Chapter 25: Transmitters and Receivers

This chapter describes the design of transmitters and receivers for radio transmission. The terms used shall have a defined meaning such that the components from the modulator up to the transmitting antenna form the transmitter, while the components from the receiver antenna up to the demodulator form the receiver.

The demands placed on the transmitter and receiver are clearly distinct since the transmitter must process only the desired signal while the receiver must separate the desired signal from the frequency mixture received by the antenna. Furthermore, the transmitter handles signal levels which are constant or which vary very slightly, while the receiver copes with extremely large level differences that depend on the distance to the transmitter. The main challenges for the transmitter include the task of converting the useful signal into a high-frequency transmission signal with as little interference as possible, to amplify this signal with the highest possible efficiency and to minimize the transmission of undesirable interference signals generated by the conversion or amplification. The main challenges for the receiver are to filter out the desired signal even from very weak levels, while at the same time receiving very strong signals from adjacent frequency ranges, and producing a clear signal with a high signal-to-noise ratio and minimum intermodulation distortions. Thus, the main obstacle for concern in transmitters is efficiency, while receivers face issues of selection, dynamics and noise.

25.1 Transmitters

First we will look at the construction of transmitters with analog modulation, followed by a description of transmitters with digital modulation. These descriptions are supported by simplified block diagrams showing only the essential components.

25.1.1 Transmitters with Analogue Modulation

Transmitters with Direct Modulation

The most simple transmitter is obtained when the carrier frequency $f_C$ of the analog modulator is identical to the transmission frequency $f_{RF}$. In this case, the modulator output signal only needs to be amplified and fed to the antenna. In practice, the transmission amplifier must be followed by an output filter that reduces the distortion products originating in the amplifier to an acceptable level. Figure 25.1a shows the construction of a transmitter with direct modulation. The signal spectra are shown in Fig. 25.2.
Transmitters and Receivers

With direct modulation

With intermediate frequency

With two intermediate frequencies

Fig. 25.1. Transmitter with analog modulation

Fig. 25.2. Signal spectra in transmitters with direct modulation
Transmitters with One Intermediate Frequency

With increasing frequencies and growing demands, it becomes more and more difficult to obtain a modulator with the required accuracy. Therefore, a lower intermediate frequency $f_{IF}$ with which the modulator can be easily built is used as carrier frequency $f_C$:

$$f_C = f_{IF} \ll f_{RF}$$

Figure 25.1b shows the construction of a transmitter with one intermediate frequency. Conversion to the transmission frequency $f_{RF}$ is done by mixer M1 which is provided with the frequency

$$f_{LO} = f_{RF} - f_{IF}$$

from a local oscillator (LO). The mixing process generates the sum and difference frequencies

$$f_{LO} + f_{IF} = f_{RF}, \quad f_{LO} - f_{IF} = f_{RF} - 2f_{IF}$$

The portion at the transmission frequency is filtered by an RF filter and fed to the transmitter amplifier. Figure 25.3 shows the signal spectra.

Owing to $f_{RF} = f_{LO} + f_{IF}$, the frequency sequence is identical in the IF and RF signals, which means that a higher IF frequency results in a higher RF frequency; this is known as noninverted mode. It is also possible to choose $f_{RF} = f_{LO} - f_{IF}$ by filtering out the signal

![Signal spectra in transmitters with one intermediate frequency](image)

Fig. 25.3. Signal spectra in transmitters with one intermediate frequency
portion below the local oscillator frequency in Fig. 25.3. Then the frequency sequence in the transmission signal is inverted; this is known as inverted mode. The receiver must take the inverted frequency operation into account in order to correctly receive the desired signal. For this purpose, the receiver uses a mixer operated in inverted mode.

The mixer output signal contains a signal portion at the local oscillator frequency $f_{LO}$ (see Fig. 25.3). Consequently, the transition region of the RF filter (transition from the pass band to the cutoff band) must not exceed the width $f_{IF} - B/2$ to ensure that the transmission signal lies fully within the pass band and the local oscillator signal is in the cutoff band. Particularly suitable are surface acoustic wave (SAW) filters with their very narrow transition region and constant group delay but whose high insertion loss (>20 dB) is disadvantageous. Where no SAW filters are available for the desired transmission frequency, LC filters or filters with dielectric resonators must be used. As these filters have unwanted group delay distortion at the borders of the transition region, it is necessary to select a clearly smaller transit region in order to prevent the transmission signal from being affected. As an alternative, one may use the entire range between the portions above and below the local oscillator frequency as the transition region and suppress the local oscillator frequency by a separate serial or parallel resonant circuit (zero transmission at $f_{LO}$).

With rising transmission frequencies, the ratio of the transmitter frequency to the width of the transition region increases; hence, the quality of the RF filter must also increase:

$$Q_{RF} \sim \frac{f_{RF}}{f_{IF} - B/2} \approx \frac{f_{RF}}{f_{C}}$$

This results in a higher filter order and increased group delay distortions. In practice, the intermediate frequency is made as high as possible so that the transition region becomes wider and the RF filter quality becomes correspondingly low.

**Transmitters with Two Intermediate Frequencies**

In transmitters with one intermediate frequency and high transmission frequencies, the quality of the RF filter becomes impermissibly high. A second intermediate frequency is then required that ranges between the carrier frequency of the modulator and the transmission frequency:

$$f_{C} = f_{IF1} < f_{IF2} < f_{RF}$$

Figure 25.1c shows the construction of a transmitter with two intermediate frequencies, while the signal spectra are presented in Fig. 25.4. Mixer M1 converts the modulator’s output signal from the first to the second intermediate frequency. This requires a local oscillator with the frequency $f_{LO1} = f_{IF2} - f_{IF1}$. Subsequently the portion above the local oscillator frequency is filtered out by an IF filter. The quality of the IF filter is proportional to the ratio of the second intermediate frequency and the width of the transition region:

$$Q_{IF} \sim \frac{f_{IF2}}{f_{IF1} - B/2} \approx \frac{f_{IF2}}{f_{C}}$$
Fig. 25.4. Signal spectra in transmitters with two intermediate frequencies

The conversion to the transmission frequency is achieved with a mixer $M2$, which is fed by a second local oscillator with the frequency $f_{LO2} = f_{RF} - f_{IF2}$. An RF filter of the quality

$$Q_{RF} \sim \frac{f_{RF}}{f_{IF2} - B/2} \quad \frac{f_{IF2} \gg B}{\approx} \quad \frac{f_{RF}}{f_{IF2}}$$

is required to filter out the transmission signal.
Obviously the overall quality is \( Q \approx f_{RF}/f_C \), which, in transmitters with one intermediate frequency, has to be generated by the RF filter and in transmitters with two intermediate frequencies can be distributed to two filters:

\[
Q = Q_{RF} Q_{IF} \sim \frac{f_{RF}}{f_C}
\]

The relative amounts can be controlled by the value of the second intermediate frequency, more specifically, if it is relatively high then \( Q_{IF} > Q_{RF} \), if it is relatively low then \( Q_{IF} < Q_{RF} \). In practice, the values selected depend on the transmission frequency and the available filters. The planned number of units also has an important influence since for high unit numbers customized dielectric or SAW filters can be used, but for mass applications such as mobile communication even the design of new filter technologies is warranted. For small batch production, on the other hand, standard filters are used. The use of LC filters with discrete components is avoided where possible for reasons of space and calibration.

In transmitters with two intermediate frequencies, one can also operate one or both mixers in inverted mode by filtering out the portions below the local oscillator frequency. If both mixers are operated in inverted mode, then the transmission signal is in noninverted mode again.

**Transmitters with Variable Transmission Frequencies**

In transmitters with a variable transmission frequency, the frequency of the last local oscillator is variable, thus allowing the transmission frequency to be altered without affecting the other components. Variations take place within the frequency range assigned to the specific application according to the channel spacing \( C \). Figure 25.5 illustrates this taking a transmitter with five channels as an example. The RF filter is rated such that all channels are within the pass band and all local oscillator frequencies are within the cutoff band. Alternatively, a tuneable RF filter may be used, but only in exceptional practical cases.

For a lower number of channels and less channel spacing, the local oscillator and transmission frequencies change very little. For such applications a transmitter with one intermediate frequency can be used as long as the transition region between the highest local oscillator frequency and the lowest limit of the channel pattern is sufficiently wide. Although in most cases this requires a transmitter with two intermediate frequencies where the second intermediate frequency is selected relatively high so that the transition region becomes as wide as possible.

![Fig. 25.5. Transmitter with variable transmission frequency](image)
25.1.2 Transmitters with Digital Modulation

In principle, transmitters with digital modulation are of the same design as transmitters with analog modulation. The essential difference is that digital modulators primarily generate the quadrature components \( i(t) \) and \( q(t) \) that are combined into a modulated carrier signal by an I/Q mixer.

Figure 25.6a shows a digital transmitter with direct modulation. It corresponds to the analog transmitter with direct modulation in Fig. 25.1a if the combination of digital modulator, I/Q mixer (MI and MQ) and the subsequent filter are regarded as being equivalent to the analog modulator. The same applies to the digital transmitter with one or two in-
Intermediate frequencies. A digital transmitter with one intermediate frequency is shown in Fig. 25.6b.

If particularly high demands are made in terms of the accuracy of the I/Q mixer, a digital I/Q mixer is used to prevent amplitude and phase errors between the two branches. The output of the digital I/Q mixer provides a digital IF signal that is converted into an analog IF signal by a D/A converter and subsequent IF filter. As the frequency of the IF signal must be comparatively low due to the limited sampling rate of the digital I/Q mixer and the D/A converter, a second intermediate frequency is usually utilized. Figure 25.6c shows the resulting transmitter.

