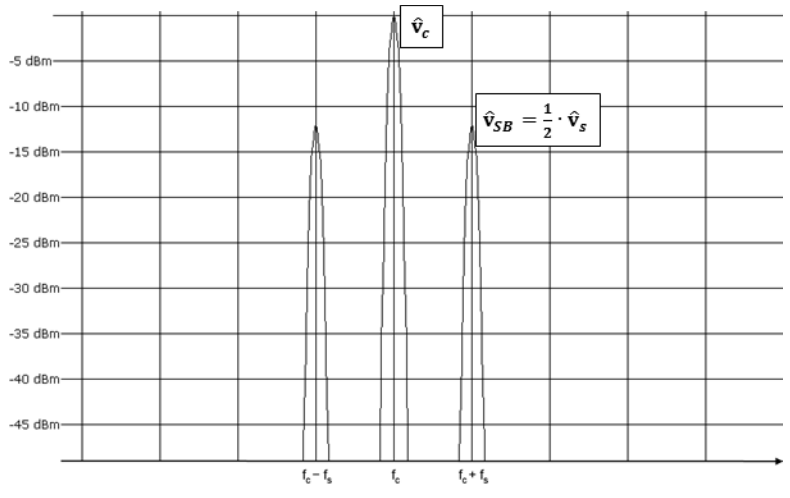


Fig. 3-6: Determining the modulation index from the spectrum.

► Determine the modulation depth from the spectrum for your setup.

► The carrier power of an AM signal is set to 7 dBm and the modulation depth is 50 %. What is the power of one sideband?

3.2 Frequency Modulation

The objective of this lab exercise is to generate and observe frequency-modulated signals in the time and spectral domain as well as to compare and contrast frequency modulation and phase modulation. An FM stereo signal is also generated, analyzed and noise applied.

3.2.1 Test Setup

In order to generate and measure the modulation signals, you will need a generator, a signal analyzer and an oscilloscope as shown in Fig. 3-7. The oscilloscope is used to measure the modulation signal at the same time as it measures the unmodulated wanted signal (output at *LF Out*).

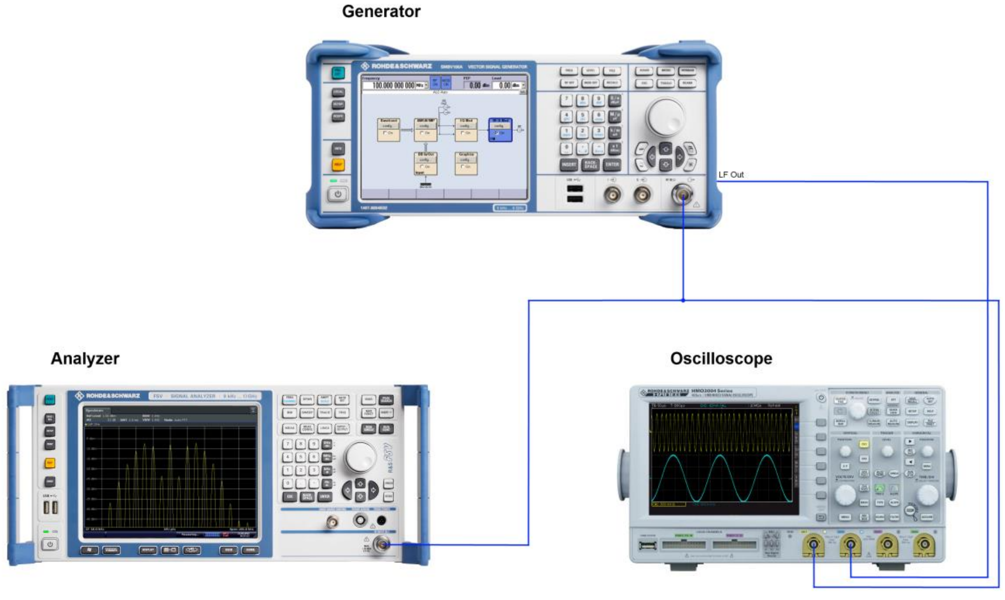


Fig. 3-7: Hardware setup for measuring FM signals.

1.1 Measurement

This first measurement uses the typical VHF FM broadcast transmitter settings for frequency modulation.

► Set up the function generator as follows:

Carrier frequency	*100 MHz*
Signal frequency	*15 kHz*
Amplitude	*0 dBm*
FM Mode	*normal*
Frequency deviation	*75 kHz*
LF Gen Shape	*Sine*

These settings are made in "RF/A Mod" > "Frequency Modulation"; see Fig. 3-8. After making the settings, press the *"TOGGLE" key* to enable them.

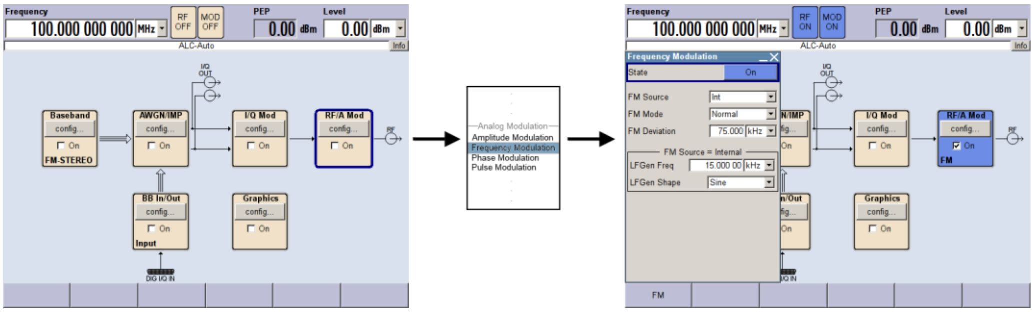


Fig. 3-8: Configuring the generator.

► Make the appropriate settings on the spectrum analyzer. What is the minimum bandwidth that should be displayed?

► How will the spectrum change if the frequency deviation or the wanted signal frequency changes?

► Reduce the wanted signal frequency in small steps down to about 13.7 kHz. What do you see? Where in the Bessel spectrum can be signal be placed?

► Return the signal frequency to 15 kHz.

Frequency and phase modulation are very closely related. The next exercise will show the differences and commonalities for the two modulation methods.

► Calculate the modulation index / phase deviation for the current settings.

► Switch to phase modulation by following the steps in Fig. 3-8 and using "*Phase Modulation*" as the mode.

► Enter your results for the modulation index under *"PhiM Deviation"*.

The spectrum should not have changed, in spite of the change in the modulation mode.

► Reduce the wanted signal frequency again. What differences do you note as compared to frequency modulation, and why?

Switch back to frequency modulation using the settings from Fig. 3-8.

The next step is to assess the simplest form of digital frequency modulation.

► In the "Frequency Modulation" screen, set "LF Gen Shape" to "Square"

► Vary the frequency deviation for the modulation and observe the resulting changes in the spectrum. What peculiarities do you notice with large and small deviations?

► What bandwidth is required and what is the modulation mode for the signal?

In order to clearly display the frequency modulation on the oscilloscope, parameters must be used that are unrealistic in practical applications.

► Set up the function generator as follows:

Carrier frequency	*100 kHz*
Signal frequency	*5 kHz*
Amplitude	*0 dBm*
FM Mode	*normal*
Frequency deviation	*30 kHz*
LF Gen Shape	*Sine*

To display the wanted signal on the oscilloscope, switch on the LF generator output on the SMBV.

