# 2

# DSP Based Radio Receiver Architectures

The incorporation of digital signal processing into the receiver has made the implementation of fully digital demodulation schemes possible. Such schemes are an effective way to improve the reception quality in legacy analog broadcasting systems.

In a scenario where the analog signal processing is to be replaced by digital functions, the ADC is a key part of a modern receiver. Several DSP based receiver architectures are described in section 2.1. The ADC performance, i.e. dynamic range, resolution, linearity, etc. determines the location of the ADC in the receiver path. These performance metrics are defined in section 2.2 and the concepts of dynamic-range extension, desensitization and blocking are discussed in section 2.3. Section 2.4 reviews the issue of image rejection in low-IF and zero-IF receivers, with emphasis on the improvements achieved by quadrature architectures. Section 2.5 presents the technical specifications of analog radio broadcasting and several examples of integrated architectures for AM/FM receivers. The technical specifications and the most important features of the IBOC digital radio broadcasting standard are presented in section 2.6. This chapter ends in section 2.7 with a description of several commercial chip-sets for AM/FM and IBOC radios.

## 2.1 Radio Receiver Architectures

From the early days of the AM radio broadcasting to the widespread use of mobile phones, the evolution of the wireless systems has followed the development of electronic technology. While both transmitters and receivers have profited from this continuous development, the latter have undergone the most revolutionary changes due to demands for increased portability and reduced power consumption. The replacement of vacuum tubes by transistors after World War II was responsible for the first radical changes leading to cost reduction and miniaturization. Further miniaturization was achieved with increased integration and digitization of the receiver back-end. However, even today, external passive components are necessary in the radio front-end. In a modern receiver, the amounts of analog and digital signal processing, and the level of integration are dictated by the characteristics of different communication systems. Each case requires the hardware to be developed with a different emphasis on cost reduction, performance (bandwidth, dynamic range, linearity, etc.), portability and programmability.

### 2.1.1 Heterodyne and Homodyne Receivers

Figure 2-1 portrays an example of the RF input spectrum received by the antenna. This spectrum is composed of the entire receiver band plus some interferers. The purpose of the receiver is to retrieve the information within a desired channel with a minimum pre-defined quality level. In order to accomplish this task, several steps of amplification, filtering and down-conversion may be used [1].

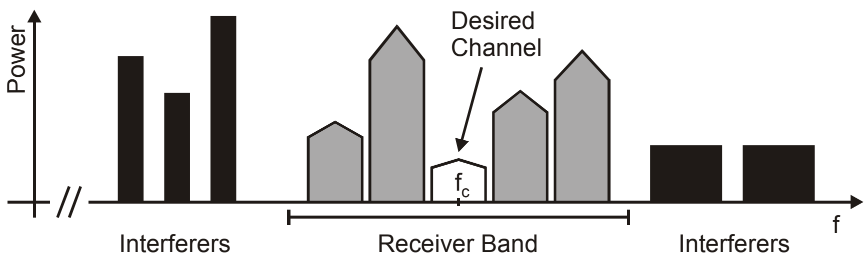


Figure 2-1: *Receiver RF input spectrum.*

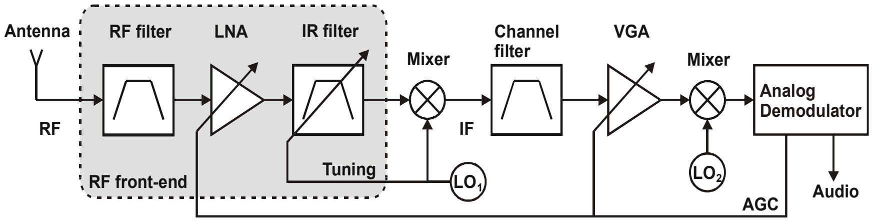


Figure 2-2: *Fully analog heterodyne receiver [1].*

Figure 2-2 shows an example of a fully analog heterodyne receiver. Because the receiver band may be surrounded by strong interferers, a first RF bandpass filtering step is required after the antenna. The received signal may be very weak and needs to be amplified by the low noise amplifier (LNA) prior to further signal conditioning. To filter the desired channel directly from the RF, very high quality-factor filters would be needed. In order to relax the quality-factor requirements for channel selection, a down-conversion mixer is used to translate the receiver band to a lower intermediate frequency (IF). The tuning of the receiver to select the desired channel is performed by changing the frequency of the first local oscillator ($LO_1$). The mixer stage however, requires an image rejection filter to attenuate interferers located at the image frequencies. At IF frequencies, the desired channel can be selected and all other channels are strongly attenuated by the channel filter. The desired channel is then amplified again by the variable gain amplifier (VGA) and then down-converted to a lower frequency. The final processing step is analog demodulation, where the baseband analog information is retrieved. In order to increase the receiver's dynamic range, both the LNA and VGA gains are controlled by an automatic gain control (AGC) loop [1].

When the $LO_1$ is chosen to be equal to the center frequency ($f_c$) of the desired channel, this channel is directly down-converted to dc and no other mixer stages are needed. This receiver architecture is known as homodyne or zero-IF, and is depicted in Figure 2-3. Because the IF frequency is dc, the image band contains a replica of the desired channel. The homodyne architecture is sensitive to low-frequency errors, such as dc offset (from amplifiers or caused by self-mixing) and *1/f* noise, which fall inside the desired channel [1]. The issue of zero-IF mixing and image rejection is discussed in more detail in section 2.4.

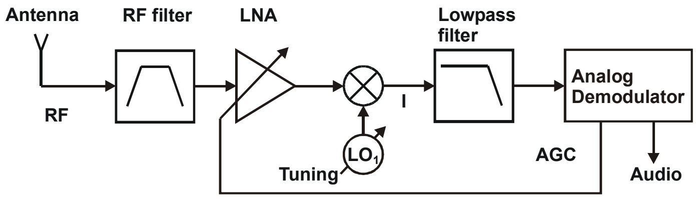


Figure 2-3: *Fully analog homodyne receiver [1].*

### 2.1.2 DSP Based Radio Receiver Architectures

The continuous development of CMOS technology and the increasing availability of digital processing power has enabled a series of mixed analog/digital radio receivers. With the first generation of DSP based receivers, the only analog operations to be implemented in software were audio baseband functions such as stereo decoding and audio control. These digital features can be added to the receiver architectures shown in Figures 2-2 and 2-3 if an ADC and a DSP [8]–[10] replace the analog audio processing circuitry. Several high quality-factor filters are typically still necessary for the full analog channel selection [11], which limits the level of integration.