25.1.3 Generating Local Oscillator Frequencies

The required local oscillator frequencies are derived by phase-locked loops (PLL) from a crystal oscillator with reference frequency $f_{REF}$. Figure 25.7 depicts this for a transmitter with one intermediate frequency and variable transmission frequency. The intermediate frequency is fixed and is determined by the divider factors $n_1$ and $n_2$:

$$f_{IF} = \frac{n_2}{n_1} f_{REF}$$

The local oscillator frequency is variable in steps according to the channel spacing $C$. For this purpose, the reference frequency is divided to the channel distance by the divider factor $n_3$ and multiplied by a PLL with the programmable divider factor $n_4$:

$$C = \frac{f_{REF}}{n_3}, \quad f_{LO} = n_4 C = \frac{n_4}{n_3} f_{REF}$$

Fig. 25.7. Generation of the local oscillator frequencies
The local oscillator frequency and thus the transmission frequency is adjusted by changing the divider factor $n_4$. If the local oscillator frequencies are not divisible by $C$, then the reference frequency must be divided by means of the divider factor $n_3$ to the largest common divisor of $C$ and the local oscillator frequencies and this common divisor must be multiplied by $n_4$.

Example: In Fig. 24.81 on page 1225, a QPSK modulator with I/Q mixer is to be converted into a transmitter with one intermediate frequency that is capable of a data rate of 200 kbit/s at a roll-off factor $r = 1$. A crystal oscillator with $f_{\text{REF}} = 10$ MHz is to be used as a reference. The data rate $f_D = 200$ kHz is obtained by division by a factor of 50. The carrier or intermediate frequency is $f_C = f_{\text{IF}} = 70$ MHz since inexpensive SAW filters are available for this frequency. Since the I/Q mixer in Fig. 24.81 must be driven with the frequency $2f_C = 140$ MHz, we select $n_1 = 1$ and $n_2 = 14$ for the IF PLL in Fig. 25.7. For QPSK, the symbol frequency is equal to half the data rate $f_S = f_D/2$, resulting in a bandwidth of $B = (1 + r)f_S = 200$ kHz. We assume that the transmitter can use 4 channels ranging from 433 to 434 MHz with a channel spacing of $C = 250$ kHz. From the transmission frequencies $f_{\text{RF}} = 433.125/433.375/433.625/433.875$ MHz we obtain the local oscillator frequencies $f_{\text{LO}} = f_{\text{RF}} - f_{\text{IF}} = 363.125/363.375/363.625/363.875$ MHz. Since these are not multiples of $C$, we calculate the largest common divisor: $\text{gcd}(K, f_{\text{LO}}) = 125$ kHz. For the LO PLL this leads to $n_3 = 10$ MHz/125 kHz = 80 and $n_4 = f_{\text{LO}}/125$ kHz = 2905/2907/2909/2911. For all channels, the RF filter must allow signal transmission without major group delay distortion and, at the same time, sufficiently attenuate the highest local oscillator frequency. The double-tuned-circuit bandpass filter described in Sect. 26.2 can be set up for a center frequency of 434.4 MHz and a bandwidth of 10 MHz. Thus the desired signal is attenuated by 6 dB, while the local oscillator frequency is reduced by more than 54 dB and the portion below the local oscillator frequency by more than 70 dB.

25.2 Receivers

The receiver has the task of filtering out the desired signal from the antenna signal and amplifying it enough to feed it to the demodulator. In most instances, the receive frequency is variable so that different channels, for example, various radio stations, can be received. As the signal level may vary widely depending on the distance between transmitter and receiver, the receiver must be provided with amplifiers of variable gain and gain control in order to compensate for the different levels of receive signals. Limiting amplifiers that convert the receive signal into a square wave signal and subsequent filtering can be used only for signals from transmitters with pure angle modulation.

First we shall describe receivers for analog modulation in which the receive signal is converted to an intermediate frequency and then demodulated in an analog demodulator (for example, detector for AM and envelope discriminator for FM). Then we shall discuss the expansions to enable the reception of digital modulated signals.
25.2.1 Direct-Detection Receivers

In the pioneer days of radio engineering, the direct-detector receiver shown in Fig. 25.8a was used. The receive signal was filtered by an RF filter and fed directly to the demodulator after a fixed or variable amplification. The RF filter needed tuning in order to receive the signals from different radio stations. The only modulation technology that could be used was amplitude modulation since the envelope detector was the only demodulator that worked satisfactorily with a variable carrier frequency $f_C = f_{RF}$. All other demodulators must be set up for a fixed carrier frequency or require frequency-synchronous tuning according to the RF filter.

Besides being limited to amplitude modulation, the direct-detection receiver has other significant drawbacks:

- The transmission frequency must be no more than two orders of magnitude greater than the bandwidth of the signal to be received; otherwise, the quality of the RF filter becomes too high. In the early days of broadcasting systems, there were only a few stations with significantly differing transmission frequencies. A simple resonant circuit was therefore sufficient to filter out the desired station.
- Tuneable filters of high quality are expensive and can only be tuned to a very limited frequency range if the bandwidth is to be maintained. On the other hand, the resonant circuits used in the early days allowed easy tuning by means of a variable capacitor.
- The entire amplification must be done at the transmission frequency, thus high-frequency transistors with high quiescent currents and relatively low gains must be used.
With increasing frequencies, the performance of envelope detectors decreases due to the parasitic capacitance of the rectifier diode.

With the growing density of transmitting stations and the use of higher frequencies, the direct-detection receiver soon reached its limits.

25.2.2 Superheterodyne Receivers

In the superheterodyne receiver, the tuning of the RF filter is replaced by the frequency conversion from a mixer with variable local oscillator frequency $f_{LO}$. This converts the signal to be received to a fixed intermediate frequency (IF frequency):

$$f_{IF} = f_{RF} - f_{LO} \ll f_{RF}$$

An intermediate frequency filter (IF filter) of a substantially lower quality

$$Q_{IF} \sim \frac{f_{IF}}{B_{IF}} \ll \frac{f_{RF}}{B_{RF}} \sim Q_{RF}$$

is used to filter out the signal. The variable amplification and the demodulation are also done at the IF frequency. Thus, all disadvantages of the direct-detection receiver are eliminated. Figure 25.8b shows the construction of a superhet receiver with one intermediate frequency.

RF Filters

In the process of frequency conversion, not only the desired receive frequency

$$f_{RF} = f_{LO} + f_{IF}$$

but also the image frequency

$$f_{RF,im} = f_{LO} - f_{IF}$$

are converted to the IF frequency (see Fig. 25.9). This causes a region located at the opposite side of the local oscillator frequency to be converted to the pass band of the IF filter. In order to prevent this, the RF filter in front of the mixer must be set up such that all desired receive frequencies are within the pass band and the related image frequencies in the cutoff region (see Fig. 25.10). The RF filter is thus also known as the image filter. In practice, the RF filter is designed such that the local oscillator frequencies are also in the cutoff region. This prevents the relatively strong signal of the local oscillator from moving backwards into the pre-amplifier and to the receiving antenna. This characteristic is of high importance because the undesirable emission of local oscillator signals from the

![Image Frequency in the Superhet Receiver](image-url)
receiving antenna is a major problem in the design of receivers which comply with EMC regulations.

In practice, the local oscillator signals are not sinusoidal but present strong harmonic distortions. This results in additional image frequencies of higher order on both sides of the harmonics of the local oscillator frequency which are also converted to the pass band region of the IF filter:

\[ f_{RF,im(n)} = nf_{LO} \pm f_{IF} \]

These image frequencies and the corresponding harmonics of the local oscillator frequency must also be suppressed by the RF filter. The RF filter must therefore provide a high stopband attenuation even above the range of reception. LC filters or filters with dielectric resonators are used in practical applications where two to four resonant circuits are typical. These filters are called 2-, 3- or 4-pole filters. The number of poles refers to the equivalent lowpass filter and is thus equal to the number of resonant circuits\(^1\).

With an increasing receive frequency and a constant IF frequency, the relative difference between the receive frequency and the image frequency becomes smaller and smaller; thus causing the quality

\[ Q_{RF} \sim \frac{f_{RF}}{f_{IF}} \]

of the RF filter to increase. Where the separation of the receive and image frequencies can no longer be achieved by reasonable means, it is necessary to either increase the IF frequency in order to reduce the quality of the RF filter or to use a superhet receiver with two intermediate frequencies.

It is also possible to configure the RF filter such that the frequency \( f_{LO} - f_{IF} \) below the local oscillator frequency is used as the receive frequency \( f_{RF} \), while the corresponding image frequency \( f_{RF,im} = f_{LO} + f_{IF} \) is suppressed. In this case, the mixer M1 operates in the inverted mode as the frequency sequence is inverted due to the relation \( f_{IF} = f_{LO} - f_{RF} \); but, with \( f_{IF} = f_{RF} - f_{LO} \), the mixer operates in the noninverted mode and the frequency sequence remains the same.

In noninverted mode, the image frequency is below the receive frequency, while in inverted mode, it is above. Therefore, the inverted mode is always used in cases where the frequency range above the receive frequency has clearly weaker signals than the frequency

\(^1\) A simple resonant circuit has two poles: \( s = \pm j\omega_0 \). A filter with four resonant circuits therefore has eight poles, but is still called a 4-pole filter in practice since bandpass filters with a lowpass/bandpass transformation are calculated on the basis of an equivalent lowpass filter with half the number of poles.
range below the receive frequency; in this way it is easier to suppress the image frequency. The inverted mode must be compensated for in the modulator or by an inverted mode in the transmitter.

**Pre-Amplifiers**

A **low-noise amplifier** (LNA) is used in front of the RF filter to keep the noise figure of the receiver low (see Fig. 25.8b). Without a pre-amplifier, the noise figure is according to (4.201):

\[
F_r' = \frac{F_{RFF}}{G_{A,RFF}} F_{M1}
\]

Here, \(F_{RFF}\) is the noise figure and \(G_{A,RFF}\) is the available power gain of the RF filter and \(F_{M1}\) is the noise figure at the input of mixer M1. The latter is calculated with (4.201) from the noise figure of the mixer and the noise figures of the subsequent components. An overall impedance matching is assumed so that the noise figure of the filter corresponds to the power attenuation \(D_{RFF}\) in the pass band region, and the available power gain corresponds to the reciprocal value of the power attenuation. With the typical values \(D_{RFF} \approx 1.6\) (2 dB) and \(F_{M1} \approx 10\) (10 dB), the noise figure becomes unacceptably high: \(F_r' \approx 16\) (12 dB). Using a pre-amplifier with noise figure \(F_{LNA}\) and available power gain \(G_{A,LNA}\) the noise figure is:

\[
F_r = \frac{F_{LNA}}{G_{A,LNA}} + \frac{F_r' - 1}{G_{A,LNA}} \left( D_{RFF} F_{M1} - 1 \right)
\]

With a sufficiently high gain this value is much smaller than the noise figure without a pre-amplifier and in the limiting case of a very high gain it approaches the noise figure of the pre-amplifier.