► Under "RF/A Mod" > "LF Generator / Output", set the "LF Output State" option to ON and set the level to 500 mV; see Fig. 3-9

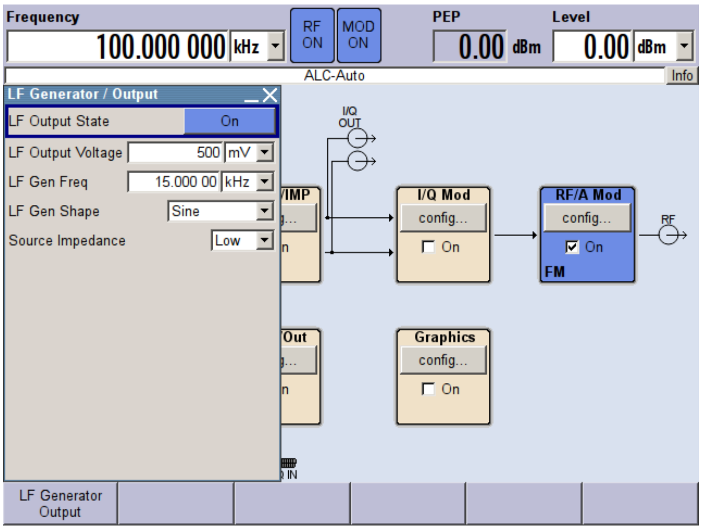


Fig. 3-9: Configuring the LF generator.

► Observe the signal on the oscilloscope.

► Switch back to a square-wave signal.

In this next exercise, you will modulate an FM stereo transmitter. To do this, configure the signal generator as follows:

► Press the "Preset" key to initialize the generator.

► Use the "TOGGLE" key to enable all blocks, then go to "Baseband" > "Broadcast Standards" > FM-STEREO and enter the settings as shown in Fig. 3-10.

Alternatively, you can specify a .wav file under "Audio Source".

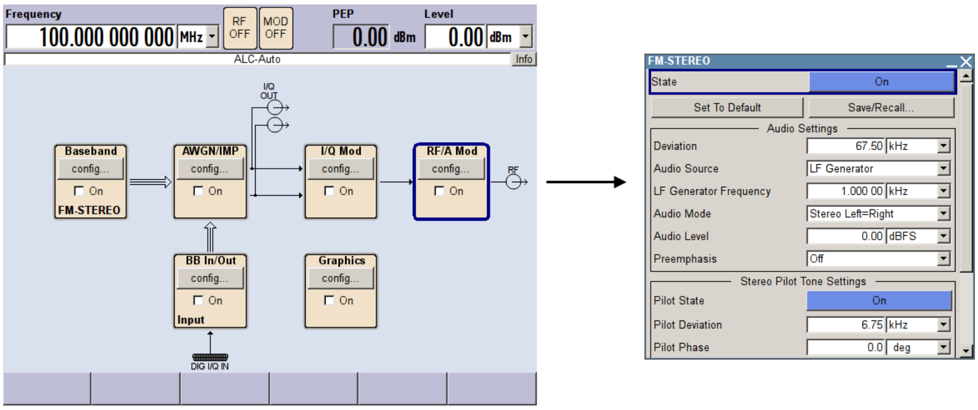


Fig. 3-10: Configuring an FM stereo transmitter.

If you are using an FSV signal analyzer with the appropriate options, the FM stereo signal can be demodulated and measured at the same time; see Fig. 3-11.

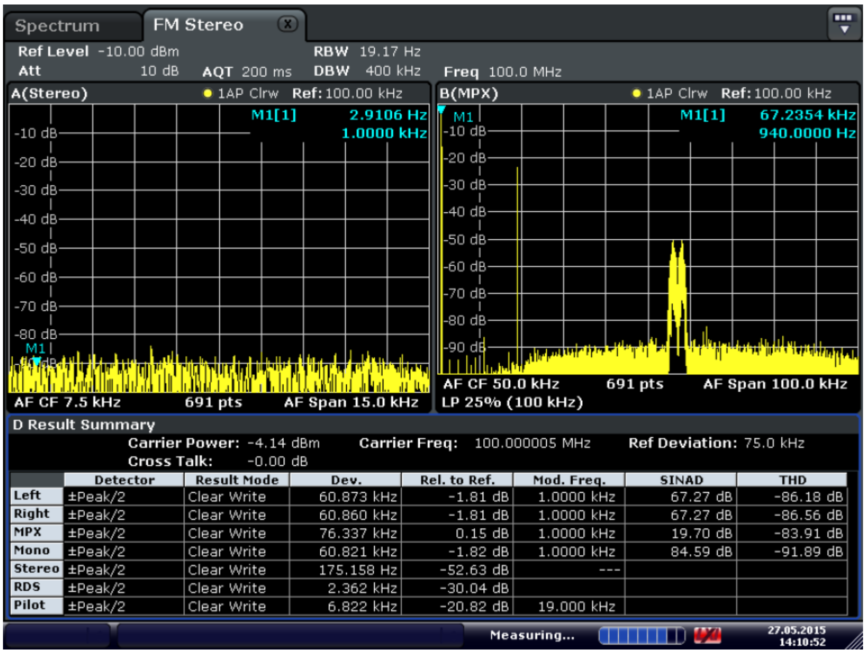


Fig. 3-11: Demodulating an FM signal.

► Press the "*MODE*" button and enable FM stereo measurements.

The FM stereo signal should now be audible and overlaid with noise.

► Switch back to the spectrum display and configure the system analyzer as follows:

Frequency	*100 MHz*
Frequency range	*Zero Span*

► Insert a marker at 100 MHz and then enable the "Marker Demod" → FM function by pressing the "*MKR FUNCT*" button.

The signal should be clearly audible over the headset or loudspeaker. The immunity of FM signals can be tested with artificially added noise.

► Set the generator output level to –15 dBm.

► Under "AWGN/IMP" > "AWGN…", make the settings as shown in Fig. 3-12.

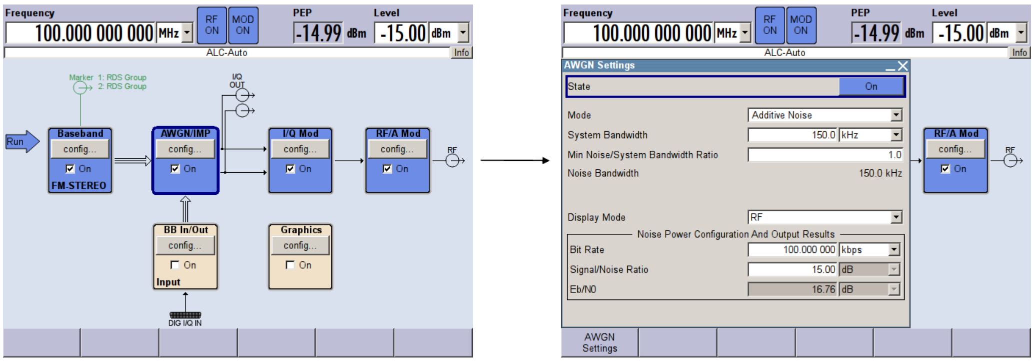


Fig. 3-12: Overlaying additive noise.

► Verify that the audio signal is still audible after being demodulated.