The next generation of DSP based receivers (Figure 2-4) incorporated digital demodulation to replace analog demodulation techniques [12]. This digital processing greatly improved the quality of the reception using several new features that could not be easily implemented in fully analog radios, such as adjacent channel suppression and multi-path interference detection/suppression. The performance required from an ADC in a receiver architecture depends on the amount of analog signal processing operations that precede the digitization. In the architecture shown in Figure 2-4, several stages of filtering, amplification and down-conversion relax the ADC's speed and resolution requirements. In order to increase the level of integration and reduce costs, some of the external channel selection filters can be replaced by integrated active filters. Because of the limited quality-factor, quadrature down-conversion is normally used to relax image rejection requirements (see section 2.4) of the integrated active filters. However, two independent in-phase and quadrature-phase (I/Q) paths and two ADCs are required by this architecture.

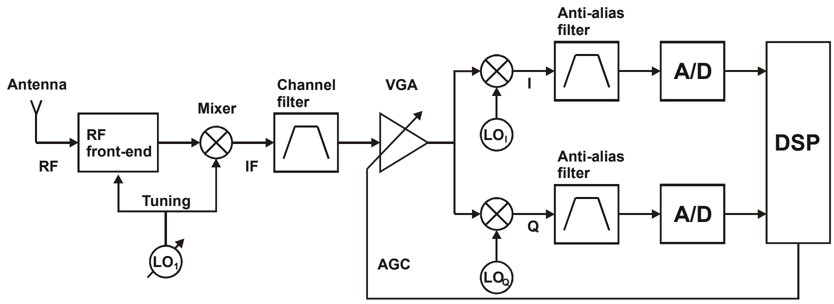


Figure 2-4: *Heterodyne quadrature receiver with baseband ADCs and digital demodulation [12].*

Another way to further reduce the number of external analog components is to move the ADC to the IF level, as shown in Figure 2-5. In this architecture, a selected narrow-band channel is digitized at IF and the quadrature mixing is performed in the digital domain with almost perfect linearity and I/Q matching [13]–[17]. Because the signal at IF is not complex, just one ADC is required. However, the direct IF digitization requires a faster ADC than the previous architecture (Figure 2-4). A high quality-factor channel filter and an AGC loop relax the DR and bandwidth requirements of the ADC. The IF ADC in this architecture is normally an oversampled bandpass ΣΔ ADC [18]. Bandpass ΣΔ ADCs are discussed in more detail in section 4.4.

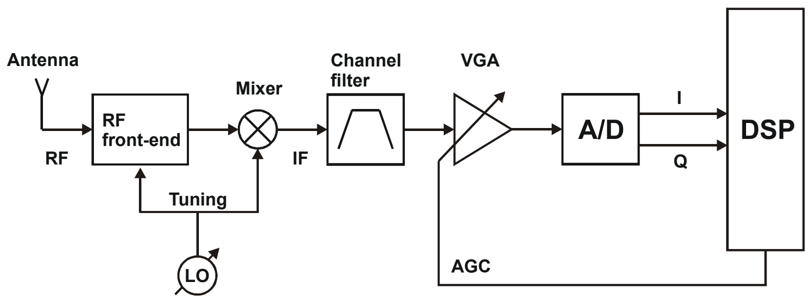


Figure 2-5: *Heterodyne receiver with bandpass IF ADC and digital demodulation [13]–[17].*

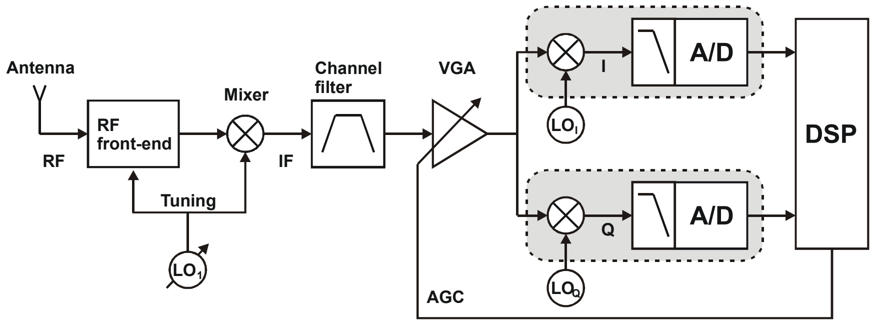


Figure 2-6: *Heterodyne receiver with IF-to-baseband ADCs and digital demodulation [19]–[21].*

An alternative approach to IF digitization was achieved by the combination of integrated down-conversion mixers and baseband ADCs. The receiver architecture shown in Figure 2-6 combines the decreased number of external filters offered by the IF digitization architecture with more linear and power efficient IF-to-baseband ADCs [19]. The drawback of this architecture is that the analog quadrature down-mixing is prone to I/Q mismatch and limits the image suppression (section 2.4). Figure 2-7 shows a receiver with higher DR IF-to-baseband ADCs [23] that do not need to be preceded by a VGA. The removal of the AGC loop prevents strong interferers from desensitizing the receiver when the desired channel is very weak (section 2.3).

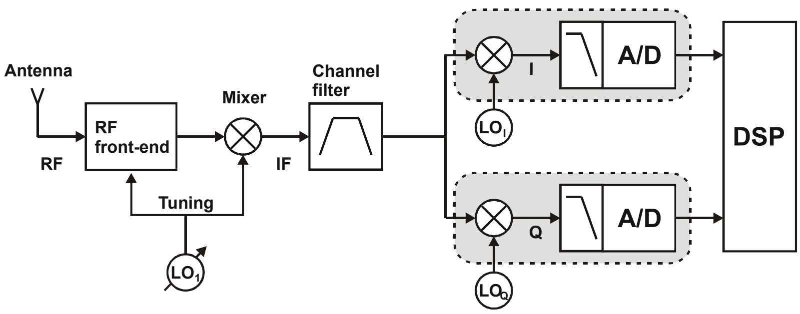


Figure 2-7: *Heterodyne receiver with higher dynamic range IF-to-baseband ADCs and digital demod. [23].*

## 2.2 ADC Performance Metrics

The most important performance metrics for ADCs used in telecommunication applications are reviewed in this section [24]. They are divided in two groups: resolution definitions and linearity definitions.