In practice, the gain of the pre-amplifier cannot be increased without limits since at this point it is still the entire receive signal of the antenna that is amplified. This means that both the signal to be received and, under good receiving conditions, the signals of neighboring channels can reach relatively high levels which may overdrive a pre-amplifier with too high a gain. In addition, a high gain in the RF range is achievable with great effort only. Therefore, the gain is selected at a level which is high enough to reduce the noise figure of the receiver to an acceptable level. Typical values are \(F_{LNA} \approx 2\) (3 dB) and \(G_{A,LNA} \approx 10\ldots100\) (10\ldots20 dB). In the above example, these values lead to \(F_r \approx 2.15\ldots3.5\) (3.3\ldots5.4 dB) compared to \(F_r' \approx 16\) (12 dB) without pre-amplification.

**IF Filters**

Due to the mixer, the entire pass band region of the RF filter is shifted to the intermediate frequency range (see Fig. 25.11). Here, the channel with the desired receive frequency is filtered out by the IF filter. For this reason the IF filter is also known as the channel filter. It must have very steep edges since the transition region between the pass band and the cutoff band must not be wider than the region between adjacent channels. Particularly well suited are surface acoustic wave (SAW) filters which, despite extremely steep edges, have almost no group delay distortions. In contrast, the group delay distortions of LC or dielectric filters increase with rising edge steepness. Filters with ceramic resonators (ceramic filters) are used in applications that are relatively insensitive to group delay distortions, such as the
Fig. 25.11. Signal spectra in a superhet receiver with one intermediate frequency

case in AM broadcasting for example. In digital modulation modes, on the other hand, group delay distortions have to be kept as low as possible and thus the use of SAW filters is usually mandatory.

**Superhet Receiver with Two Intermediate Frequencies**

In the superhet receiver with two intermediate frequencies as shown in Fig. 25.12, the receive frequency is converted into a relatively high first intermediate frequency $f_{IF1}$, which is selected such that the separation of receive and image frequencies can occur with an RF filter of acceptable quality:

$$Q_{RF} \sim \frac{f_{RF}}{f_{IF1}}$$

Fig. 25.13 shows the signal spectra.

*IF filter 1* filters out a portion that contains the desired channel. It is not possible to filter out the desired channel alone at this point because of the necessary high quality. IF filter 1 serves as the image frequency filter for the second mixer, this means that the image frequency

$$f_{IF1,im} = f_{IF1} - 2f_{IF2}$$
must be within the cutoff band of the filter. To prevent a backwards transmission of the second local oscillator frequency

\[ f_{LO2} = f_{IF1} - f_{IF2} \]

this frequency must also be within the cutoff band; consequently, the quality of the filter is:

\[ Q_{IF1} \sim \frac{f_{IF1}}{f_{IF2}} \]

After conversion to the second intermediate frequency with the mixer M2, the desired channel is filtered out by means of IF filter 2 which acts as the channel filter.

It is possible to operate one or both mixers in inverted mode by regarding the frequencies \( f_{LO1} - f_{IF1} \) or \( f_{LO2} - f_{IF2} \) below the local oscillator frequencies as the receive frequencies. In this case, the RF filter suppresses the image frequency \( f_{RF,im} = f_{LO1} + f_{IF1} \) while IF filter 1 suppresses the image frequency \( f_{IF1,im} = f_{LO2} + f_{IF2} \). If only one of the mixers is operated in inverted mode, then the frequency sequence is inverted, due to \( f_{IF1} = f_{LO1} - f_{RF} \) or \( f_{IF2} = f_{LO2} - f_{IF1} \). This must be taken into account in the demodulator or must be compensated via an inverted mode in the transmitter. If both mixers are operated in inverted mode, the overall receiver operates in noninverted mode.

The advantage of the superhet receiver with two intermediate frequencies is that the quality for filtering out the desired channel can be distributed to two IF filters

\[ Q_{IF} \sim \frac{f_{IF1}}{B} = \frac{f_{IF1}}{f_{IF2}} \frac{f_{IF2}}{B} \sim Q_{IF1} Q_{IF2} \]

which is in contrast to the superhet receiver with one intermediate frequency where the task must be performed by one IF filter. This is required whenever the receive frequency \( f_{RF} \) is very high, meaning that a high (first) intermediate frequency \( f_{IF1} \) is required in order to limit the quality of the RF filter or if the bandwidth \( B \) of the receive signal is very low.

**Generating the Local Oscillator Frequencies**

The local oscillator frequencies required are derived from a crystal oscillator by means of a phase-locked loop (PLL) which has already been explained in the description of transmitters (see page 1244 and Fig. 25.7). In receivers with a variable receive frequency,
the frequency of the first local oscillator is varied by adapting the divider factors of the corresponding PLL.
25.2 Receivers

25.2.3 Gain Control

A variable gain amplifier (VGA) and an amplitude detector are used for gain control as shown by a simplified diagram in Fig. 25.14a. The VGA generates the voltage

\[ v_o(t) = A(V_R) v_i(t) \Rightarrow \hat{v}_o = |A(V_R)| \hat{v}_i \]  

(25.1)

with the variable gain \( A(V_R) \) and the control voltage \( V_R \). A peak value rectifier is usually used to determine the amplitude. By comparing the rectifier output with the setpoint value an integrator generates the control voltage \( V_R \) from the difference. Figure 25.14b shows the equivalent circuit for the gain control.

**Control Characteristic**

In steady state (operating point A) we have \( \hat{v}_o = \hat{v}_{\text{setpoint}} \) and \( V_R = V_{R,A} \) with:

\[ |A(V_{R,A})| = \frac{\hat{v}_{\text{setpoint}}}{\hat{v}_i} \]

For examination of the dynamic response we linearize (25.1) at the operating point:

\[ d\hat{v}_o = \left( \hat{v}_i \frac{d|A|}{dV_R} \right) dV_R + |A(V_R)| d\hat{v}_i \]

\[ = \hat{v}_{i,A} \frac{d|A|}{dV_R} \frac{dV_R}{k_R} + k_F d\hat{v}_i \]  

(25.2)
Using factors $k_R$ and $k_F$ and the Laplace transforms 

$$V_i(s) = \mathcal{L}(d\hat{v}_i), \quad V_o(s) = \mathcal{L}(d\hat{v}_o), \quad V_R(s) = \mathcal{L}(dV_R)$$

we obtain the linear gain control model as shown in Fig. 25.15 featuring the transfer function:

$$H_R(s) = \frac{V_o(s)}{V_i(s)} = k_F \frac{s T_I / k_R}{1 + s T_I / k_R} = k_F \frac{s T_R}{1 + s T_R} \quad (25.3)$$

Here, $T_I$ is the time constant of the integrator and $T_R$ the resulting time constant of the control circuit. This results in a highpass filter with gain $k_F$ and a $-3$ dB cutoff frequency of:

$$f_{-3dB} = \frac{1}{2\pi T_R} = \frac{k_R}{2\pi T_I} = \frac{\hat{v}_{i,A}}{2\pi T_I} \frac{d|A|}{dV_R} \Big|_{A} \quad (25.4)$$

Figure 25.16 shows the frequency response. Changes to the input amplitude with a frequency, that is below the cutoff frequency, are better suppressed with decreasing frequency; while changes with frequencies above the cutoff frequency are amplified with $k_F = |A(V_R,A)|$. The cutoff frequency must be less than the lower cutoff frequency of the amplitude modulation contained in the desired signal to prevent the desired signal from being invalidated.

According to (25.3) the cutoff frequency is proportional to the input amplitude $\hat{v}_i$ and to the derivative of the gain characteristic $|A(V_R)|$. In order to prevent the cutoff frequency from being dependent on the operating point, the condition

$$k_R = \frac{\hat{v}_i}{dV_R} = \frac{\hat{v}_{setpoint}}{|A(V_R)|} d|A| = \text{const.}$$

must be met; it follows:

$$\frac{d|A|}{dV_R} = \frac{k_R}{\hat{v}_{setpoint}} |A(V_R)| \Rightarrow |A(V_R)| = A_0 e^{k_R V_R} \quad (25.4)$$
Therefore, the VGA must have an exponential gain characteristic. In practice, the gain is quoted in decibel, i.e. logarithmically, thus producing a linear relationship:

\[ A(V_R) [\text{dB}] = A_0 [\text{dB}] + \frac{k_R V_R}{\text{setpoint}} \cdot 8.68 \text{ dB} \]

**Variable Gain Amplifier (VGA)**

There are several circuit designs for constructing a *variable gain amplifier* (VGA). In integrated circuits, the VGA with differential amplifiers for current distribution as shown in Fig. 25.17 is used almost exclusively. It offers a control range of approximately 60 dB with the required exponential characteristic.