At this point, it is convenient to have selected a .wav file as the audio source in the generator.

► Change the AWGN settings and determine the minimally required SNR.

3.3 Digital Modulation Exercises

The objective for this lab exercise is to generate and observe digitally modulated signals in the time and spectral domain. It also includes descriptions of current display and measurement methods, such as I/Q diagrams and BER measurements.

3.3.1 Test Setup

The test setup consists of a generator and a signal analyzer. This exercise uses the signal analyzer to measure and analyze several different modulation signals produced using the generator.

In order to be able to generate and measure the modulation signals, the analyzer must be connected to the generator via the external trigger (Marker Out 1 / Ext. Trigger In) and the external reference input (Ref Out / Ref In) as shown in Fig. 3-13.

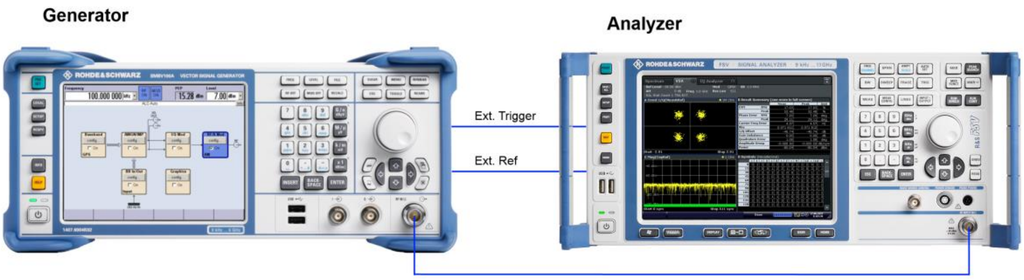


Fig. 3-13: Hardware setup for measuring digitally modulated signals.

3.3.2 Instrument Configurations

Both instruments must be initialized before starting the measurements.

► Press the green "*PRESET"* key on each of the instruments.

This first exercise analyzes QPSK modulation.

Preparing the generator

► Set up the function generator as follows:

Frequency	*1 GHz*
Level	*–30 dBm*
Data Source	*PRBS*
PRBS Type	*PRBS 9*
Standard	*GSM (automatically set -> User)*
Symbol rate	*1 Msym/s*
Coding	*Off*
Modulation Type	*PSK -> QPSK*
Filter	*Root Cosine*
Roll Off Factor	*0.1*

These settings are made under "*BASEBAND*" > "*Custom Digital Mod…*"; see Fig. 3-14.

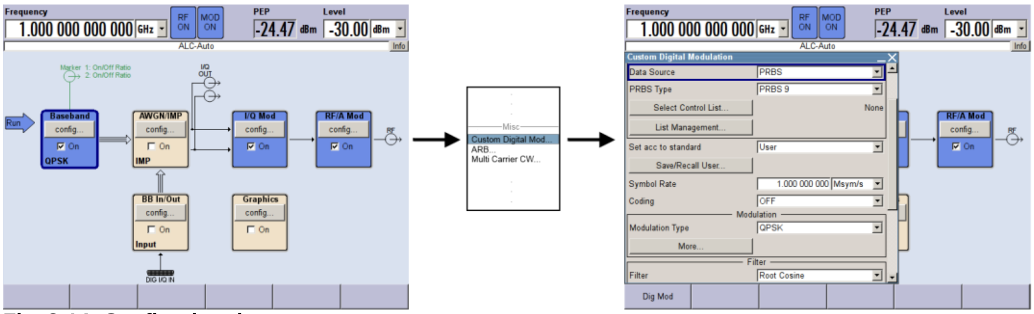


Fig. 3-14: Configuring the generator.

The external reference, external trigger and several synchronization options must be configured in order to perform precision measurements.

Data source PRBS9 generates a random bit sequence with a length of $2^9 - 1$. The generator sends all 511 bits to an external trigger signal at once.

► In the *Custom Digital Modulation* submenu, select the *Trigger/Marker …* option and make the following settings; see Fig. 3-15 :

Mode	*Auto*	
Marker 1:	*On Time*	*1 sym*
	Off Time	*510 sym*

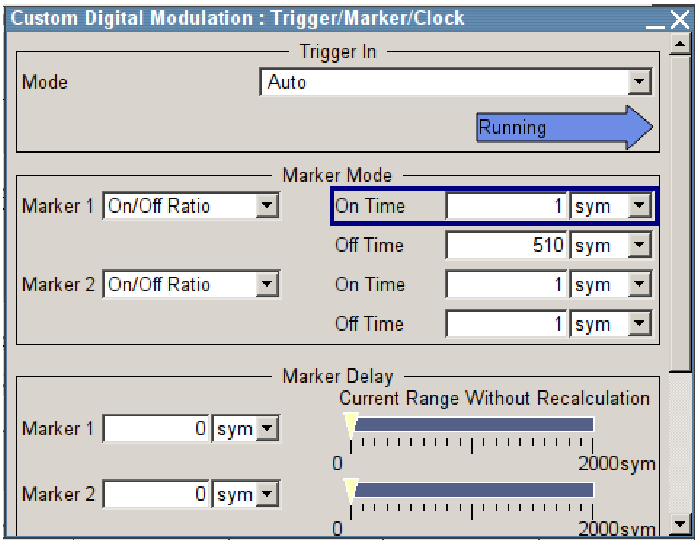


Fig. 3-15: Setting the external trigger.

Preparing the Analyzer

Enable *VSA* (vector signal analyzer) mode on the analyzer.

► Press the *Mode* hardkey and then click the screen to switch to *VSA* mode.

The analyzer must additionally be connected to an external reference:

► Press the *Setup* hardkey and then under *Reference,* select the *Ext* option.

Pressing the *MEAS* hardkey displays several softkeys relevant for the measurements on the screen.

► Select the *Settings Overview* softkey; see Fig. 3-16.