### 2.2.1 Measures of Resolution

The resolution of an ADC expresses the minimum detectable change of the analog input related to the maximum input. The most important measures of resolution are dynamic range, signal-to-noise ratio and effective number of bits. Figure 2-8 shows the signal-to-noise ratio plotted as a function of the power of a sinusoidal input, and the relation between several measures of resolution.

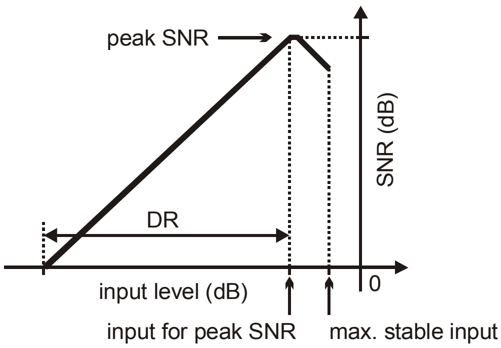


Figure 2-8: *Definition of dynamic range (DR).*

***Dynamic range*** **(DR)** – ratio between maximum input power and minimum detectable input power within a certain bandwidth. Generally the minimum input is determined by the noise power in the bandwidth of interest.

***Signal-to-noise ratio*** **(SNR)** – ratio between the input power (normally a sinusoidal signal) and the noise power inside a certain bandwidth.

***Peak*** **SNR** – peak value of the SNR plot. The ADC resolution is very often expressed by the peak SNR in number of bits.

***Effective number of bits* (ENOB)** – the peak SNR at the ADC output expressed as a number of bits:

$$ENOB = \frac{SNR_{peak}(dB) - 1.76}{6.02} \tag{2-1}$$

### 2.2.2 Measures of Linearity

ADC linearity is a very important performance specification in DSP based receivers. Receiver architectures with multi-band digitization are especially sensitive to distortion components induced by strong unwanted channels inside the bandwidth of a weak wanted channel. Ideally, non-linearity induced spectral components should be below the minimum detectable input signal. The most important linearity definitions are harmonic distortion, intermodulation distortion, spurious-free dynamic range, signal-to-noise-and-distortion ratio (SNDR), peak SNDR, intermodulation intercept point and cross-modulation distortion.

***Harmonic distortion* ($HD_x$)** – ratio between maximum input sinusoidal power and the power of the $x^{th}$ harmonic of the input tone. The second ($HD_2$) and the third ($HD_3$) harmonic components are normally the most important (Figure 2-9).

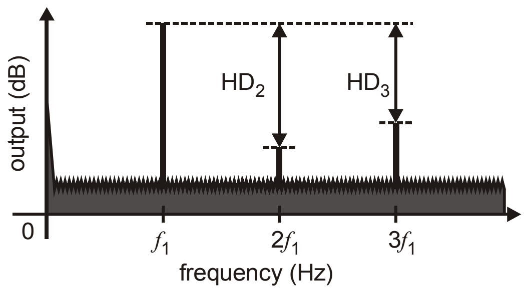


Figure 2-9: *Definition of harmonic distortion ($HD_x$).*

***Intermodulation distortion* ($IM_x$)** – defined for a maximum power two tone input test, where distortion components due to non-linearity are present in spectral positions which are combinations of the input signal frequencies $f_1$ and $f_2$. The intermodulation distortion is defined as the ratio between the power of one of the signal tones and the power of $x^{th}$-order intermodulation distortion tone. The second ($IM_2$) and the third ($IM_3$) intermodulation components, respectively located at the spectral positions ($f_2$-$f_1$),($f_1$+$f_2$) and (2$f_1$-$f_2$),(2$f_2$-$f_1$), are the most important (Figure 2-10).

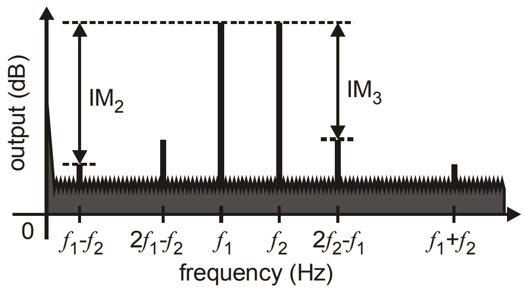


Figure 2-10: *Definition of intermodulation distortion ($IM_x$).*

***Spurious-free dynamic range*** **(SFDR)** – ratio between maximum power input sinusoidal and the strongest in-band spurious tone power (Figure 2-11).

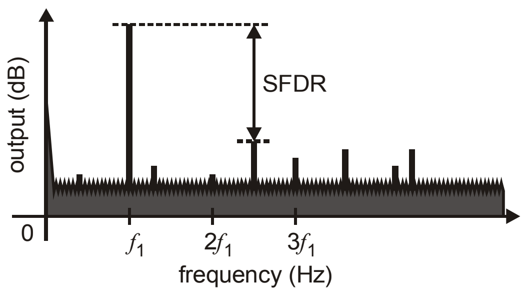


Figure 2-11: *Definition of spurious free DR (SFDR).*

***Signal-to-noise-and-distortion ratio*** **(SNDR)** – ratio between the input power (normally a sinusoidal signal) and total noise and distortion power inside a certain bandwidth.

***Peak*** **SNDR** – ratio between the power of the sinusoidal input for peak SNR and the total noise and distortion power inside the ADC bandwidth.

***Intermodulation intercept point* (IP$_x$)** – theoretical sinusoidal input carrier power for which the $x^{th}$-order intermodulation product is equal to the signal carrier power. $IP_2$ and $IP_3$ (Figure 2-12) are the most important.