The VGA cell consists of a common-emitter circuit with current feedback \((T_1, R_1)\) and a differential amplifier \((T_2, T_3)\). The quiescent current is adjusted with resistances \(R_2\) and \(R_3\), while \(R_7\) serves as the load resistance. The output current

\[ I_{C1}(t) = I_{C1,A} + i_{C1}(t) = I_{C1,A} + \frac{g_{m1}}{1 + g_{m1}R_1} v_i(t) \]

of the common-emitter circuit is distributed by the differential amplifier to the load resistance and the supply voltage; according to (4.61)

\[ I_{C3} = \frac{I_{C1}}{2} \left( 1 + \tanh \frac{V_R}{2V_T} \right) = \frac{I_{C1}}{1 + e^{-\frac{V_R}{V_T}}} \]

Taking the subsequent amplifier with gain \(A_V\) into account, the small-signal output voltage is:

\[ v_o(t) = -A_V i_{C3}(t)R_7 = -A_V \frac{i_{C3}(t)R_7}{1 + e^{-\frac{V_R}{V_T}}} = -A_V g_{m1}R_7 \frac{v_i(t)}{1 + e^{-\frac{V_R}{V_T}}} \]

\[ \text{Fig. 25.17. VGA with differential amplifiers for current distribution} \]

\[ ^2 \text{Current } I_{C1} \text{ corresponds to the quiescent current } 2I_0 \text{ of the differential amplifier.} \]
The control range is $V_R < -2V_T$. Here, the constant value of one is negligible with respect to the exponential function, and the desired exponential gain characteristic is:

$$v_o(t) \approx -AV \frac{g_m R_7}{1 + g_m R_1} e^{V_R / V_T} v_i(t) \Rightarrow A(V_R) \approx -AV \frac{g_m R_7}{1 + g_m R_1} e^{V_R / V_T} \quad (25.5)$$

Figure 25.18 shows the characteristic of the VGA in Fig. 25.17 for a signal frequency of 3 MHz. The control range covers 60 dB with a slope of 0.33 dB/mV. It is limited upward by the deviation from the exponential shape and downward by the reverse attenuation of the VGA cell. The latter depends on the parasitic capacitances and becomes worse with a frequency increase. Figure 25.19 shows the frequency response for different control voltages. Above 10 MHz the gain drops at a rate of 20 dB/decade; thus, the control range narrows accordingly. In this region, the minimum gain increases to 25 dB due to the declining reverse attenuation of the VGA cell.

The change in the current distribution also changes the DC voltage at the output of the VGA cell making the galvanic coupling with the subsequent amplifier difficult. The change can be compensated by connecting a second VGA cell with the same quiescent current ($T_4 \ldots T_6, R_4 \ldots R_6$) in parallel and inversely controlling the differential amplifier. Then we have

$$I_{R7,A} = I_{C3,A} + I_{C6,A} = I_{C1,A} = I_{C4,A}$$
and the DC voltage remains constant.

Dimensioning the control circuit according to (25.3) requires factor $k_R$ to be determined. A comparison between (25.4) and (25.5) yields:

$$k_R = \frac{\hat{v}_{\text{setpoint}}}{V_T} \quad (25.6)$$

Here, $\hat{v}_{\text{setpoint}}$ is the desired amplitude at the VGA output (see Fig. 25.14b). The time constant $T_I$ of the integrator can be calculated from $\hat{v}_{\text{setpoint}}$ and the cutoff frequency $f_{-3dB}$:

$$T_I = \frac{k_R}{2\pi f_{-3dB}} = \frac{\hat{v}_{\text{setpoint}}}{2\pi f_{-3dB} V_T} \quad (25.7)$$

**Localization of Gain Control in the Receiver**

In the direct-detection receiver of Fig. 25.8a, the gain control must be located in the RF section. This is inconvenient because the control range decreases with rising frequencies and the RF frequency is variable. In the superhet receiver with one intermediate frequency shown in Fig. 25.8b, the gain control is located in the IF section behind the IF filter. This arrangement is compulsory since, before the IF filter, the signal contains not only the desired channel but also all the adjacent channels with frequencies in the pass band region of the RF filter.

In systems with received levels that vary extremely, the high levels require an additional gain reduction of the pre-amplifier in order to prevent the subsequent components from being overloaded. The gain switchover of Fig. 25.12 serves this purpose. However, it only works well under the presumption that the high level is caused solely by the desirable channel. Thus, overdriving of the pre-amplifier by a neighboring channel can not be prevented.

From these considerations, it follows that an optimum operating range for all components is only possible if all amplifiers are made controllable by the level at their own output. This provides maximum sensitivity for the desired channel independent of the levels of adjacent channels. Such an elaborate design for the gain control is used in exceptional cases only. For most applications, a control system based on the level of the desired signal, as described here, is sufficient.

**Level Detection**

In addition to the amplitude-controlled useful signal, many systems require a measure for the received level of the useful signal. Typical examples include the VHF broadcasting system with automatic stereo/mono switchover controlled by the received level, and mobile communication in which several base stations receive a signal transmitted from a mobile unit and then the base station with the highest received level takes over the communication.

Level detection can be based on the control voltage of the gain control. If the controllable amplifier has an exponential characteristic, the control voltage $V_R$ is a logarithmic measure for the received level. In steady state, (25.4) provides:

$$\hat{v}_{\text{setpoint}} \equiv |A(V_R)| \hat{v}_i = A_0 \hat{v}_i e^{V_{\text{setpoint}}} \Rightarrow V_R = \frac{k_R V_R}{k_R} \ln \left( \frac{V_{\text{setpoint}}}{A_0 \hat{v}_i} \right)$$
Using (25.6), it follows for the VGA of Fig. 25.17:

\[ V_R = V_T \ln \left( \frac{\hat{v}_{\text{setpoint}}}{A_0 \cdot 1\,\text{V}} \right) - V_T \ln \left( \frac{\hat{v}_i}{1\,\text{V}} \right) \]

If \( \hat{v}_i \) increases by a factor of 10 (20 dB), \( V_R \) decreases by \( V_T \ln 10 \approx 60 \, \text{mV} \). Therefore, the slope of the level detection is \( -3 \, \text{mV/} \text{dB} \).

This simple level detection is confined to the exponential portion of the characteristic and depends on the temperature. Integrated receiver circuits usually provide a temperature compensated level signal positive with a slope which is called the *received signal strength indicator* (RSSI).

**Digital Gain Control**

With respect to the cutoff frequency of the gain control there are contradicting demands. On one hand, it should be as low as possible so that an amplitude modulation contained in the useful signal is not invalidated; whereas, on the other hand, it should be as high as possible so that following a channel switchover, the steady state is reached in the shortest time possible. One method of optimization is to switch over the time constant of the integrator. In normal operation a large time constant with a correspondingly low cutoff frequency is used, but in the case of large deviations, for example following a switchover to another channel, the system changes to a smaller time constant.

A more flexible and suitable solution is to use a digital gain control according to Fig. 25.20. Here, a microcontroller evaluates the *received signal strength indicator* (RSSI) of the last IF amplifier and performs a gain adaptation of the RF and IF amplifiers. Here, too, the majority of the control range has to be covered by the last IF amplifier because all other amplifiers also boost the neighboring channels. If, in addition to the desired channel, the neighboring channels have comparably high levels, then the risk of overdriving exists. Switching the three amplifiers on the input side of Fig. 25.20 is optional. In practice, usually only one amplifier is switched over.

Very often the digital gain control is performed in steps of 2...4 dB resolution in accordance with the gain graduation of the last IF amplifier. The gain is adjusted by a binary command (\( n_{\text{VGA}} \) Bit in Fig. 25.20). The change in gain is done either by a gain switchover in the individual amplifier stages or by using programmable attenuators between the stages.

![Fig. 25.20. Digital gain control](image-url)
The microcontroller can evaluate the received level by averaging relatively quickly the \textit{RSSI} signal, while at the same time taking into account the current amplifier setting. The microcontroller can thus programme all controllable amplifiers in one step with high accuracy, thus significantly reducing the transient time. Following this pre-setting, the duration of averaging is increased so that only amplitude variations with frequencies below the lower limit of the desired signal amplitude modulation are adjusted. In practice, the gain is set by a central microcontroller that controls the overall system. Therefore, it is particularly easy to adapt the performance to the given mode of operation (normal reception, channel switching, search mode, etc.).

25.2.4 Dynamic Range of a Receiver

The dynamic range of a receiver corresponds to the difference between the maximum and minimum received level. The maximum received level is determined by the maximum permissible intermodulation distortions and depends on the \textit{intercept point} of the receiver. The minimum received level follows from the minimum signal-to-noise ratio at the input of the demodulator and depends on the \textit{noise figure} of the receiver. In turn, the intercept point and noise figure of the receiver are dependent on the intercept points, the noise figures and the gain factors of the individual components. Therefore, the main task in designing a receiver is the selection of components with suitable characteristics. On one hand, the performance of the signal processing chain is limited by its weakest member, while on the other hand, components with unnecessarily high characteristics are either expensive or have a high power consumption. Thus, the selection of components must be balanced between the two extremes in order to achieve an optimum result.

In the example below, the dynamic range of the receiver shown in Fig. 25.21 is calculated. It is assumed that the receiver picks up channels in the range of 434 MHz with a bandwidth of $B = 200$ kHz and a channel spacing of $C = 250$ kHz. We use a receiver with one intermediate frequency $f_{IF} = 70$ MHz. Two identical RF amplifiers with gain $A = 12$ dB are used in the RF stage where RF amplifier 1 corresponds to the pre-amplifier in Fig. 25.8a. The RF filter for suppressing the image frequency

$$f_{RF,im} = f_{RF} - 2f_{IF} = 434 \text{ MHz} - 2 \cdot 70 \text{ MHz} = 294 \text{ MHz}$$

is arranged between the two RF amplifiers and is designed as a two-circuit bandpass filter with an attenuation of 6 dB ($A = -6$ dB). A programmable attenuator performs the gain switching to adapt the received level. The attenuator performance can be switched between 1 dB and 25 dB ($A_1 = -1$ dB, $A_2 = -25$ dB). It should be noted in this respect that the noise figures of a passive reactive filter and an attenuator correspond to the respective attenuation. A diode mixer with a conversion loss of 7 dB ($A = -7$ dB) and a noise figure of 7 dB is used as the mixer. Two identical IF amplifiers with gain $A = 25$ dB, and the IF filter arranged between them, follow in the IF stage. The IF filter is a surface acoustic wave (SAW) filter with a center frequency of 70 MHz and a bandwidth of 200 kHz. The attenuation is 24 dB ($A = -24$ dB). This is followed by a variable gain IF amplifier that provides the
Fig. 25.21. Example for calculating the dynamic range of a receiver
subsequent demodulator with a constant output level of 0 dBm \(v_{\text{eff}} = 224 \text{ mV}\). It is based on the VGA of Fig. 25.17 and has a high noise figure of 20 dB which is typical for VGA cells.