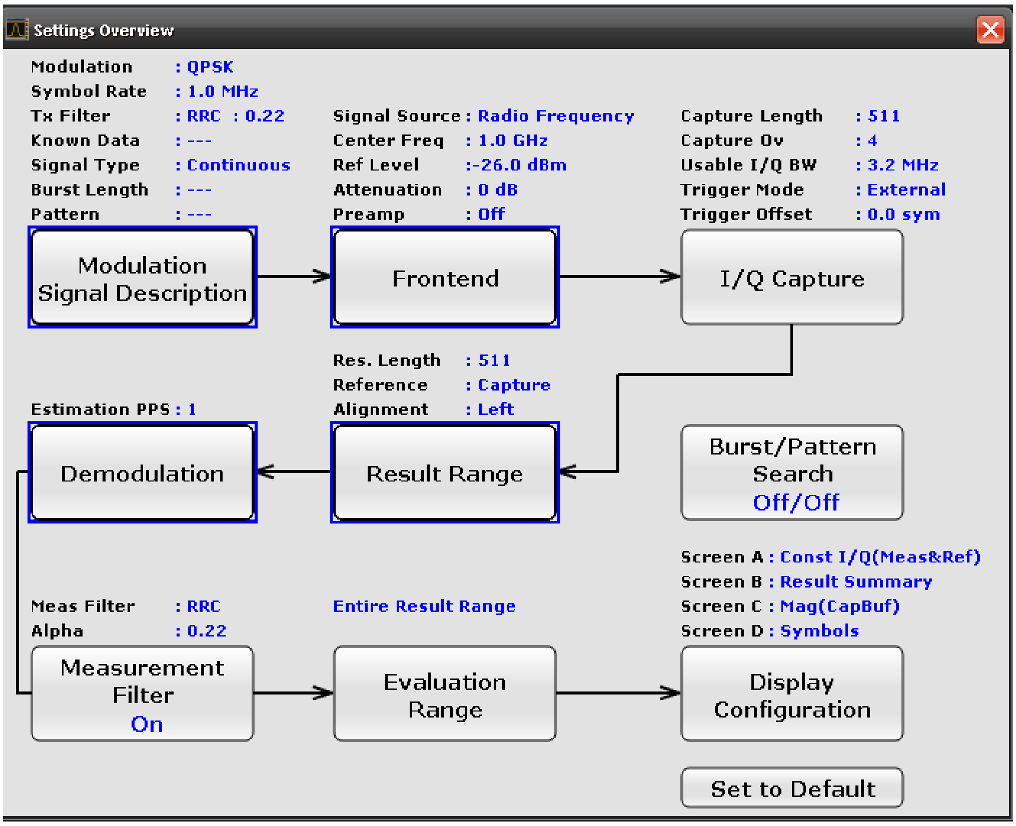


Fig. 3-16: Configuring the analyzer.

This screen provides an overview of the entire processing chain.

► Configure the following options in the *Modulation Signal Description* block:

Modulation Settings:	
Type	*PSK*
Order	*QPSK*
Mapping	*Natural*
Symbol Rate	*1.0 MHz*
Transmit Filter:	
Type	*RRC*
Alpha/BT	*0.1*

► Configure the following option in the *Frontend* block:

Frequency	*1.0 GHz*

► Configure the following options in *Frontend* > *I/Q Capture*:

Trigger Settings:	
Trigger Mode	*external*
Data Capture Settings:	
Capture Length	*511 sym*

► Configure the following option in *Demodulation* > *Measurement Filter*:

Alpha/BT *0.1*

To ensure that all 511 bits are analyzed, the *Result Range* must be configured appropriately.

► Configure the following parameter in *Result Range* > *Measurement Filter*:

Result Length *511*

3.3.3 Measuring Digital Signals

With the current configuration, the generator transmits random QPSK-modulated data to the analyzer, which repeats every 511 bits.

In order to perform the measurement, you must first enable all blocks in the generator.

► To do this, use the *TOGGLE* key.

A fixed image should be displayed on the analyzer, similar to that shown in Fig. 3-17.

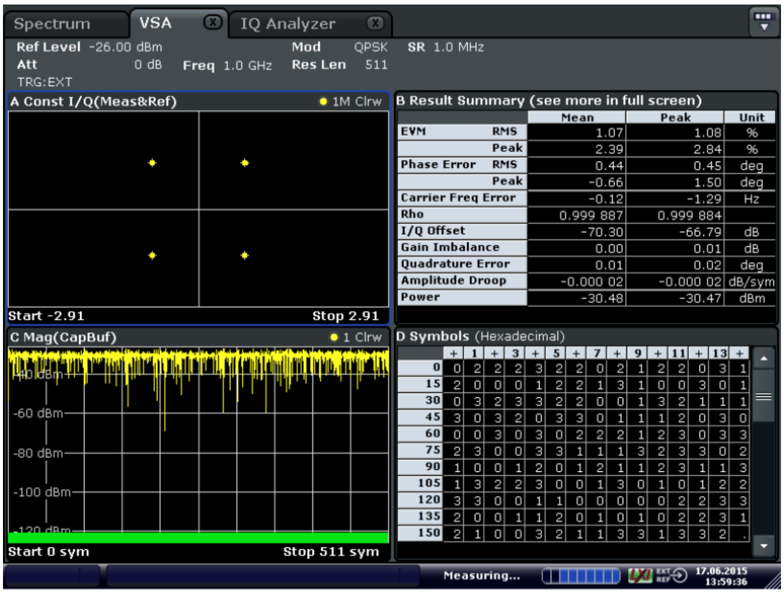


Fig. 3-17: QPSK measurement.

The analyzer splits the display into the following four panes (from left to right) with the current settings:

- The constellation diagram showing the symbols as points in the complex layer.
- A table with the various measurement results.
- The signal amplitudes of the 511 captured symbols.
- A table with the decoded symbols.

Observe the spectrum for the transmission.

► Open the *Spectrum* tab. What bandwidth would have to be displayed for an optimal representation?

► On the generator, change the rolloff factor for the filter. How does this change affect the spectrum, and what effects can be expected for the signal transmission?

Transitions between the possible states are often of interest as well. The *IQ Analyzer* displays all measured signal states.

► Press the *Mode* hardkey and select *IQ Analyzer*.

► Under *Display Config*, select the *IQ Vector* option and then use the *Level* function to optimally adjust the results. Your display should look something like Fig. 3-18.

The effects of the rolloff factor are not limited to the spectrum; the influence of these variables is also clearly visible in the I/Q vector display.

► What changes would you expect at various rolloff factors?

► Change the factor to various positions across the entire range and verify your assumptions.

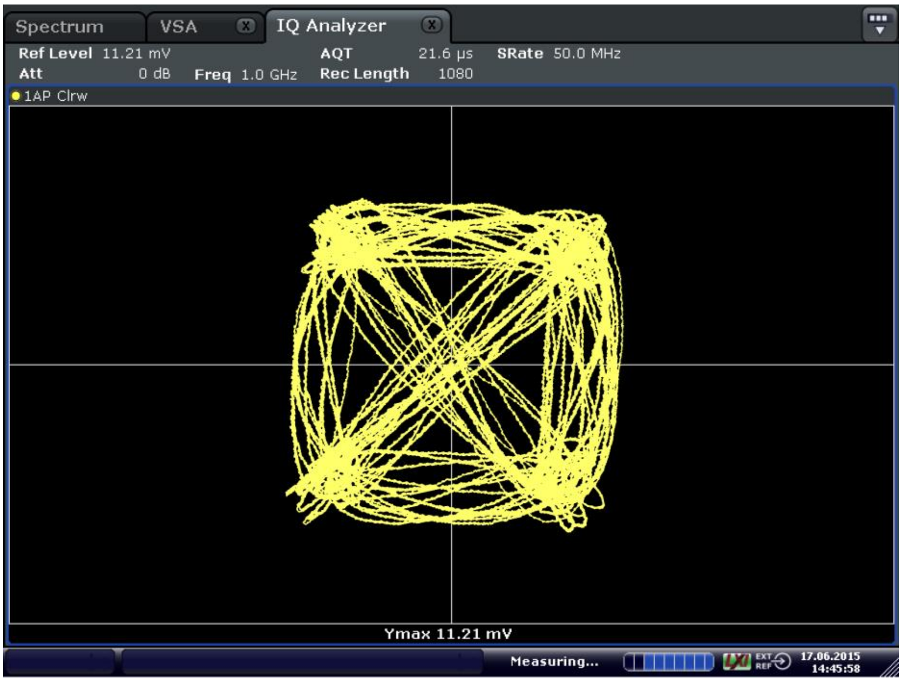


Fig. 3-18: I/Q diagram for QPSK.

Influence of I/Q impairments during QPSK modulation

The R&S® SMBV generator supports a wide variety of options for influencing the signal quality.