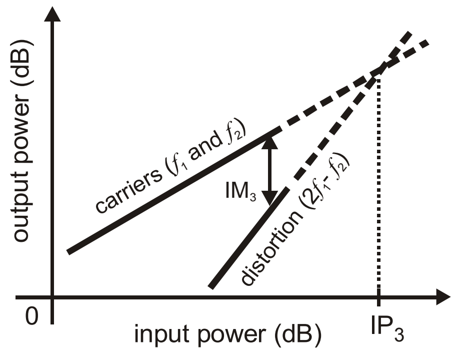


Figure 2-12: *Definition of $IP_3$.*

***Cross-modulation distortion* (CM)** – modulation of the spectrum around the carrier ($f_1$) of the wanted channel by the spectral content of an unwanted channel due to non-linearity (Figure 2-13). Cross-modulation distortion is defined as the distance between the desired carrier and the strongest cross-modulation distortion component.

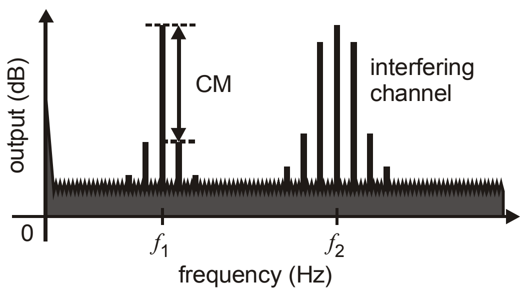


Figure 2-13: *Definition of cross-modulation distortion (CM).*

## 2.3 Desensitization and Blocking

The input-referred DR of an ADC can be extended if it is preceded by a variable gain amplifier and a filter (Figure 2-14).

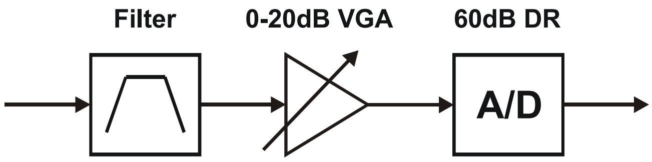


Figure 2-14: *ADC with extended dynamic range.*

The input-referred DR of a 60dB DR ADC combined with a VGA with 3 programmable gains steps (0, 10, 20 dB) is shown in Figure 2-15. In this picture the SNR for a sinusoidal input is plotted for several amplitudes, varying from 0 to -80$dB_{FS}$. For larger inputs, the VGA gain is set to 0dB and the peak SNR of 60dB is achieved. For very small signals, the ADC input is amplified by 20dB, extending the input-referred DR from 60 to 80 dB. The combination VGA+ADC works properly if the VGA is perfectly linear and no strong interferers are present.

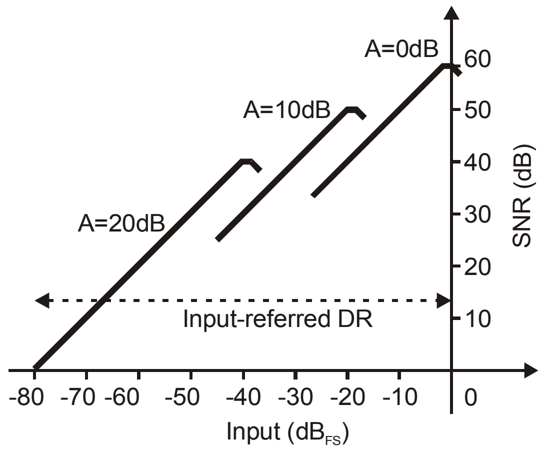


Figure 2-15: *Input DR of the combination VGA + ADC.*

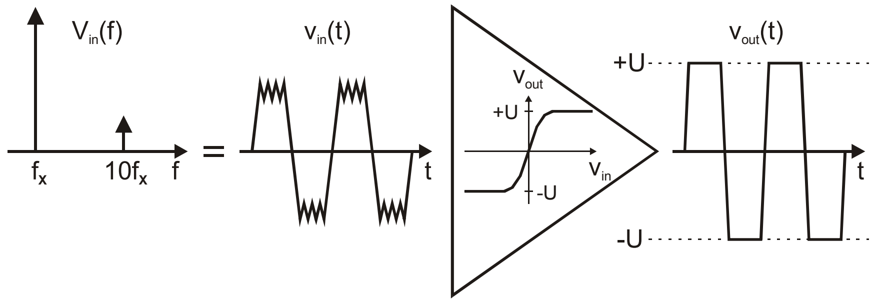


Figure 2-16: *Saturated output of an amplifier with a compressive characteristic (worst case).*

However, the situation described in the last sentence of the previous paragraph is unrealistic: strong interferers and adjacent channels are often present in the received spectrum, and real amplifiers present non-linearities like saturation. Figure 2-16 shows a typical amplifier with a (non-linear) compressive input-output characteristic with a two-tone input and a clipped output. If the desired information is carried by the weak tone at a high frequency and the strong tone is part of an interference channel, the large amplitude signal forces the amplifier to saturate. In the worst case, the high-frequency tone is not present at the output (gain 0). In this situation the receiver is said to be desensitized and the desired signal is blocked by the strong interference [1].

Therefore, in a practical receiver, the combination VGA+ADC is always preceded by filtering as in Figure 2-14. The effect of a filter preceding an AGC controlled VGA is better understood with the aid of Figure 2-17. The wanted channel centred at the IF is surrounded by unwanted adjacent channels. When the wanted channel is very strong the AGC loop sets the gain of the VGA to 0dB (Figure 2-17a). The unwanted channels are strongly attenuated by the filter and the desired channel is not blocked. Figure 2-17b shows the other extreme case: the wanted channel is so weak that the VGA is set to its maximum gain. The wanted signal can still be properly amplified because strong adjacent channels are filtered out before the VGA. In the case the filter is absent, the whole receiver could be desensitized by strong adjacent channels like the amplifier in Figure 2-16.