**Noise Figure of the Receiver**

To calculate the noise figure \(F_r\) of the receiver, we assume that all components are matched to the characteristic impedance and the quoted gain factors in decibel correspond to the available power gain \(G_A\); it thus follows:

\[
G_A [\text{dB}] = A [\text{dB}] \Rightarrow G_A = |A|^2
\]

The noise figure can be calculated with (4.201):

\[
F_r = F_1 + \frac{F_2 - 1}{G_{A1}} + \frac{F_3 - 1}{G_{A1}G_{A2}} + \cdots = 1 + F_{Z1} + \frac{F_{Z2}}{|A_1|^2} + \frac{F_{Z3}}{|A_1A_2|^2} + \cdots
\]

Here, \(F_Z = F - 1\) is the supplementary noise figure of the respective component. In Fig. 25.21 the noise figures of the components are quoted in decibel, and with

\[
F(Z) [\text{dB}] = 10^{F/10} - 1
\]

we obtain the supplementary noise figures stated in the upper portion of the table. Beneath the supplementary noise figures, the power gains at the input of the receiver, up to the input of the given component, are stated \((\Pi |A|^2)\). This allows us to convert the supplementary noise figures to the input of the receiver:

\[
F_{Z,i} = F_Z / \Pi |A|^2
\]

The supplementary noise figure and the noise figure of the receiver are obtained by arithmetic addition:

\[
F_{Z,r} = \Sigma F_{Z,i} \Rightarrow F_r = F_{Z,r} + 1
\]

For the receiver in Fig. 25.21, \(F_{Z,r} \approx 2\) and \(F_r \approx 3\) (4.8 dB).

After conversion to the input, the supplementary noise figures of the components indicate their individual contribution to the supplementary noise figure of the receiver. This shows which of the components must be of low-noise design to ensure that the noise figure of the receiver is markedly reduced and which of the components may have a higher noise figure without causing a noticeable increase in the noise figure of the receiver. In the receiver of Fig. 25.21, the contribution of the first RF amplifier dominates, followed by the contributions of the second RF amplifier and the RF filter. Under practical considerations the receiver appears to be well balanced since the noise figures of the RF amplifiers may only be decreased with a great effort. The first RF amplifier often requires a compromise between a low noise figure and a high intercept point. A high intercept point necessitates feedback and this causes an increase in the noise figure.

---

3 The 0 dBm level corresponds to a power of 1 mW at 50 \(\Omega\):

\[
P = \frac{v_{\text{eff}}^2}{50 \Omega} = 1 \text{ mW} \Rightarrow v_{\text{eff}} = 223.6 \text{ mV} \Rightarrow v_{\text{eff}} [\text{dBm}] = 20 \log \frac{v_{\text{eff}} [\text{V}]}{0.2236 \text{ V}}
\]
Minimum Received Level

The minimum received level $P_{i,\text{min}}$ is determined by the effective noise power $P_{n,i}$ at the receiver input and the minimum required signal-to-noise ratio $\text{SNR}_{i,\text{min}}$ for an error-free demodulation of the received signal:

$$\text{SNR}_{i,\text{min}} = \frac{P_{i,\text{min}}}{P_{n,i}} \Rightarrow P_{i,\text{min}} = \text{SNR}_{i,\text{min}} P_{n,i}$$ (25.8)

The minimum received level is also called the sensitivity, where a lower minimum received level is the same as an increased sensitivity.

The effective noise power results from the thermal noise power density $N_0$, the bandwidth $B$ and the noise figure $F_r$ of the receiver:

$$P_{n,i} = N_0 B F_r = kT B F_r T=300 \text{K} = 4.14 \cdot 10^{-21} \frac{\text{W}}{\text{Hz}} \cdot B F_r$$ (25.9)

Consequently:

$$P_{n,i} [\text{dBm}] = -174 \text{ dBm} + 10 \text{ dB} \cdot \log_2 \frac{B}{\text{Hz}} + F_r [\text{dB}]$$ (25.10)

Insertion into (25.8) yields the minimum received level:

$$P_{i,\text{min}} [\text{dBm}] = -174 \text{ dBm} + 10 \text{ dB} \cdot \log_2 \frac{B}{\text{Hz}} + F_r [\text{dB}] + \text{SNR}_{i,\text{min}} [\text{dB}]$$ (25.11)

The minimum received level essentially depends on the bandwidth. Therefore, the minimum received level of a system with a high data rate and a resulting high bandwidth is greater than that of systems with a low data rate, provided the systems use the same modulation mode (same as $\text{SNR}_{i,\text{min}}$) and receivers with the same noise figure. An increase in the data rate by a factor of 10 increases the minimum received level by 10 dB.

We assume that the receiver in Fig. 25.21 receives a QPSK-modulated signal with a maximum symbol failure rate of $10^{-6}$. According to [25.2] this requires a power efficiency of $E_b/N_0 = 13 \text{ dB}$. With the required power efficiency, the assumed data frequency $f_D = 200 \text{ kHz}$, and the bandwidth $B = 200 \text{ kHz}$, equation (24.83) provides the required signal-to-noise ratio:

$$\text{SNR}_{i,\text{min}} [\text{dB}] = \left( \frac{E_b f_D}{N_0 B} \right) [\text{dB}] = 13 \text{ dB}$$

Insertion into (25.11) with $B = 200 \text{ kHz}$ and $F_r \approx 5 \text{ dB}$ leads to the minimum received level:

$$P_{i,\text{min}} [\text{dBm}] = -174 \text{ dBm} + 53 \text{ dB} + 5 \text{ dB} + 13 \text{ dB} = -103 \text{ dBm}$$

This corresponds to an rms voltage of 1.6 $\mu$V.

Maximum Received Level

The maximum received level depends on the permissible intermodulation distortions. The dominating intermodulation of 3rd order (IM3) is as described by the intermodulation ratio

\[ \text{IM3} = \frac{P_3}{P_1^2} \]

We presume a QPSK system with a data rate $r_D = 200 \text{ kbit/s}$ and a roll-off factor $r = 1$. This results in a data frequency of $f_D = 200 \text{ kHz}$, the symbol frequency $f_S = f_D/2 = 100 \text{ kHz}$ (two bits per symbol) and the bandwidth $B = (1 + r) f_S = 200 \text{ kHz}$ (see (24.84)).
$IM3$ which is characterized by the intercept point $IP3$. These relationships are described in Sect. 4.2.3 on page 426 by way of the amplitude of sinusoidal signals. In telecommunication engineering, the levels are usually given in dBm or the corresponding rms values, but this does not influence the intermodulation ratio $IM3$. From (4.184) it follows:

$$IM3 \approx \left( \frac{\hat{v}_{I,IP3}}{\hat{v}_i} \right)^2 = \left( \frac{\hat{v}_{I,IP3}}{v_i} \right)^2$$  \hspace{1cm} (25.12)

Here, $v_{I,IP3}$ and $v_i$ are the rms values and $\hat{v}_{I,IP3} = \sqrt{2}v_{I,IP3}$ and $\hat{v}_i = \sqrt{2}v_i$ are the amplitudes of the intercept point $IP3$ and the received signal, both of which are related to the input of the receiver. In practice, the intermodulation ratio is quoted in decibel and the rms values of the intercept point and the received signals are quoted in dBm. It thus follows:

$$IM3 \ [\text{dB}] \approx 2 \left( v_{I,IP3} \ [\text{dBm}] - v_i \ [\text{dBm}] \right)$$  \hspace{1cm} (25.13)

The intercept point is determined by means of a two-tone signal; therefore, the intermodulation ratios according to (25.12) and (25.13) are also only valid for a two-tone signal. The receiver receives a very complex signal that is composed of the desired receive signal and the signals of adjacent channels. For this reason, the intermodulation ratio cannot be given; whereas, in practice, the two-tone intermodulation ratio is used as a substitute. For this purpose, the permissible nonlinearity is determined for the case of two neighboring channels with the same level, then the relevant two-tone intermodulation ratio and the intercept point are calculated from this value. We will not go into details on this and assume that the required two-tone intermodulation ratio is known.

The intercept point $v_{I,IP3}$ of the receiver is calculated from the intercept points of the components, but only the components up to the last IF filter are taken into account since, behind this filter, the signals of all adjacent channels are suppressed. Figure 25.21 shows the output intercept points of the components in dBm. The lower portion of the table contains the resulting effective values $v_{o,IP3}$, which are converted to the input by means of the related gain factors from the receiver input to the output of each component ($\Pi |A|$):

$$v_{o,IP3}^{(i)} = \frac{v_{o,IP3}}{\Pi |A|}$$

In Sect. 4.2.3 we demonstrated that the inverse square of the intercept points of 3rd order of a series connection must be added (see page 439):

$$\frac{1}{v_{I,IP3}^2} = \sum \frac{1}{v_{o,IP3}^{(i)}^2}$$

For the receiver in Fig. 25.21, this leads to $v_{I,IP3} = 0.124 \text{ V} (-5.1 \text{ dBm})$. QPSK usually requires an intermodulation ratio of $IM3 \approx 10000 \ (40 \text{ dB})$; the maximum received level is obtained with (25.13):

$$P_{r,\text{max} \ [\text{dBm}]} = v_i \ [\text{dBm}] = v_{I,IP3} \ [\text{dBm}] - \frac{IM3 \ [\text{dB}]}{2}$$

$$= -5.1 \text{ dBm} - \frac{40 \text{ dB}}{2} \approx -25 \text{ dBm}$$

This corresponds to an rms value of 12.6 mV.

The component intercept points converted to the input indicate the contribution which the components make to the intercept point of the receiver; a higher value is better than
a lower value. In Fig. 25.21, the contribution of the first RF amplifier dominates and is enhanced by squaring the values for the inverted quadratic addition. The dominance of the intercept point of the first RF amplifier is typical of receivers and an improvement at this point is difficult to achieve and only possible at the cost of the noise figure or the current consumption.