► What would be the result of switching on the quadrature offset?

► On the generator, go to *I/Q Mod* > *I/Q Settings.* Set the options as follows:

Impairments:	
State	☒ *On*
Quadrature Offset	*10 deg*

Note that the signal analyzer does not automatically compensate for the offset.

► Open the *Settings Overview* menu (if the option is not displayed as a softkey, press the *MEAS* hardkey once). Disable the option in the *Demodulation* submenu.

► Is the signal still recognizable with these settings? Change to VSA mode and verify your assumptions.

► Pay special attention to the effects on the EVM measurement.

► Set the quadrature offset back to *0 deg.*

The next exercise shows how to add an I/Q offset.

► Make the following settings and observe their effects on the signal in VSA and I/Q analyzer mode:

Impairments:	
I Offset	*10 deg*
Q Offset	*10 deg*

► Which methods are particularly impacted by these changes?

Levels in the order of –30 dBm are rarely seen at receivers during transmissions over the air interface. The modulation methods are therefore designed for significantly worse conditions. The next exercise replicates transmission under more realistic conditions.

► Set the I and the Q offsets back to 0.

Reduce the transmission level to –70 dBm.

► Press the *Level* key, enter –70, and confirm with the *x1/ dB(m)* key.

Along with the lower receive power, you will also apply additive noise.

► Change to *AWGN/IMP* > *AWGN…*; see Fig. 3-19.

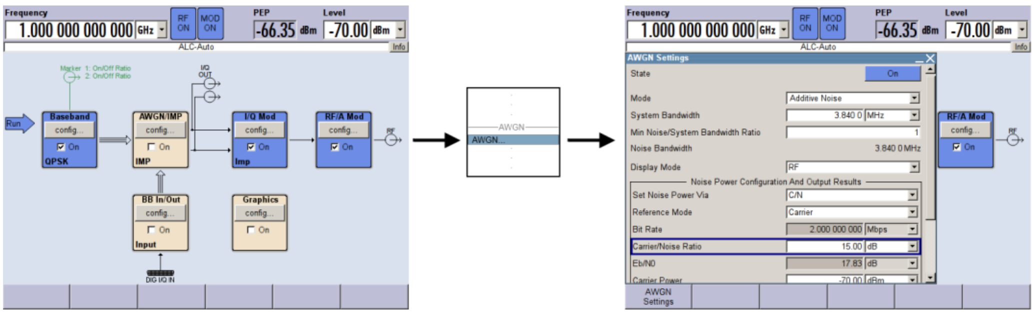


Fig. 3-19: Configuring the AWGN.

Set the options as follows:

State	*On*
Carrier/Noise Ratio	*15 dB*

► Observe the effects on the constellation diagram and on the spectrum.

► How low does the SNR need to be for this transmission? Check the result against your assumptions.

► Can interference-free transmission still be expected for a signal as shown in Fig. 3-20. What exactly does the EVM measurement show, and where can this value be found?

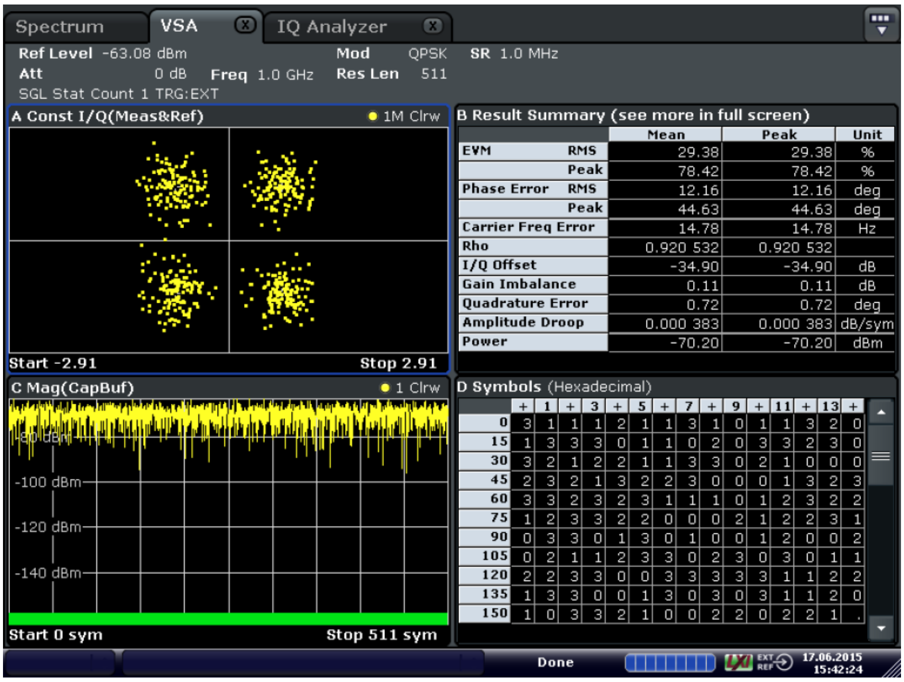


Fig. 3-20: Low SNR with QPSK.

Frequently, an RRC filter is used during transmission to suppress intersymbol interference (ISI). This filter is typically split into two parts in order to achieve the best possible noise suppression: One filter is located in the transmitter and one in the receiver, and the two combined provide a raised cosine transfer function.

► Does changing the rolloff factor affect the EVM measurement? What factor provides the best result readings, and why is this factor not used in practice?

► Test other filters and their behavior with interference. Make note of the required settings on the transmitter and the receiver end.

► Switch back to the RRC filter.

Using QPSK modulation places special demands on the amplifier that are particularly disadvantageous for mobile devices.

► What are the disadvantages of QPSK and how can they be circumvented?

$\frac{\pi}{4}$-QPSK inserts an additional phase shift after every transmit symbol in order to compensate for the disadvantages of QPSK.

Bring the transmission level to back up to –30 dBm.

► Set the $\frac{\pi}{4}$QPSK modulation mode on both the generator and the analyzer.

► Observe how the modulation behaves when *IQ Analyzer* mode is set on the analyzer; see Fig. 3-21. (The transitions can be better seen when higher signal levels are used.)

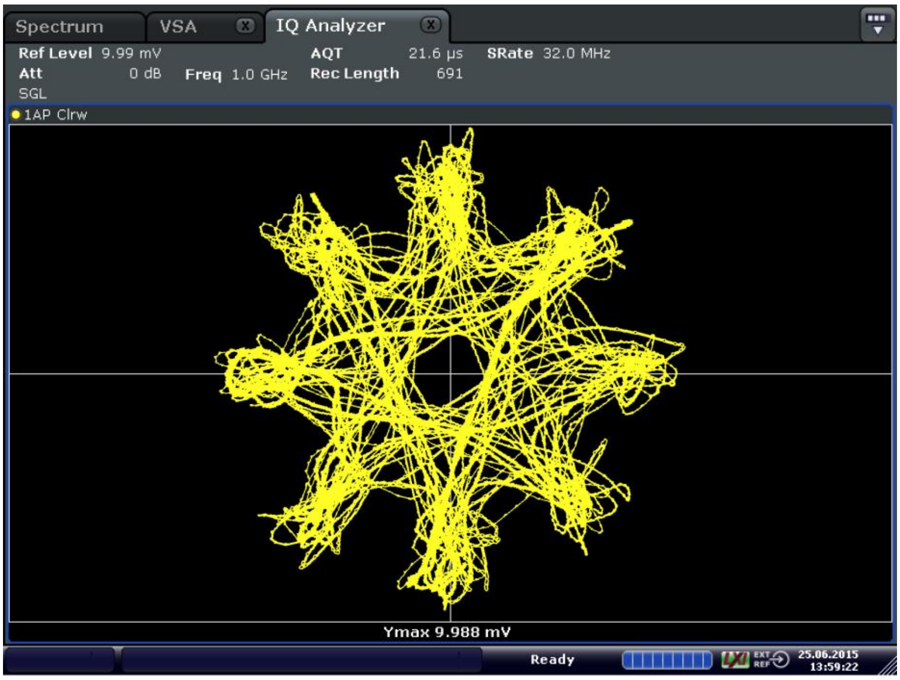


Fig. 3-21: I/Q diagram of $\frac{\pi}{4}$-QPSK.