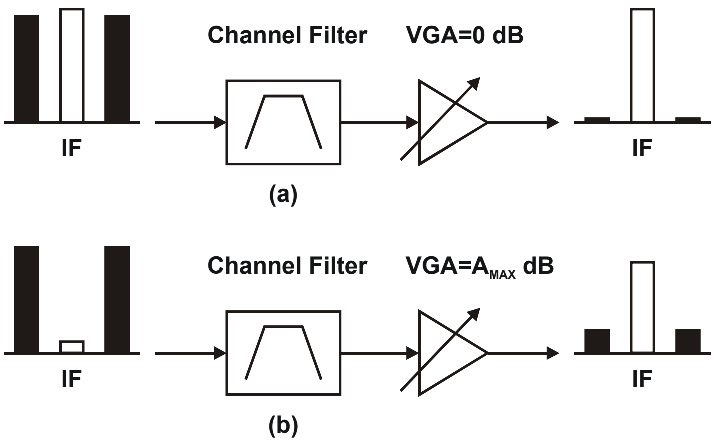


Figure 2-17: *Combination of a filter and a VGA: strong desired channel and VGA=0dB (a), weak desired channel and VGA=$A_{MAX}$ dB (b).*

In Figure 2-17 it is assumed that no interferers are present in the transition band of the filter. Figure 2-18 shows the common situation when some adjacent channels are present in the frequency bands where the filter attenuation is not strong. This is the case of a narrow-band filter with limited quality-factor centred around a high IF, for example an AM channel selection filter with 30 kHz bandwidth for the 10.7 MHz IF. The final channel selection is often performed in the digital domain to compensate for the lack of attenuation around the desired channel. When the wanted channel and the adjacent channels are all weak (Figure 2-18a), the maximum VGA gain is selected and the wanted channel is properly received. However, when the adjacent channels are very strong (Figure 2-18b, neglecting harmonic distortion), the insufficiently attenuated adjacent signals can saturate the VGA and significantly reduce the amplification of the wanted channel. In an extreme case, the wanted channel can be blocked. The sensitivity of the receiver can be improved by a higher quality-factor channel filter. However, this option increases the overall cost of the radio. The most elegant solution to increase the receiver sensitivity is to use an ADC with higher DR, in such a way that the VGA and AGC loop are not required anymore. Of course, this solution poses some challenging DR specifications on the ADC design (see Chapter 5).

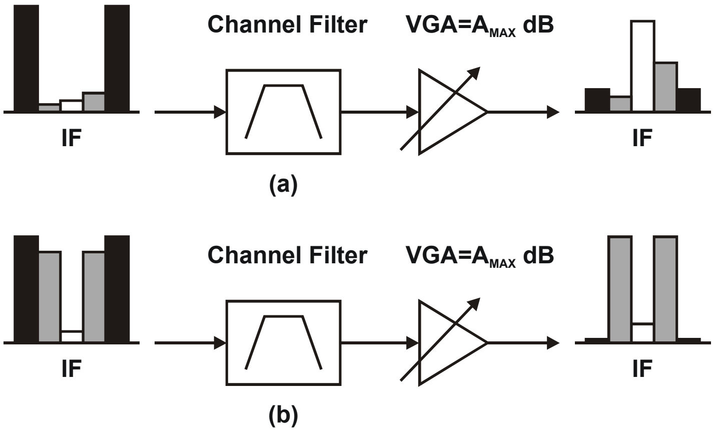


Figure 2-18: *Filter and VGA with weak desired channel surrounded by non-attenuated weak adjacent channels (a), by strong adjacent channels (b).*

## 2.4 Image Rejection

All architectures discussed in section 2.1 rely on one or more down-conversion stages to successfully select the wanted channel. The single-path (real) mixing operation (shown in Figure 2-19a) is described mathematically by a multiplication of the mixer RF input spectrum with a single-tone sinusoidal at the radian frequency $\omega_{LO}$ [1]:

$$RF \times LO = IF = \cos(\omega_{RF}t) \cdot \cos(\omega_{LO}t) = \frac{\cos[(\omega_{RF}-\omega_{LO})t]}{2} + \frac{\cos[(\omega_{RF}+\omega_{LO})t]}{2} \quad (2\text{-}2)$$

Equation (2-2) reveals that the input RF spectrum is down-converted to an IF and up-converted to a higher RF. In a receiver, the up-converted components are not a problem if the mixing operation is followed by lowpass filtering. More important is the fact that the RF frequency components located at ($\omega_{LO}+\omega_{IF}$) and ($\omega_{LO}-\omega_{IF}$) are both converted to the same IF. If the first is the center frequency of the signal band, the second is named the image band, and vice-versa. Figure 2-20 shows the image problem in single-path low-IF (Figure 2-2) and zero-IF (Figure 2-3) receivers.

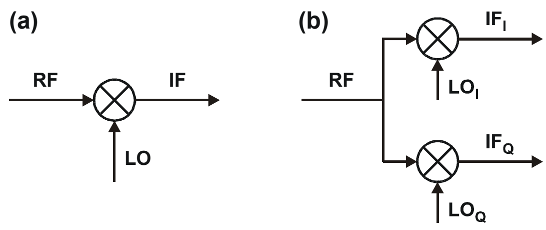


Figure 2-19: *Single-path (real) mixing (a), quadrature-paths (complex) mixing (b).*

Figure 2-20a describes the image problem in the low-IF down-conversion: if the power at the image band is stronger than the signal channel, the wanted information is completely corrupted. This is the reason why single-path mixers have to be preceded by a high selective and often tuneable image rejection filter in heterodyne receivers (Figure 2-2). In the case of single-path zero-IF down-mixing (Figure 2-20b), the image band contains a mirrored version of the wanted band. Both are converted to dc and the desired information is always corrupted if a single sideband (SSB) modulation scheme, as in citizens band (CB) amateur radio, is not used [1].