**Dynamic Range**

The maximum dynamic range of the receiver can be determined from the minimum and maximum received levels:

$$D_{\text{max}} \quad [\text{dB}] = P_{i,\text{max}} \quad [\text{dBm}] - P_{i,\text{min}} \quad [\text{dBm}]$$

(25.14)

For the receiver in Fig. 25.21 it follows:

$$D_{\text{max}} = -25 \text{ dBm} - (-103 \text{ dBm}) = 78 \text{ dB}$$

The signal levels for borderline cases are given in Fig. 25.21 in both dBm and Volt. One should note that the signal levels are related to the portion of the received useful signal. The overall levels may be significantly higher if adjacent channels with higher levels exist. Only after the last IF filter are the levels of the useful signal and the overall system equal; in this case, the signals of all adjacent channels are suppressed.

The available dynamic range depends on the signal levels in the adjacent channels and may be much lower than the maximum dynamic range. We examine the case of the receiver in Fig. 25.21 receiving an adjacent channel with a maximum level of $P_{i,\text{max}} = -25 \text{ dBm}$. In this case, there are intermodulation distortions, some of which add to the received channel and cause a noise-like interference with a level that is clearly higher than the thermal noise level. Therefore, the level of the useful signal must be above the minimum received level $P_{i,\text{min}} = -103 \text{ dBm}$ by the same factor to guarantee the required signal-to-noise ratio. This reduction in sensitivity is particularly undesirable in radio receivers and is the reason why weak stations which are located close to powerful stations cannot be received. The same problem occurs in base stations of mobile communication systems that must be capable of receiving very different signal levels from several mobile units. The mobile units themselves are less critical since they normally use the highest received level in their communication with the base station. Blocking of one mobile unit by other mobile units operating in the immediate vicinity is prevented by using a different frequency range for communication from the mobile units to the base stations (uplink) than from the base stations to the mobile units (downlink) (see Fig. 24.21). The separation of uplink and downlink ranges is achieved by a duplexer consisting of two bandpass filters. Figure 25.22 shows an example of a mobile unit for GSM900. The two ranges are separated by a frequency gap which is needed as the transition region for the bandpass filters of the duplexer. One disadvantage of this is the increase in the noise figure caused by the duplexer. The noise figure is increased by the power attenuation $D_D$ of the duplexer:

$$F_r' \quad [\text{dB}] = D_D \ [\text{dB}] + F_r \ [\text{dB}]$$

A typical value for duplexers is $D_D \approx 3 \ldots 4 \text{ dB}$. This means that the maximum dynamic range is reduced by the factor $D_D$ when a duplexer is used. On the other hand, the available
dynamic range increases significantly when the unit is operated in the vicinity of other mobile units since their comparably strong transmission signals can no longer reach the receiver.

The available dynamic range depends on the stop-band attenuation of the RF and IF filters. If, for example, the last IF filter has a stop-band attenuation of 50 dB, but the level of the adjacent channel is 50 dB higher, then the levels of the desired and the adjacent channels at the output of the filter are the same, and no reception is possible. The position of the image frequencies and the levels that occur at these frequencies, which are determined by the selected IF frequencies, also have an effect on the available dynamic range. Therefore, besides the above-mentioned considerations, a multitude of additional considerations must be taken into account when designing a receiver.

25.2.5 Receivers for Digital Modulation

Receivers for digital modulation methods have basically the same design as receivers for analog modulation but differ in terms of the demodulator. While analog demodulators process the IF signal directly, digital demodulators perform an additional frequency conversion by means of an I/Q mixer in order to provide the quadrature components $i(t)$ and $q(t)$.

The principle construction of a demodulator for digital modulation methods is illustrated in Fig. 24.69; Fig. 25.23a shows the same version with an additional gain control. The input signal is given by the IF signal $r_{IF}(t)$ of a superhet receiver with one or two intermediate frequencies (see Fig. 25.8b or Fig. 25.12) and corresponds to the carrier signal $s_C(t)$ of Fig. 24.69. The quadrature components $i(t)$ and $q(t)$ are derived with the help of an I/Q mixer and two lowpass filters and fed to the demodulator.

Compared to a receiver for analog modulation, the lowpass filters behind the I/Q mixer act like an additional filter. Therefore, in a receiver for digital modulation, the desired channel is normally not filtered out by the last IF filter but by the lowpass filters behind the I/Q mixer; this is the reason they are called channel filters in Fig. 25.23a. With respect to the filtering function, a receiver for digital modulation with one intermediate frequency already has the same characteristics as a receiver for analog modulation with two intermediate frequencies. Figure 25.24 shows the relevant signal spectra for the $i$ branch which are the same as for the $q$ branch.

However, channel filtering behind the I/Q mixer has two disadvantages:
- The gain control can be performed only after the lowpass filters since the IF signal may still contain adjacent channels with considerably higher signal levels. Gain control requires two variable gain amplifiers that amplify the mean of the absolute value

\[ |r_{CB}(t)| = \sqrt{i^2(t) + q^2(t)} \]

of the complex baseband signal \( r_{CB}(t) = i(t) + j q(t) \) to a setpoint value. An analog realization of this gain control is rather complex.

- The lowpass filters for channel filtering must have very steep edges as the frequency gap between the useful and the adjacent channels is very narrow. At the same time, the group...
delay in the useful channel must be as constant as possible since digital modulation is very sensitive to group delay distortions. It is difficult to meet these demands with analog lowpass filters.

Due to these disadvantages, a demodulator with analog inputs usually is used in combination with channel filtering and gain control in the IF range. In this case, the lowpass filters in Fig. 25.23a are only required to suppress the portions of the double IF frequency and the gain control for \( i \) and \( q \) is unnecessary.

**Receivers with Digital Channel Filters**

A version which is more suitable for practical use is obtained with digital filters as channel filters and a demodulator with digital inputs (see Fig. 25.23b). The output signals of the I/Q mixer undergo anti-aliasing filtering and are digitized by two A/D converters. The digital channel filters function as linear-phase FIR filters which prevent group delay distortions. The gain control is integrated in the demodulator and adapted to the modulation method used. Figure 25.25 shows the signal spectra for branch \( i \) which is identical to branch \( q \).

The demands on the anti-aliasing filter are comparatively low as the range \( 2f_{IF} - (B_{IF} + B)/2 \) is available for the transition from the pass-band to the stop-band according to Fig. 25.25; an LC filter of the 2nd or 3rd order is usually sufficient. In practice, an attenuated portion of the IF signal and the local oscillator signal are contained in the
output signal of the mixers which is caused by asymmetries and crosstalk. Usually the attenuation of the IF signal is high enough to prevent any interference. The level of the local oscillator signal is substantially higher and must be reduced. This can be done in two ways:

- Stop filters with a resonant frequency that is matched to the IF frequency are added to the anti-aliasing filters (see Fig. 25.26).
The sampling frequency of the A/D converter is selected such that the difference between the IF frequency and the harmonics of the sampling frequency is larger than half the bandwidth of the desired signal \((= B/2)\). After sampling, the IF frequency is then within the stop-band of the digital channel filters. A combination of both methods is also possible.

After anti-aliasing filtering, the signal has an upper cutoff frequency corresponding to half the bandwidth of the IF filter \((= B_{IF}/2)\) (see Fig. 25.25). Therefore, a sampling frequency \(f_{AD} > B_{IF}\) would be required for nonaliasing A/D conversion. Since the subsequent digital channel filter suppresses all portions above a frequency of half the bandwidth of the desired signal \((= B/2)\), aliasing may be permitted in this region. Consequently the sampling frequency is:

\[
 f_{AD} > \frac{B_{IF} + B}{2} \tag{25.15}
\]

Figure 25.25 is the borderline case of the minimum sampling frequency. The aliasing components shown as a broken line go right up to the desired channel.

The IF signal and the signals behind the mixers still contain several adjacent channels; therefore, the overall level of these signals may be significantly higher than the level of the desired channel. In order to prevent overdriving of the A/D converters, it is necessary to use a gain control for the IF signal in addition to the gain control for the desired channel that is integrated in the demodulator. For this purpose, the gain control in the IF range, which exists in the superhet receivers shown in Figs. 25.8b and 25.12, is used.

**Dynamic range:** The available dynamic range of the receiver depends primarily on the resolution of the A/D converter. We demonstrate this for a desired channel of power \(P_C\) and an adjacent channel of power \(P_{AC}\). The corresponding spectrum at the output of an A/D converter is plotted in Fig. 25.27. The powers of the channels correspond to the areas below the respective curves\(^5\). \(P_{n,Q}\) is the power of the quantization noise of the A/D converter and is evenly distributed in the frequency interval from zero up to half the sampling frequency. We assume that the power in the neighboring channel is significantly higher than the power

\[^5\text{The power of a signal } \phi(t) \text{ with the Fourier transform (two-sided spectrum) } \Phi(f) \text{ is:}

\[P_{\phi} = \int_{-\infty}^{\infty} \left| \Phi(f) \right|^2 df\]

This is called the Parseval equation. We use unilateral absolute spectra; this eliminates the negative frequencies, and the lower limit of the integral becomes zero.\]
Fig. 25.27. Spectrum at the output of the A/D converter for a desired channel with the power $P_C$ and an adjacent channel with the power $P_{AC}$. $P_{n,Q}$ is the power of the quantization noise. $P_{n,C}$ is the portion in the desired channel.

in the desired channel. Thus, the overall power is approximately equal to the power in the adjacent channel:

$$P = P_C + P_{AC} + P_{n,Q} \approx P_{AC}$$

When fully modulated, an ideal A/D converter with a resolution of $N$ bits achieves the signal-to-noise ratio:

$$SNR = \frac{3 \cdot 2^{2N}}{C^2} \quad \Rightarrow \quad SNR [\text{dB}] = N \cdot 6 \text{ dB} + 4.8 \text{ dB} - CF [\text{dB}] \quad (25.16)$$

The crest factor of the signal is

$$CF = \frac{\text{Peak value}}{\text{rms value}} = \frac{v_{\text{max}}}{v_{\text{eff}}} \quad (25.17)$$

The crest factor ranges from $CF = 1$ (0 dB) for a square-wave signal to $CF \approx 4$ (12 dB) for a noise-like signal. This means that the achievable signal-to-noise ratio depends on the signal in the adjacent channel. The power of the quantization noise can be calculated from the overall power $P$ and the signal-to-noise ratio:

$$SNR = \frac{P}{P_{n,Q}} \quad \Rightarrow \quad P_{n,Q} = \frac{P}{SNR} = \frac{PCF^2}{3 \cdot 2^{2N}}$$