► Is there a discernible difference in the immunity to noise?

QAM signals

With QAM modulation, the available signal space is improved by adding signal states.

► Set modulation mode 16-QAM on both the generator and the analyzer.

► Add the same I and Q offsets as used above for QPSK. Evaluate the results; see also Fig. 3-22.

► Test a variety of SNR values and compare them against the result gained during the QPSK measurements.

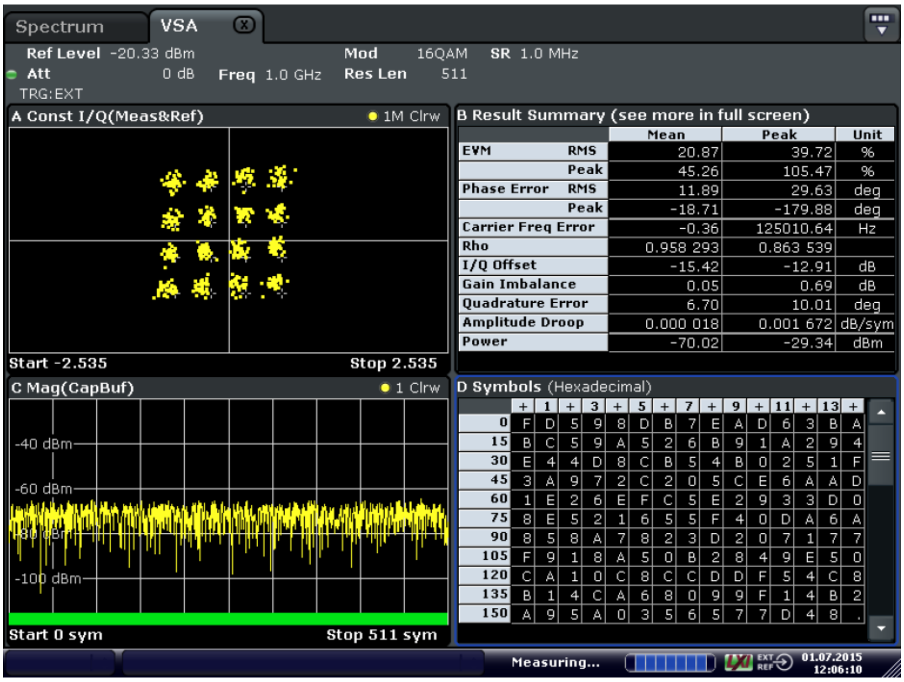


Fig. 3-22: 16-QAM with I and Q distortions.

This next exercise shows how to work with data lists, which can replace random sequences. Data lists provide for better synchronization and unambiguous demodulation, and can also be used for bit error rate measurements.

In this section, you will create a data list on the generator and demodulate the data using the signal generator and an appropriate synchronization sequence.

► In the *Custom Digital Modulation* screen, select submenu *Data Source* and then the *List Management* option.

► Press *Select Data List to Edit* > *New List,* then select a memory location and assign a name for the new list.

Select the new list and fill it with 32 hexadecimal values; see Fig. 3-23.

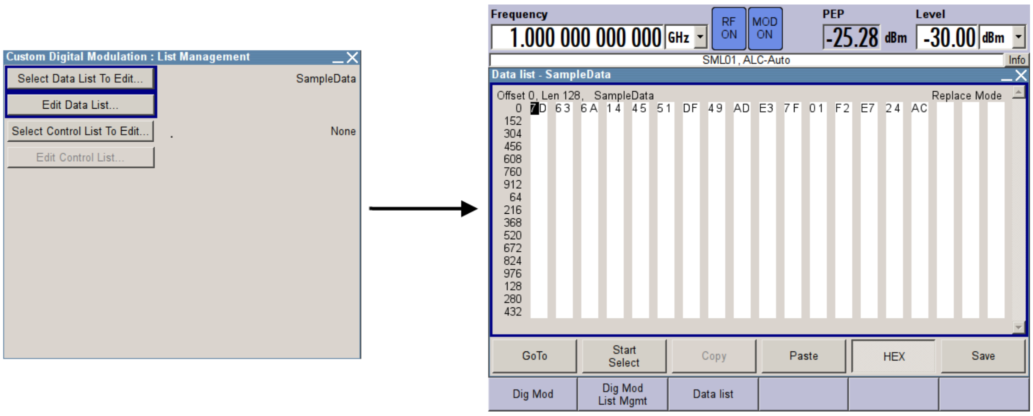


Fig. 3-23: Creating a data list with 32 hexadecimal values.

Because the bit sequence is known, there is no longer any need for an external trigger.

► In submenu *I/Q Capture*, switch to trigger mode *Free Run*; see Fig. 3-16.

► Set a *Capture Length* of 256 symbols.

By using a known pattern, it is possible to demodulate the QPSK-modulated signal unambiguously. This requires that a segment of the data be supplied to the analyzer.

► Press the *MEAS CONFIG* hardkey and select submenu *Config Pattern*.

► Under *All Patterns*, select *New* to create a new pattern. Select a segment of the previously created list and fill in the remaining fields as appropriate; see Fig. 3-24.

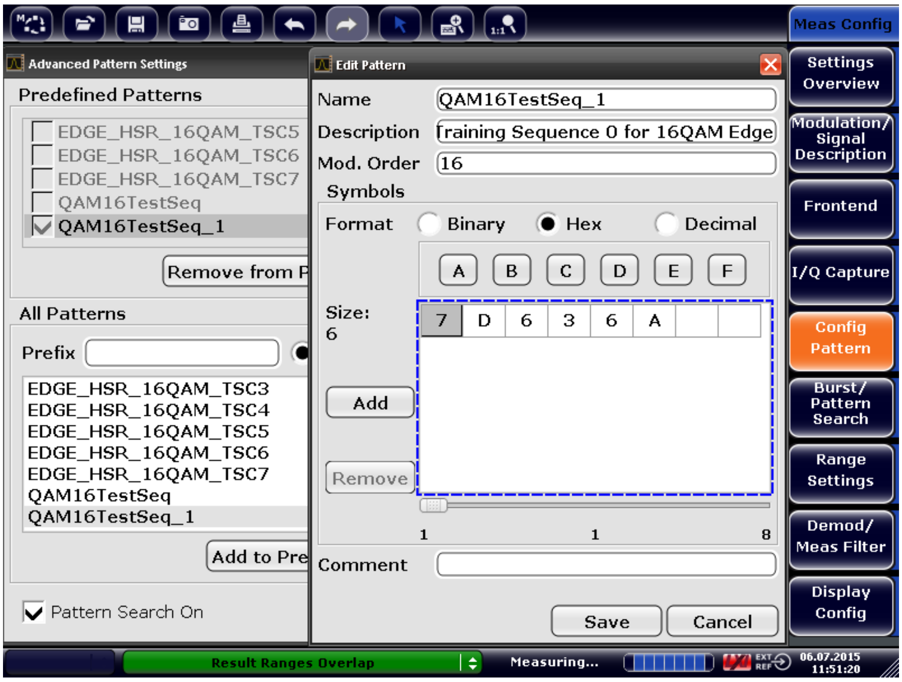


Fig. 3-24: Creating a known pattern.