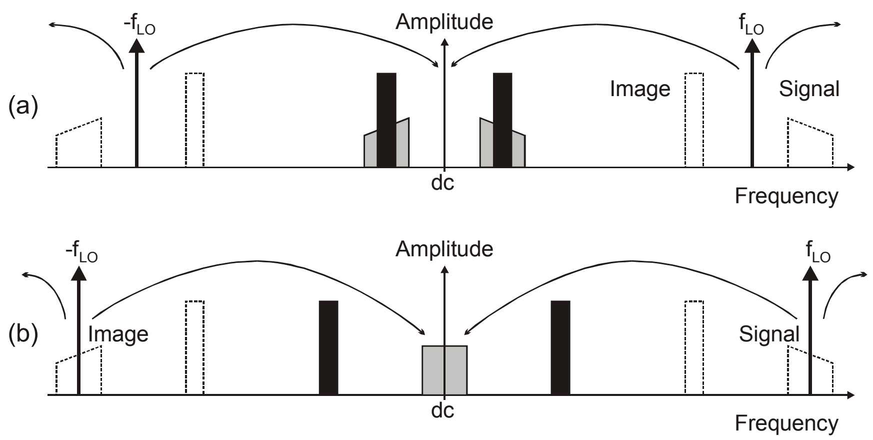


Figure 2-20: *Single-path (real) low-IF down-conversion (a) and single-path zero-IF down-conversion (b).*

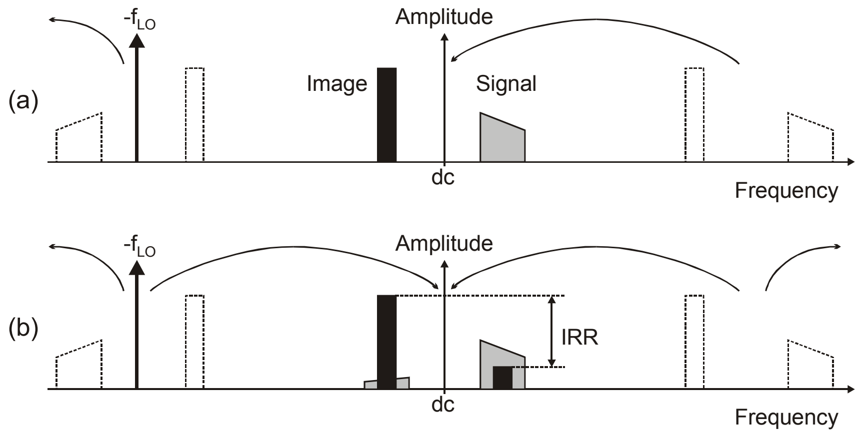


Figure 2-21: *Quadrature-paths low-IF down-conversion with ideal matching (a), with I/Q mismatch (b).*

The quadrature (complex) mixing operation depicted in Figure 2-19b is able to alleviate the image problem. The mixers' oscillator in-phase ($LO_I$) and quadrature-phase ($LO_Q$) inputs are $90^o$ phased from each other:

$$
\begin{aligned}
LO_I &= \cos(\omega_{LO}t) = \frac{e^{j\omega_{LO}t}}{2} + \frac{e^{-j\omega_{LO}t}}{2} \\
LO_Q &= -\sin(\omega_{LO}t) = \frac{e^{j\omega_{LO}t}}{-2j} + \frac{e^{-j\omega_{LO}t}}{2j}
\end{aligned}
\tag{2-3}
$$

The signals $LO_I$ and $LO_Q$ can be considered the real and imaginary parts of the complex signal $LO = LO_I + jLO_Q$. The multiplication of the real RF input with LO results in another complex signal at the IF output:

$$
LO = \cos(\omega_{LO}t) - j\sin(\omega_{LO}t) = e^{-j\omega_{LO}t} \tag{2-4}
$$

$$
RF \times LO = IF = \cos(\omega_{RF}t) \cdot e^{-j\omega_{LO}t} = \frac{e^{j(\omega_{RF}-\omega_{LO})t}}{2} + \frac{e^{j(-\omega_{RF}-\omega_{LO})t}}{2} \tag{2-5}
$$

Equation (2-5) reveals how the quadrature mixer distinguishes between positive and negative frequencies. Frequencies above $\omega_{LO}$ (signal band) are down-converted to the right side of the spectrum, while frequencies below $\omega_{LO}$ (image band) are down-converted to the left side. Figure 2-21a depicts the ideal quadrature low-IF down-conversion. If the phase-shift between the local oscillator outputs is exactly $90^o$ and their amplitudes are the same, no fraction of the image band is down-converted to the positive frequencies [1]. In the case of a phase-shift different of $90^o$ and/or amplitude mismatch between $LO_I$ and $LO_Q$, some power will leak from the image band into the signal band, and vice-versa (Figure 2-21b).

The image rejection ratio (*IRR*) quantifies the relation between the negative-half-plane image band and positive-half-plane leaked image power (Figure 2-21b) in quadrature receivers. The *IRR* can be expressed as a function of the finite mismatch between $LO_I$ and $LO_Q$ oscillator inputs, present in any analog implementation. This mismatch can be modelled as a gain error $a_e$ and phase error $\phi_e$. The effect of gain error in the complex mixing can then be calculated:

$$RF \times LO = IF = \cos(\omega_{RF}t) \cdot [(1 + a_e) \cdot \cos(\omega_{LO}t) - j\sin(\omega_{LO}t)] \quad (2\text{-}6)$$

$$IF \approx \frac{1}{2}(e^{j(\omega_{RF}-\omega_{LO})t} + e^{j(-\omega_{RF}-\omega_{LO})t}) + \frac{a_e}{2}(e^{j(\omega_{RF}+\omega_{LO})t} + e^{j(\omega_{LO}-\omega_{RF})t}) \quad (2\text{-}7)$$

Equation (2-7) quantifies the fraction of the down-converted spectrum power that leaks from the positive-half-plane into the negative-half-plane, and vice-versa, due to the gain error $a_e$. The effect of the phase error $\phi_e$ in the complex mixing can be calculated as well:

$$RF \times LO = IF = \cos(\omega_{RF}t) \cdot [\cos(\omega_{LO}t) - j\sin(\omega_{LO}t + \phi_e)] \quad (2\text{-}8)$$

$$IF \approx \frac{1}{2}(e^{j(\omega_{RF}-\omega_{LO})t} + e^{j(-\omega_{RF}-\omega_{LO})t}) - \frac{j\phi_e}{2}(e^{j(\omega_{RF}+\omega_{LO})t} + e^{j(\omega_{LO}-\omega_{RF})t}) \quad (2\text{-}9)$$

Equation (2-9) quantifies the fraction of the down-converted spectrum power that leaks from the positive-half-plane into the negative-half-plane, and vice-versa, due to the phase error $\phi_e$. Equations (2-6) and (2-9) can be combined to give an approximate expression for the *IRR* [1]:

$$IRR_{dB} \approx 10 \cdot \log\left[\frac{4}{a_e^2 + \phi_e^2}\right] \tag{2-10}$$

Figure 2-22a depicts the zero-IF down-conversion in a quadrature receiver. The complex mixer output contains the wanted signal band shifted directly to dc, while the image band is up-converted to a higher negative band around $-2\omega_{LO}$. Because of their inherent simplicity, homodyne receivers are a very attractive solution for fully integrated receivers.