The portion is within the desired channel (see Fig. 25.27):

$$P_{n,C} = P_{n,Q} \cdot \frac{B}{f_{AD}} = \frac{PCF^2}{3 \cdot 2^{2N}} \frac{B}{f_{AD}}$$

In order to ensure correct demodulation of the desired signal, the signal-to-noise ratio $SNR_C$ in the desired channel must be higher than the minimum signal-to-noise ratio $SNR_{i\text{,min}}$ of the modulation method used:

$$SNR_C = \frac{PC}{P_{n,C}} > SNR_{i\text{,min}}$$

---

6 For a sinusoidal signal with $CF = \sqrt{2}$ (3 dB) (25.16) produces the relationship $SNR = N \cdot 6 \text{ dB} + 1.8 \text{ dB}$ (see (18.9) on page 950).
Thus, the power in the desired channel is

\[ P_C > \frac{SNR_{i,\text{min}} PC^2}{3 \cdot 2^N} \cdot \frac{B}{f_{AD}} \]  

(25.18)

and the permitted ratio of the powers in the adjacent channel and the desired channel (available dynamic range) is:

\[ \frac{P_{AC}}{P_C} \approx \frac{P}{P_C} < \frac{3 \cdot 2^N}{SNR_{i,\text{min}} CF^2} \cdot \frac{f_{AD}}{B} \]  

(25.19)

The parameters \( SNR_{i,\text{min}}, CF \) and \( B \) are defined by the modulation method used. Therefore, the available dynamic range is essentially determined by the resolution \( N \) of the A/D converter and the sampling frequency \( f_{AD} \). While the sampling rate is often increased in audio applications to achieve a better signal-to-noise ratio (oversampling), this is usually not possible in receivers due to their very high minimum sampling rate; here, it is the resolution that must be increased if the available dynamic range is not large enough.

Owing to a number of interferences, the signal-to-noise ratio of a real A/D converter is lower than that of an ideal A/D converter according to (25.16). In practice, one must use the effective resolution \( N_{eff} < N \), which is quoted in data sheets, in place of the resolution \( N \). Instead of the effective resolution, many data sheets specify the signal-to-noise ratio for a sinusoidal signal as a function of the signal and the sampling frequencies; in this case, the effective resolution is

\[ N_{eff} = \frac{SNR [\text{dB}]}{6 \text{dB}} - 1.8 \text{dB} \]  

(25.20)

Example: We consider a receiver for a QPSK system with a data rate \( r_D = 200 \text{ kbit/s} \), a roll-off factor \( r = 1 \) and a bandwidth of \( B = 200 \text{ kHz} \). The bandwidth of the last IF filter is assumed to be \( B_{IF} = 1 \text{ MHz} \). According to (25.15) the sampling frequency is

\[ f_{AD} > \frac{B_{IF} + B}{2} = 600 \text{ kHz} \]

We choose \( f_{AD} = 800 \text{ kHz} \). At a bit error rate of \( 10^{-6} \) QPSK requires a minimum signal-to-noise ratio \( SNR_{i,\text{min}} = 20 \text{ (13 dB)} \); with \( r = 1 \) the crest factor is \( CF \approx 1.25 \text{ (2 dB)} \). Furthermore, we assume an available dynamic range of \( P_{AC}/P_C = 10^6 \text{ (60 dB)} \). Solving (25.19) for \( N \) we obtain:

\[ N > \frac{1}{2} \log \left( \frac{P_{AC}}{P_C} \cdot \frac{SNR_{i,\text{min}} CF^2}{3} \cdot \frac{B}{f_{AD}} \right) = \frac{1}{2} \log \left( \frac{10^6 \cdot 10.4 \cdot 1}{4} \right) \approx 10.7 \]

Thus, an A/D converter with an effective resolution of at least 10.7 bits at \( f_{AD} = 800 \text{ kHz} \) is required. If operated with a sinusoidal signal, the signal-to-noise ratio according to (25.20) is \( SNR = 10.7 \cdot 6 \text{ dB} + 1.8 \text{ dB} = 66 \text{ dB} \). In practice, this means that a 12 bit converter is required.

This example is typical of receivers with digital channel filters. A/D converters with a comparably high resolution are required despite the fact that the signal-to-noise ratio \( SNR_{i,\text{min}} \) required in the desired channel is very low. This is necessary due to the high signal levels in adjacent channels.
Receivers with IF Sampling and Digital Channel Filters

Using digital I/Q mixers in addition to digital channel filters results in a receiver with IF sampling in which the IF signal is already digitized as shown in Fig. 25.23c. As the bandwidth $B_{IF}$ of the IF signal is usually significantly lower than the IF frequency, sub-sampling is possible, which means that the sampling frequency $f_{AD}$ is selected lower than the IF frequency without compromising the demand that $f_{AD} > 2B_{IF}$. The aliasing effect converts the IF signal to a lower frequency. Figure 25.28 illustrates this with sampling in the first, second and third aliasing region as compared to sampling in the main region.

Sampling in the main region means that it is necessary to comply with the sampling theorem in its standard form, i.e. the upper cutoff frequency must be lower than half the

---

Fig. 25.28. Frequency conversion in IF sampling
sampling frequency:

\[ f_g = f_{IF} + \frac{B_{IF}}{2} < f_{AD} \]

For subsampling in the \( m \)th aliasing region, the IF signal must be fully within this frequency range. Thus, at the lower limit this requires

\[ f_{IF} - \frac{B_{IF}}{2} > m \frac{f_{AD}}{2} \]

and at the upper limit:

\[ f_{IF} + \frac{B_{IF}}{2} < (m + 1) \frac{f_{AD}}{2} \]

In summary, the general condition for the sampling frequency \( f_{AD} \) is:

\[
\frac{2f_{IF} + B_{IF}}{m + 1} < f_{AD} < \frac{2f_{IF} - B_{IF}}{m}
\]

with \( m \leq \frac{f_{IF}}{B_{IF}} - \frac{1}{2} \) (25.21)

For \( m = 0 \) this also applies to the main region but in this case the upper limit is excluded. By inserting the maximum possible integer for \( m \) into (25.21), the minimum sampling frequency \( f_{AD,\text{min}} \) is obtained. This minimum value depends on the ratio \( f_{IF}/B_{IF} \) and lies within the range:

\[ 2B_{IF} < f_{AD,\text{min}} < 2B_{IF}\left(1 + \frac{B_{IF}}{2f_{IF}}\right) \]

The digital IF frequency \( f_{IF,D} \) at the output of the A/D converter is:

\[
f_{IF,D} = \begin{cases} 
  f_{IF} - m \frac{f_{AD}}{2} & \text{if } m \text{ is even} \\
  (m + 1) \frac{f_{AD}}{2} - f_{IF} & \text{if } m \text{ is odd}
\end{cases}
\]

It follows that with even \( m \) values, the IF signal is converted in noninverted mode and with odd \( m \) values in inverted mode (see Fig. 25.28). The inverted mode must either be taken into account in the demodulator or compensated by an inverted mode in the transmitter or in the mixers of the previous superhet receiver.

The I/Q mixer generates the signals from the digital output signal \( r_{AD}(n) \) of the A/D converter:

\[
i_M(n) = r_{AD}(n) \cos\left(2\pi n \frac{f_{IF,D}}{f_{AD}}\right)
\]

\[
q_M(n) = -r_{AD}(n) \sin\left(2\pi n \frac{f_{IF,D}}{f_{AD}}\right)
\]

The digital quadrature components \( i(n) \) and \( q(n) \) are obtained after channel filtering. The digital I/Q mixer becomes particularly simple if the digital IF frequency is equal to a

\[7\) This condition applies only to cases where the entire IF signal is to be processed digitally. If limited to the desired channel, aliasing can be allowed as long as the desired channel is not affected. This will be detailed further below.\]
Fig. 25.29. Digital receiver with IF sampling where $f_{IF,D} = f_{AD}/4$. The switches operate in synchronization with the A/D converter.

fraction of the sampling frequency; then we have

$$f_{IF,D} = \frac{f_{AD}}{4} \Rightarrow \begin{cases} i_M(n) = r_{AD}(n) \cos \left( \frac{\pi n}{2} \right) \\ q_M(n) = -r_{AD}(n) \sin \left( \frac{\pi n}{2} \right) \end{cases}$$

(25.23)

with:

$$\cos \left( \frac{\pi n}{2} \right) = 1, 0, -1, 0, \ldots \quad \text{for } n = 0, 1, 2, 3, \ldots$$

$$\sin \left( \frac{\pi n}{2} \right) = 0, 1, 0, -1, \ldots \quad \text{for } n = 0, 1, 2, 3, \ldots$$

In this case, only the factors 0 (value is suppressed), 1 (value is taken over) and $-1$ (value is taken over with the sign inverted) occur and no multiplication is required. Equation (25.23) specifies the relationship:

$$i_M(n) = [ r_{AD}(0), 0, -r_{AD}(2), 0, r_{AD}(4), 0, \ldots ]$$

$$q_M(n) = [ 0, -r_{AD}(1), 0, r_{AD}(3), 0, -r_{AD}(5), \ldots ]$$

Accordingly, the signal sequence $r_{AD}(n)$ must pass a controlled inverter and then be distributed to the two outputs by a demultiplexer resulting in a digital receiver with IF sampling as shown in Fig. 25.29.

The condition for the sampling frequency is obtained by inserting (25.23) into (25.22):

$$f_{AD} = \frac{4f_{IF}}{2m+1} \quad \text{with } m \leq \frac{f_{IF}}{B_{IF}} - \frac{1}{2}$$

(25.24)

It follows that $f_{AD} = 4f_{IF}$ in the main region ($m = 0$, noninverted mode), $f_{AD} = 4f_{IF}/3$ in the first aliasing region ($m = 1$, inverted mode), and $f_{AD} = 4f_{IF}/5$ in the second aliasing region ($m = 2$, noninverted mode), etc. This condition is met in Fig. 25.28. Figure 25.30 lists some of the commonly used IF frequencies together with the corresponding sampling frequencies for $m = 0 \ldots 4$.