► Save the pattern and apply the settings to the *Predefined Patterns* column.

► Use *Burst/Pattern Search* to load the saved pattern. The analyzer should detect the bit sequence in the signal and synchronize it correctly; see Fig. 3-25.

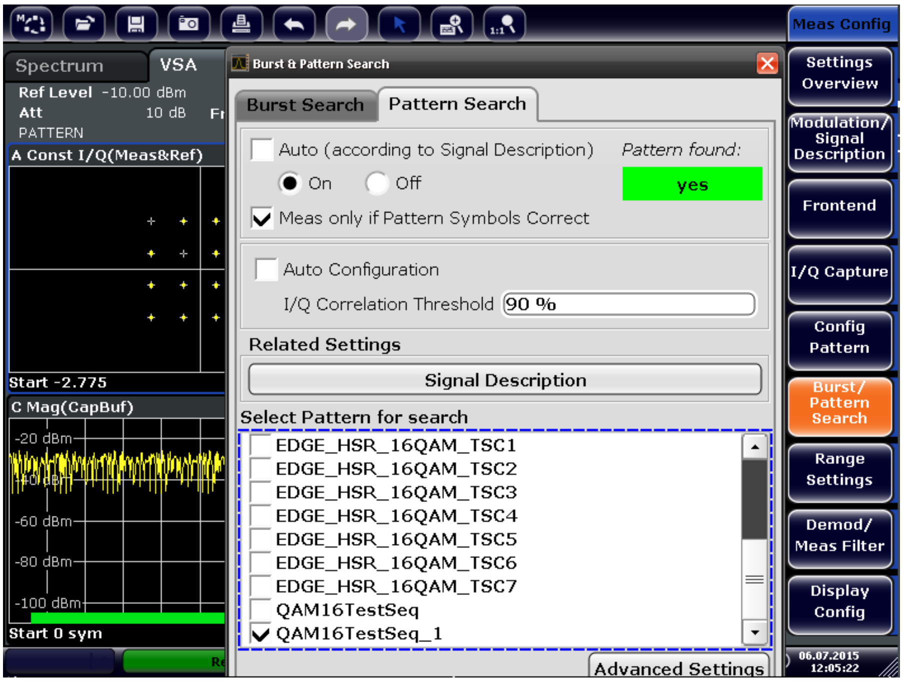


Fig. 3-25: Synchronizing to a bit pattern.

Different patterns have a varying degree of effectiveness for synchronization purposes.

► How can the pattern be optimized?

► Add noise in the generator and test various patterns and correlation settings; see Fig. 3-25.

BER measurements

Once the content of the data signal is known, it is possible to perform a bit error rate measurement in order to assess the quality of the transmission.

To do this, a file in .xml format that contains information about the signal characteristics must be provided to the analyzer. The following example is used for the pattern from Fig. 3-23:


```
<RS_VSA_KNOWN_DATA_FILE
Version="01.00">
<Comment> 16 QAM for BLER Testing </Comment>
<Base>16</Base>
<ModulationOrder>16</ModulationOrder>
<ResultLength>32</ResultLength>
<Data>7D 63 6A 14 45 51 DF 49 AD E3 7F 01 F2 E7 24 AC</Data>
</RS_VSA_KNOWN_DATA_FILE>
```

Load the file into the R&S® FSV.

► In the *Modulation Signal Description* menu, use the *Known Data* option to load the data.

Complete the remaining settings as shown in Fig. 3-26.

► The *QAM16TestSeq* pattern includes the full contents for the transmission ($7D_h$ … AC_h).

► Configure the BER measurement under *Display Configuration* > *Modulation Accuracy* > *Bit Error Rate*.

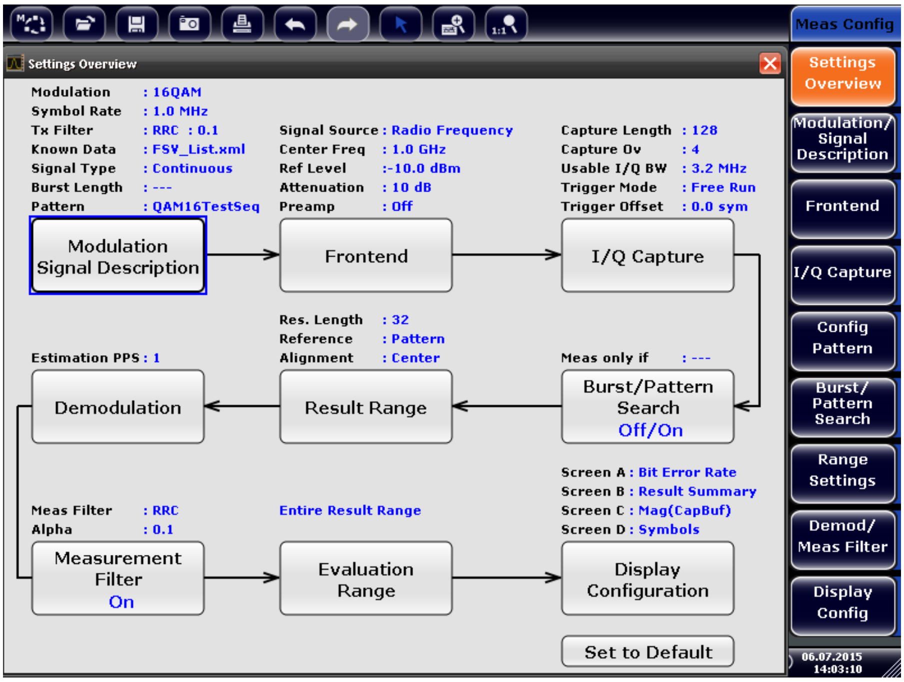


Fig. 3-26: Configuring the BER measurement.

At low noise, the results will be as shown in Fig. 3-27.