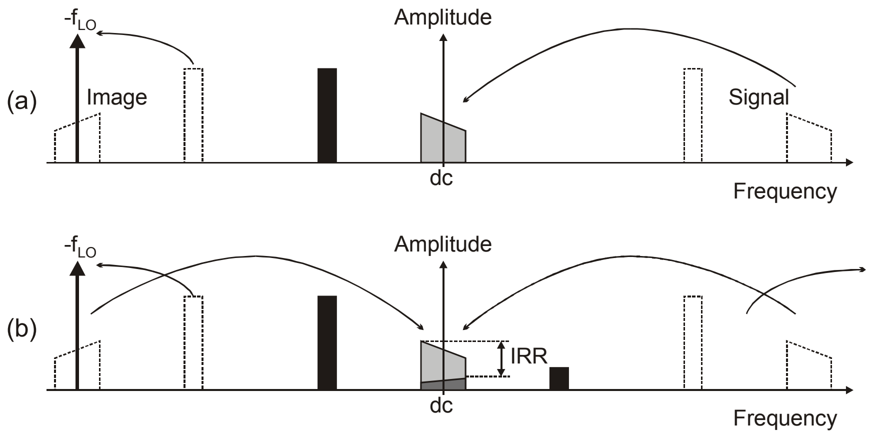


Figure 2-22: *Quadrature-paths zero-IF down-conversion with ideal matching (a), with I/Q mismatch (b).*

Figure 2-22b shows the effect of I/Q mismatch in the quadrature homodyne receiver. Because the image band is a mirrored version of the desired band for most modulation schemes, quadrature zero-IF receivers have much more relaxed image rejection requirements. The major drawback of the zero-IF architecture is that dc offset and 1/*f* noise can corrupt the baseband information. Furthermore, the fact that the mixer oscillator operates at same frequency as the IF may cause undesirable interferences [1].

## 2.5 Analog Radio Broadcasting

AM and FM are the most popular analog broadcasting radio services, being introduced, respectively, in 1920 and 1941 [25]. By far, the most widespread AM broadcasting band is the medium wave (MW), from 520 to 1,710kHz world wide. The channel width and the carrier spacing are 10kHz in the Americas, and 9kHz in the rest of the world. The long wave (LW) band, from 153 to 279kHz, has also been used in Europe, Africa and the Middle-East for AM broadcasting. The short wave (SW), from 2.3 to 26.1MHhz, has been used worldwide for very long distances broadcasting based on ionospheric reflection.

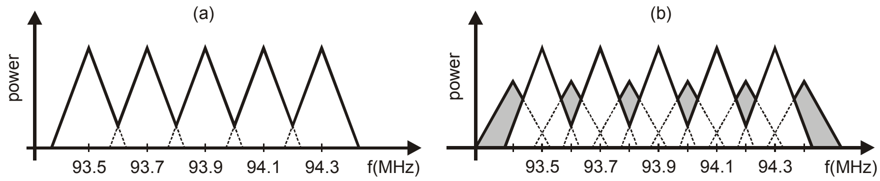


Figure 2-23: *FM channel distribution in the Americas (a) and in Europe (b).*

The FM radio broadcasting band ranges from 87.5 to 108.5MHz in most countries. An additional band, ranging from 65.9 to 74MHz, was also assigned in the former communist block, while Japan has its own FM radio allocation to the 76 to 90MHz band. In the Americas, the FM band carrier spacing and the modulated channel width are 200kHz (Figure 2-23a). In Europe, the carrier spacing is 100kHz and the channel width is 150kHz (Figure 2-23b). Because the FM band is overcrowded in Europe, adjacent FM channels are not used in the same region.

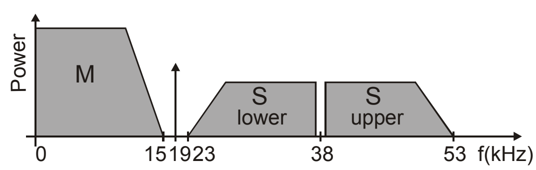


Figure 2-24: *FM stereo baseband channel.*

In order to accommodate stereo audio – left (L) and right (R) outputs – and maintain compatibility with mono receivers, a 55 kHz FM baseband stereo channel consists of a mono compatible M=(L+R)/2 band and a double-sideband suppressed carrier S=(L-R)/2 band. The M band ranges from 30 Hz to 15 kHz, and a pilot tone is present at 19 kHz. The S band, centred around 38 kHz, ranges from 23 to 53 kHz (Figure 2-24).

## 2.6 Digital Radio Broadcasting

In spite of the widespread use of AM/FM radios, the quality of the reception is limited by the characteristics of the old analog broadcasting standards. Several digital telecommunication techniques, developed since the 1980s, could have been employed for audio broadcasting. However, due to the lack of available spectrum, most of the proposals for terrestrial digital audio broadcasting require the shut down of the traditional analog transmission. As a result of the lack of backwards compatibility with AM/FM radio, in most of the world no full digital broadcasting solution has been adopted up to now for terrestrial broadcasting.