For subsampling, it is necessary to use special A/D converters suitable for subsampling because the analog bandwidth (i.e. the bandwidth of the analog input circuitry and the sample-and-hold element) must be higher than the sampling frequency.
<table>
<thead>
<tr>
<th>IF-frequencies</th>
<th>Sampling frequencies</th>
</tr>
</thead>
<tbody>
<tr>
<td>455 kHz</td>
<td>1.82 MHz  606.67 kHz</td>
</tr>
<tr>
<td>10.7 MHz</td>
<td>42.8 MHz  14.267 MHz</td>
</tr>
<tr>
<td>21.4 MHz</td>
<td>85.6 MHz  28.533 MHz</td>
</tr>
<tr>
<td>70 MHz</td>
<td>280 MHz  93.33 MHz</td>
</tr>
</tbody>
</table>

Table 25.30. Sampling frequencies for some commonly used IF frequencies

Figure 25.31 shows the signal spectra of a digital receiver with IF sampling for \( f_{IF,D} = f_{AD}/4 \) and \( f_{AD} = 4f_{IF}/5 \) \((m = 2)\). We can see that no aliasing occurs when the condition of (25.24) is met; this means that the entire IF signal is digitized without any loss. Thus, it is possible to receive adjacent channels by using bandpass filters, instead of lowpass filters, as channel filters and by once again converting the frequency of the output signal. This enables
the reception of all channels that are completely within the pass band of the IF filter without changing the local oscillator frequency. Switching the channel filters is particularly easy in practice as the channel filtering is generally done by a digital signal processor (DSP); only the coefficients for the filter must be exchanged. This method is of particular importance in narrow-band systems since it enables an entire group of channels to be received with the same local oscillator frequency. In extreme cases the entire frequency band of the application is within the IF bandwidth; here, one can use a fixed local oscillator frequency and perform the channel selection solely by switching the channel filters. If, however, only the desired channel is to be processed, as in Fig. 25.31, aliasing can be permitted as long as the desired channel is not affected; thus the condition for \( m \) in (25.24) can then be widened. We illustrate this by enlarging the IF bandwidth in Fig. 25.31 to the point just before aliasing in the desired channel occurs (see Fig. 25.32). It thus follows:

\[
B_{IF,\text{max}} = f_{AD} - B \quad \Rightarrow \quad f_{AD} > B_{IF} + B
\]  

(25.25)

Inserting (25.24) into (25.25) and solving for \( m \) yields the condition:

\[
m < \frac{2f_{IF}}{B_{IF} + B} - \frac{1}{2}
\]  

(25.26)

A comparison of (25.25) and (25.15) shows that with IF sampling the minimum sampling frequency is twice as high than for sampling of the quadrature components after analog I/Q mixing. The reason for this is that the IF signal contains both quadrature components:

\[
r_{IF}(t) = i(t) \cos(2\pi f_{IF}t) - q(t) \sin(2\pi f_{IF}t)
\]

This shows that it is possible to perform IF sampling with one A/D converter and a sampling rate according to (25.25) or to perform sampling of the quadrature components with two A/D converters and half the sampling rate.

**Comparison of Receivers with Digital Modulation Methods**

The receiver with analog channel filters shown in Fig. 25.23a is not used in practice. Only the version with channel filtering and gain control in the IF range is of importance and the analog lowpass filters are only used for suppressing the components with twice the IF frequency. This version is often found in uncomplicated systems with simple modulation methods and comparably low data rates.

The receiver with digital channel filters is a widespread system. It allows a much better separation from adjacent channels thus allowing very narrow frequency gaps between the channels so that better use is made of the frequency band available for the application. Sampling of the quadrature components is possible with A/D converters with lower analog bandwidth thus keeping the power loss in the analog section of the converter low. As
the modulation method becomes more complex, the disturbing effects of the unavoidable asymmetries in the analog I/Q mixer increase which in turn increases the bit error rate. Thorough tuning of the I/Q mixer, in regards to the amplitude and phase of the two signal paths, is essential if complex modulation methods are used. This adjustment must have a high degree of stability in terms of temperature sensitivity and durability in order to permanently satisfy the demands.

The digital I/Q mixer in the receiver with IF sampling works accurately. Thus, this receiver yields the best results. If the condition $f_{IF, D} = f_{AD}/4$ is met, the mixer only consists of three multiplexers and one inverter.

**Direct Conversion Receiver**

If, instead of the IF signal, the RF signal is used as the input signal for receivers with digital modulation as shown in Fig. 25.23 on page 1266, then the receiver becomes a direct conversion receiver. The previous superhet receiver is reduced to the pre-amplifier and the RF filter and all IF components are eliminated. In practice, the receiver with digital channel filters, according to Fig. 25.23b, is used almost exclusively. For optimum utilization of the A/D converter, a gain control is necessary after the I/Q mixer. The gain control for the desired channel takes place as usual in the demodulator. This gives us the direct conversion receiver, a typical version of which is shown in Fig. 25.33. Figure 25.34 shows the related signal spectra for branch $i$, which also apply for branch $q$.
There are no image frequencies in the direct conversion receiver. Therefore, the RF filter is only required to limit the received band with the aim of limiting the received power. In the superhet receiver, the bandwidth of the RF filter must be at least as large as the frequency range to be received; it may be even larger as long as the additional received power does not restrict the dynamic range of the subsequent components too much.
In addition to the portions with differential frequencies in the range of \( 0 \leq f \leq B_{RF}/2 \), the output signals of the I/Q mixer also contain portions of the sum frequencies in the range of \( 2f_{RF} \). Furthermore, there are portions at the frequency \( f_{RF} \) generated by crosstalk in the mixers that are suppressed by the anti-aliasing filter.

The minimum sampling frequency of the A/D converter depends on the bandwidth \( B \) of the desired channel and the bandwidth \( B_{AAF} \) of the anti-aliasing filter or \( B_{RF} \) of the RF filter depending on which of the two bandwidths is lower:

\[
f_{AD} > \begin{cases} 
\frac{B + B_{AAF}}{2} & \text{for } B_{AAF} < B_{RF} \\
\frac{B + B_{RF}}{2} & \text{for } B_{AAF} \geq B_{RF}
\end{cases}
\] (25.27)

In both cases, the desired signal remains free of any aliasing contents. The situation of \( B_{AAF} < B_{RF} \) is shown in Fig. 25.34. However, the sampling frequency can also be selected such that all channels are digitized in the pass band region of the RF filter without aliasing, and the channel selection can be achieved by switching the digital channel filters where, in this case, the following must apply: \( f_{AD} > B_{RF} \). The anti-aliasing filter is then used exclusively for suppressing the signal portions in the frequency range \( f_{RF} \) and \( 2f_{RF} \).

The primary advantage of a direct conversion receiver is its reduced number of filters. It is particularly suitable for monolithic integration as only the RF filter is required as an external component and RC filters are used as the anti-aliasing filters. At the same time, only one local oscillator with an RC quadrature network \( (0^\circ/90^\circ) \) is required, which may also be integrated with the exception of a resonant circuit that determines the frequency, and a variable capacitance diode for frequency tuning. The elimination of the IF components substantially reduces the current consumption of the receiver. In particular, power-consuming drivers required for the SAW-IF filters in the superhet receiver and the subsequent amplifiers for compensating the relatively high attenuation of these filters are eliminated.

Besides the advantages mentioned, the direct conversion amplifier also has three problems with negative effects that must be reduced to a noncritical level by additional circuitry:

- The local oscillator frequency corresponds to the received frequency which causes the risk of the relatively strong local oscillator signal reaching the antenna via the RF filter and the pre-amplifier and then being transmitted (see Fig. 25.35). In order to avoid this, the pre-amplifier must have a particularly low reverse transmission. As an alternative, a 3-gate circulator can be introduced between the pre-amplifier and the RF filter. This eliminates the local oscillator signal at the third gate so that it no longer reaches the output of the pre-amplifier (see Fig. 25.36).
- If the local oscillator signal reaches the RF path and is reflected, the result is a self-mixing effect. This results in a DC component at the outputs of the I/Q mixer which
superimposes the DC component of the desired signal. Since the removal of this disturbing DC component is not possible, the total DC component must be removed in the demodulator by a digital highpass filter with a very low cutoff frequency. This must be done in such a way that the useful signal is affected as little as possible.

- The variable gain amplifiers operate as LF amplifiers in the frequency of 1/f noise, which increases the noise figure significantly above that of an IF amplifier. The influence from the noise figure of the receiver can, of course, be reduced by making the gain of the RF pre-amplifier as high as possible, but this is limited by the fact that a high gain in the RF range is only possible with several amplifier stages and comparably high power consumption; at the same time this reduces the overload margin. Figure 25.37 shows the plot of the spectral noise figure $F(f)$ of the variable gain amplifier over the channel sequence. One possible way of improving the noise figure is to reduce the spectral noise figure by using the $m$th adjacent channel at $f = mC$ as the desired channel instead of the channel at $f = 0$.

---

8 According to Fig. 25.37 the bandwidth of the adjacent channels is twice as high as the bandwidth of the channel at $f = 0$. However, these channels contain two RF channels, i.e. $f_{RF} + C$ and $f_{RF} - C$, which are separated by combining the quadrature components in the following digital processing; only half of the noise power becomes effective so that factor 2 in the bandwidth is compensated.
Maintaining the required amplitude and phase position accuracy of the I/Q mixer poses an additional problem. The demands on I/Q mixers in direct conversion receivers and in superhet receivers are identical, but meeting them is much more difficult in the I/Q mixer with RF input than in the I/Q mixer with IF input due to the higher frequency.

Today, the problems outlined for direct conversion receivers are dealt with well. Therefore, it may be assumed that direct conversion receivers will replace superhet receivers. In this respect, considerations arise as to whether the receiver with IF sampling, according to Fig. 25.23c, shall be also used as a direct conversion receiver by placing only an additional pre-amplifier and an RF filter in front of the A/D converter. This is known as RF sampling.
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