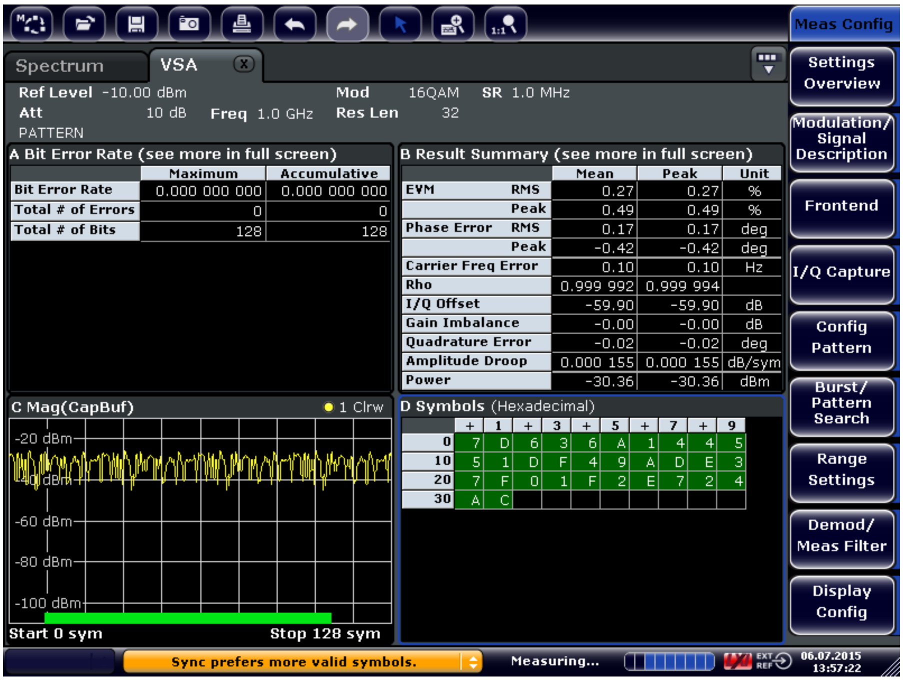


Fig. 3-27: BER measurement under good conditions.

UMTS signals

In this last set of exercises, a simple UMTS cellular signal will be generated and several parameters measured. The objective is to gain an understanding of the spectrum and the basic structure of WCDMA signals.

Configure the signal generator for UMTS signals:

► In the *Baseband* block, open *3GPP FDD* (found under *CDMA Standards*).

► Use *State* > *On* to enable the generation of UMTS signals.

► Increase the frequency to 2110 MHz (band 1).

The generator produces downlink UMTS signals from a simulated base station.

Ensure that the analyzer is set to the correct frequency and observe the spectrum of the WCDMA transmission.

The analyzer must be in *3G FDD BTS* mode in order to measure a UMTS base station.

► Press the *MODE* hardkey to change to *3G FDD BTS* mode.

Before starting a measurement, you should familiarize yourself with the possible analyzer settings in this mode.

► Use the *Settings Overview* softkey to open the summary display; see Fig. 3-28.

The first measurement (Fig. 3-29) displayed when you change to *3G FDD BTS* mode shows the level for all detected channels (yellow) in the selected slot. Codes that cannot be allocated to a channel are shown in turquoise.

► What is the purpose of the *Channelization Codes* and the *Scrambling Codes*?

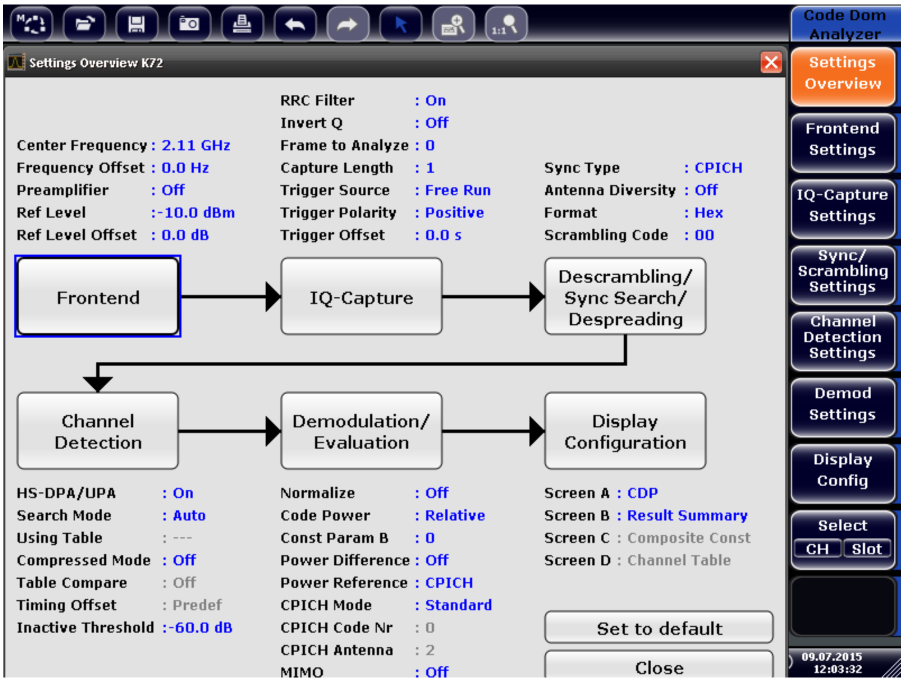


Fig. 3-28: Overview of UMTS settings.

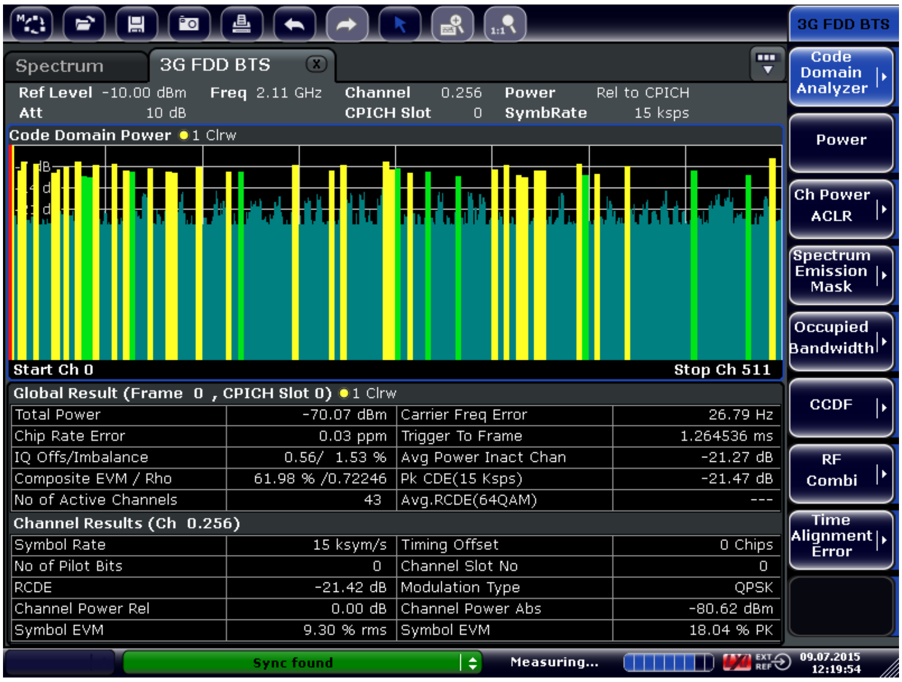


Fig. 3-29: Display of all detected channels and their level.

The R&S®FSV also allows simultaneous measurement of the levels of adjacent channels (ACLR), the allocated bandwidth and spectrum masks (SEM).

► Start the *RF Combi* measurement.

The channel being used (TX1; blue) is displayed in the center and the adjacent channels (Adj, Alt1) are displayed to either side. The channel spacing is 5 MHz and the signal bandwidth is 3.84 MHz for each channel. The analyzer measures the average signal level for each channel and displays the results in a table.

The spectrum and the associated masks make it possible to check the conformity of the UMTS signal.

4 Literature

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