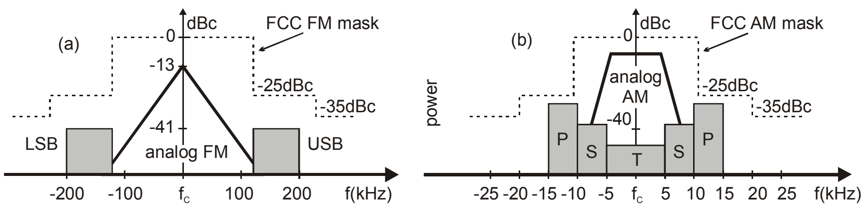


Figure 2-25: *Hybrid IBOC spectra. FM (a), AM (b).*

The *In-Band, On-Channel* (IBOC) standard, developed during the late 1990s, proposes a hybrid digital broadcasting solution within the traditional AM and FM bands [26]. The digital information is transmitted according to the FCC transmission masks for AM and FM broadcasting (Figure 2-25). Traditional receivers are able to receive the analog signal without noticeable quality loss on the reception because the digital broadcasting is perceived as an additional source of in-band noise. Due to the blend-to-analog feature, if the quality of the digital broadcasting is inferior to the analog broadcasting, an IBOC compatible receiver switches back to the analog reception mode.

A hybrid FM-IBOC channel is shown in Figure 2-25a. The upper and lower digital sidebands are present together with the analog information. The digital audio is transmitted using *Orthogonal Frequency Division Multiplexing* (OFDM) and *Forward Error Correction* (FEC) coding. Both digital sidebands contain the same baseband information to increase robustness against interferers from the 1st adjacent analog carrier, located 200kHz away from the tuned channel carrier. When both digital sidebands are properly received, because of the *Complementary Pair Convolution* (CPC) coding, additional audio quality improvement is achieved [27]. However, outside the Americas region, where the carrier distance is just 100kHz, FM-IBOC transmission is much less feasible.

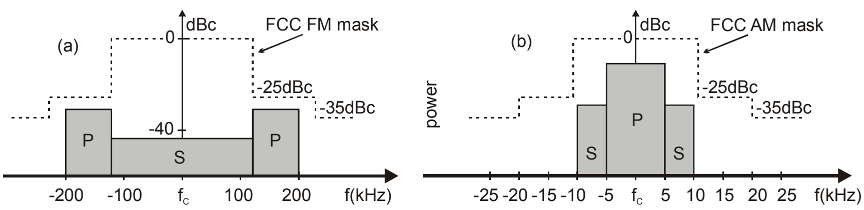


Figure 2-26: *All-digital FM (a) and AM (b) IBOC spectra.*

A hybrid AM-IBOC channel is shown in Figure 2-25b. Again OFDM and FEC coding are used for the primary (P) and secondary (S) digital audio sidebands. Because the analog information is amplitude modulated, some terciary (T) digital information is transmitted as a quadrature-phase component of the analog broadcasting. However, due to non-satisfactory performance caused by strong adjacent channel interferers, hybrid AM-IBOC is yet to be adopted [28].

Both FM and AM IBOC standards are prepared for transition to full digital broadcasting, if and when the analog broadcasting is shut-down. Figure 2-26 shows the all-digital FM and AM IBOC channel spectra. An alternative solution to increase the FM band digital broadcasting data rate, is to discontinue the stereo FM analog broadcasting. In this scenario, the stereo S channel bandwidth would be allocated to another pair of IBOC OFDM channels. The FM analog mono (M) channel service would continue available without interruption for backwards compatibility.

## 2.7 Integrated Solutions for AM/FM Receivers

Following the development of integrated circuit technology, several LSI (large scale of integration) based AM/FM receiver solutions became available by the mid-1970s [4]. This section presents an overview of commercial chip-sets available for car radios until the present time. Figure 2-27 shows a simplified diagram of a 4 ICs bipolar technology receiver chip-set. All active components required for AM demodulation were already integrated together. FM demodulation was partitioned between RF and IF ICs, while stereo decoding required its own dedicated analog processor. A 5th IC was employed in car radios to implement pulsed interference (e.g. from the ignition spark) cancellation. Several non-integrated components like filters, passives and crystal oscillators were also needed.

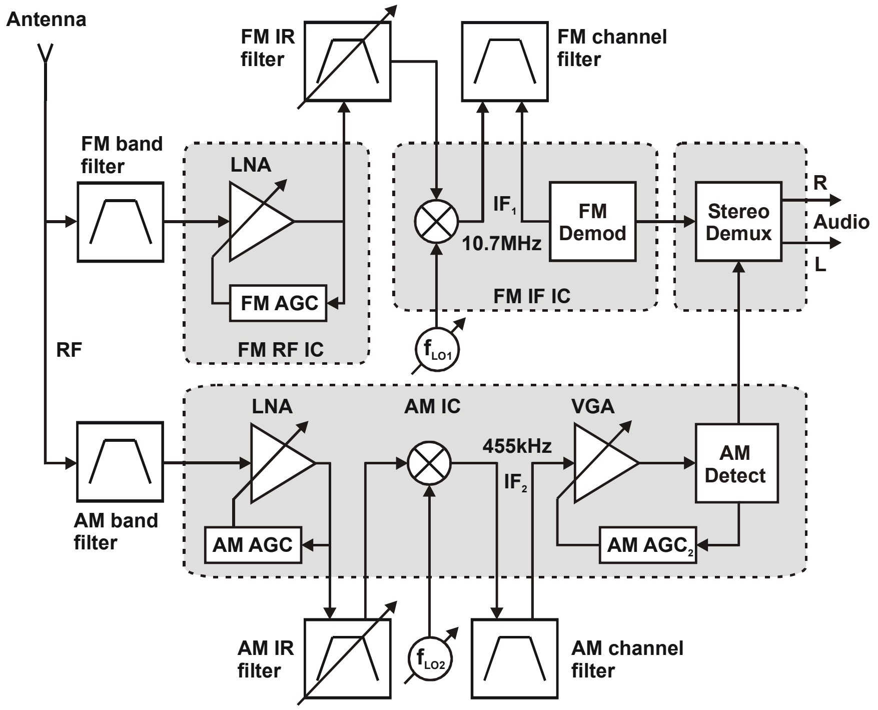


Figure 2-27: *Analog AM/FM receiver LSI chip-set [4].*